

REPORT ITU-R SM.2028

**Monte Carlo simulation methodology for the use in sharing
and compatibility studies between different
radio services or systems**

(Question ITU-R 211/1)

(2001)

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Summary

In this Report background information on a Monte Carlo radio simulation methodology is given. Apart from giving general information this text also constitutes a specification for the first generation of spectrum engineering advanced Monte Carlo analysis tool (SEAMCAT) software which implements the Monte Carlo methodology applied to radio communication scenarios.

General

The problem of unwanted emissions, as a serious factor affecting the efficiency of radio spectrum use, is being treated in depth in various fora, internal and external to the European Conference of Postal and Telecommunications Administrations (CEPT). As the need to reassess the limits for unwanted emissions within Appendix 3 of the Radio Regulations (RR) is observed, it is widely recognized that a generic method is preferable for this purpose.

One of numerous reasons why generic methods are favoured is their a priori potential to treat new communication systems and technologies as they emerge. Another reason is that only a generic method can aspire to become a basis for a widely recognized analysis tool.

The Monte Carlo radio simulation tool described in this Report was developed, based on the above considerations, within the European Radiocommunication Committee (ERC) process.

SEAMCAT

SEAMCAT is the implementation of a Monte Carlo radio simulation model developed by the group of CEPT administrations, European Telecommunications Standards Institute (ETSI) members and international scientific bodies. SEAMCAT is a public object code software distributed by the CEPT European Radiocommunications Office (ERO), Copenhagen. The Web address is as follows:

<http://www.ero.dk> Folder: Project Activities

The software is also available in the ITU-R software library. Further details can be provided by the SEAMCAT Management Committee (SMC), which can be contacted via the ERO e-mail: ero@ero.dk. The source code is only available for members of SMC.

1 Background

In order to reassess the limits for unwanted emissions within RR Appendix 3, it is desirable to develop an analytical tool to enable us to evaluate the level of interference which would be experienced by representative receivers. It has been agreed in the ITU-R that level of interference should be expressed in terms of the probability that reception capability of the receiver under consideration is impaired by the presence of an interferer. To arrive at this probability of interference, statistical modelling of interference scenarios will be required and this Report describes the methodology and offers a proposal for the tool architecture.

The statistical methodology described here and used for the tool development is best known as Monte Carlo technique. The term “Monte Carlo” was adopted by von Neumann and Ulan during World War II, as a code-name for the secret work on solving statistical problems related to atomic bomb design. Since that time, the Monte Carlo method has been used for the simulation of random processes and is based upon the principle of taking samples of random variables from their defined probability density functions. The method may be described as the most powerful and commonly used technique for analysing complex statistical problems. The Monte Carlo approach does not have an alternative in the development of a methodology for analysing unwanted emission interference.

The approach is:

- generic: a diversity of possible interference scenarios can be handled by a single model.
- flexible: the approach is very flexible, and may be easily devised in a such way as to handle the composite interference scenarios.

2 Monte Carlo simulation methodology: An overview

This methodology is appropriate for addressing the following items in spectrum engineering:

- sharing and compatibility studies between different radio systems operating in the same or adjacent frequency bands, respectively;
- evaluation of transmitter and receiver masks;
- evaluation of limits for parameters such as unwanted (spurious and out-of-band) blocking or intermodulation levels.

The Monte Carlo method can address virtually all radio-interference scenarios. This flexibility is achieved by the way the parameters of the system are defined. The input form of each variable parameter (antenna pattern, radiated power, propagation path,...) is its statistical distribution function. It is therefore possible to model even very complex situations by relatively simple elementary functions. A number of diverse systems can be treated, such as:

- broadcasting (terrestrial and satellite);
- mobile (terrestrial and satellite);
- point-to-point;
- point-to-multipoint, etc.

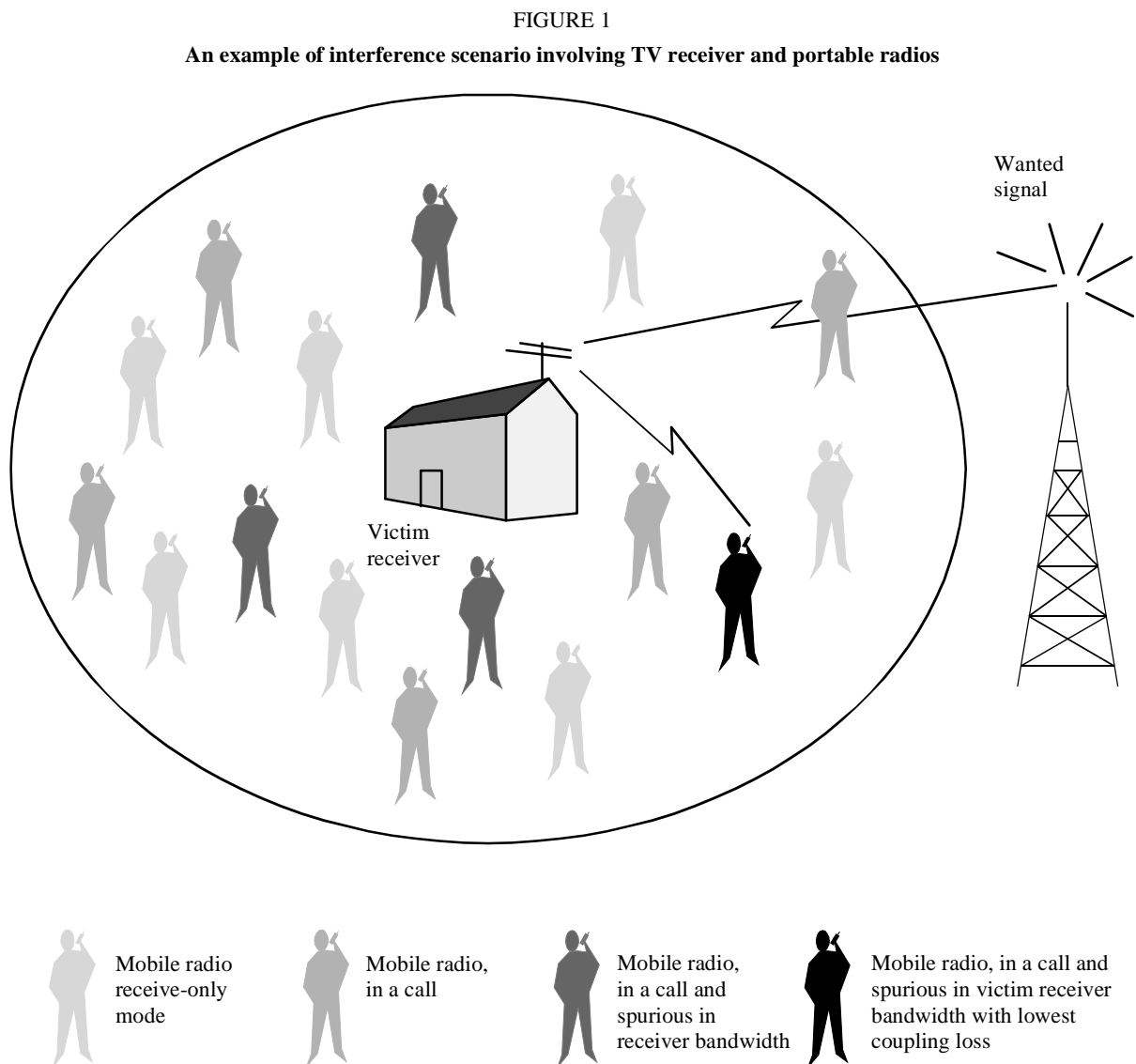
The principle is best explained with the following example, which considers only unwanted emissions as the interfering mechanism. In general the Monte Carlo method addresses also other effects present in the radio environment such as out-of-band emissions, receiver blocking and intermodulation.

Some examples of actual applications of this methodology are:

- sharing study between short range devices (Bluetooth) and radio local area networks (RLANs) in the industrial, scientific and medical (ISM) band at 2.4 GHz;
- compatibility study for international mobile telecommunications-2000 (IMT-2000) and PCS1900 around 1.9 GHz.

2.1 Illustrative example (only unwanted emissions, most influential interferer)

For interference to occur, it has been assumed that the minimum carrier-to-interference ratio, C/I , is not satisfied at the receiver input. In order to calculate the C/I experienced by the receiver, it is necessary to establish statistics of both the wanted signal and unwanted signal levels. Unwanted emissions considered in this simulation are assumed to result from active transmitters. Moreover, only spurious falling into the receiving bandwidth have been considered to contribute towards interference. For the mobile to fixed interference scenario, an example is shown in Fig. 1.

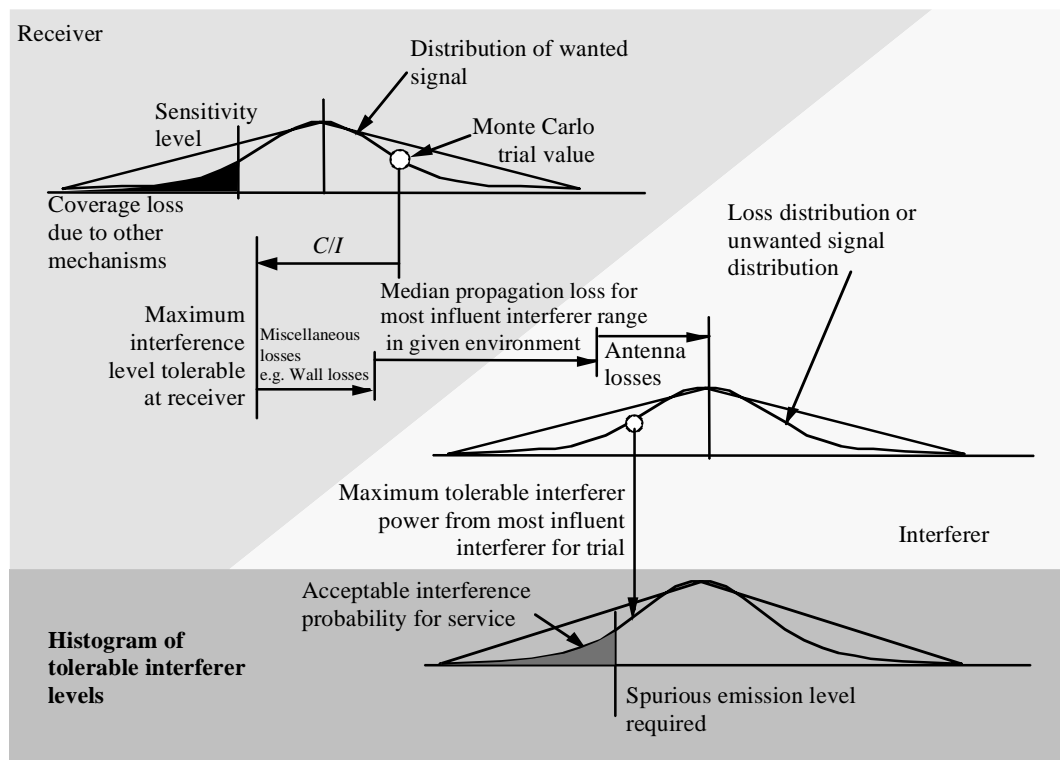


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Many potential mobile transmitters are illustrated. Only some of the transmitters are actively transmitting and still fewer emit unwanted energy in the victim receiver bandwidth. It is assumed that interference occurs as a result of unwanted emissions from the most influential transmitter with the lowest path loss (median propagation loss + additional attenuation variation + variation in transmit power) to the receiver.

An example of Monte Carlo simulation process as applied to calculating the probability of interference due to unwanted emission is given in Fig. 2. For each trial, a random draw of the wanted signal level is made from an appropriate distribution. For a given wanted signal level, the maximum tolerable unwanted level at the receiver input is derived from the receiver's C/I figure.

FIGURE 2
An example formulation of the Monte Carlo evaluation process



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For the many interferers surrounding the victim, the isolation due to position, propagation loss (including any variations and additional losses) and antenna discrimination is computed. The lowest isolation determines the maximum unwanted level which may be radiated by any of the transmitters during this trial.

From many trials, it is then possible to derive a histogram of the unwanted levels and for a given probability of interference, then to determine the corresponding unwanted level.

By varying the values of the different input parameters to the model and given an appropriate density of interferers, it is possible to analyse a large spectra of interference scenarios.

3 Architecture requirements

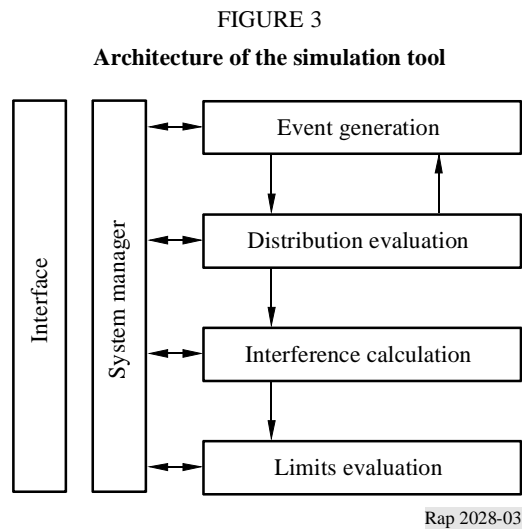
One of the main requirements is to select such an architectural structure for the simulation tool which would be flexible enough to accommodate analysis of composite interference scenarios in which a mixture of radio equipment sharing the same habitat and/or multiple sources of interference (e.g. out-of-band emission, spurious emission, intermodulation, ...) are involved and can be treated concurrently.

Other requirements would be that the proposed architecture consists of modular elements and is versatile enough to allow treatment of the composite interference scenarios.

The proposed Monte Carlo architecture which meets these constraints is presented in Fig. 3. The proposed architecture is basically of a sequential type and consists of four processing engines:

- event generation engine;
- distribution evaluation engine;
- interference calculation engine;
- limits evaluation engine.

The schematic view of the entire tool is in Fig. 3.



The list of interference parameters and their relevance to one or more of the processing engines is shown in Annex 1.

3.1 Event generation engine

The event generation engine (EGE) takes the relevant parameters from the submitted interference scenario and generates information on the received signal strength (RSS) of the desired, as well as on the strength for each of the interfering, signals included in the composite interference scenario. This process is repeated N times, where N is a number of trials which should be large enough to produce statistically significant results. Generated samples of the desired, as well as all interfering, signals are stored in separate data arrays of the length N .

The trials on parameters being common for desired and interfering radio paths are done concurrently in order to capture possible correlation between desired and interfering signals. Such an implementation will not cover those seldom cases of interference in which one interference mechanism is excited by another interference (e.g. a strong emission of the first transmitter mixes with a spurious emission of the second transmitter and produces an intermodulation type of interference).

The flow chart description and detailed algorithm description for the EGE are presented in Annex 2.

List of potential sources of interference to be found in a radio environment includes:

Transmitter interference phenomena:

- unwanted (spurious and out-of-band) emissions;
- wideband noise;
- intermodulation;
- adjacent channel;
- co-channel.

Receiver interference phenomena:

- spurious emission.

Background noise:

- antenna noise;
- man-made noise.

Other receiver interference susceptibility parameters:

- blocking;
- intermodulation rejection;
- adjacent and co-channel rejections;
- spurious response rejection.

All of the above sources can be classified into three generic interference mechanism categories: undesired emission, intermodulation and receiver susceptibility. Each of the above three categories requires a different model for physical processes being characteristic for that interfering mechanism. The man-made noise and the antenna temperature noise can be considered as an increase of the thermal noise level, decreasing thus the sensitivity of a receiver, and can be entered in the simulation when the criteria of interference is I/N (interference-to-noise ratio) or $C/(I + N)$.

3.2 Distribution evaluation engine

The distribution evaluation engine (DEE) takes arrays of the data generated by the EGE and processes the data with the aim of:

- a) assessing whether or not the number of samples is sufficient to produce statistically stable results;
- b) calculating correlation between the desired signal and interfering signal data and between different types of the interfering signals (e.g. blocking vs. unwanted emissions);
- c) calculating a known continuous distribution function, e.g. Gaussian, as the best fit to the generated distributions of the desired and interfering signal data.

Items a) and c) can be achieved using well known goodness-of-fit algorithms for general distributions such as the Kolmogorov-Smirnov test. Applicability of the fit to this specific task is to be further investigated in the planned phase 2 of the development of the methodology.

If DEE detects unacceptable variation in discrete distribution parameters estimated in two successive estimations using N and $N + \Delta N$ sample sizes, the EGE is instructed to generate another ΔN of additional samples. This test is repeated until a tolerable variation of the parameters is measured over the pre-defined number of successive tests.

Three different kinds of outputs are possible from the DEE engine:

- data arrays of the wanted and interfering signals. This is the output in the case that a high degree of correlation is detected between the wanted and any of the interfering signals;
- discrete distributions of the wanted and interfering signals are passed in the case of a weak correlation between the signals or in the case that there was no correlation between the signals but no continuous distribution approximation with satisfactory accuracy was possible;
- continuous distribution functions of the wanted and interfering signals are passed to the interference calculation engine (ICE) in the case that signals were de-correlated and discrete distributions were successfully approximated with continuous distribution functions.

The proposed flow chart and detailed algorithm specification are presented in Annex 3.

3.3 Interference calculation engine (ICE)

The ICE is the heart of the proposed architecture. Here, information gathered by the EGE and processed by DEE are used to calculate probability of interference. Depending on which kind of information was passed from DEE to ICE, three possible modes of calculating the probability of interference are identified, as shown in Annex 4.

Mode 1: Data arrays for $dRSS$ (wanted signal) and i_nRSS (interfering signal resulting from n different systems) passed by the DEE to the ICE, and vector representing the composite interfering signal I is calculated as a sum of the i_nRSS data vectors.

Mode 2: Distribution function for the composite interfering signal is calculated by taking random samples for i_nRSS distributions and linearly adding them up.

Mode 3: The i_nRSS is calculated using numerical or analytical integration of the supplied distribution functions for each of the interference sources.

Mode 4: All signals are assumed to be mutually independent and the overall probability for interference is identified as the probability to be disturbed by at least one kind of interference.

Different criteria for calculation of interference probability can be accommodated within the processing engine. A cumulative probability functions (cpf) can be calculated for C/I , $C/(N + I)$ or $N/(N + I)$ random variables.

The flow of information together with associated processes is shown in the form of a flow chart in Annex 4.

All interfering signal distributions are calculated with respect to reference levels, or functions, of unwanted (emission mask), blocking (receiver mask) or intermodulation attenuation. Interfering signal distributions for some other reference levels or functions can be derived by first order (unwanted or blocking) or third order (intermodulation) linear translation of the reference distributions (see Annex 4).

3.4 Limits evaluation engine (LEE)

The LEE is to play a very important role in two aspects of the tool development:

- selection of optimal values for the limits;
- verification of the tool.

Output from the ICE is presented as a multi-dimensional surface characterising the dependence of the probability of interference versus the radio parameters. Two main features of the probability surface are:

- the same probability of interference is achieved by different sets of the limit values for the radio parameters under consideration;
- probability of interference parameter is not used in the radio system design and as such does not lend itself nicely for the validation through the system performance measurements. Instead, degradation in system coverage or traffic capacity seems to be more appropriate for understanding impact of a particular probability of interference to the radio system performance.

The radio variables are transformed from the probabilistic space into a system performance space enabling us to evaluate the system performance degradation due to presence of interference. When the inter-system compatibility is analysed (e.g. unwanted emission), radio coverage and/or traffic capacity can be used to evaluate the impact of the radio parameters limits. For the case of intra-system compatibility study (e.g. out-of-band emission), spectrum efficiency should be used to derive appropriate values for the radio parameters.

The limit values are derived by means of an optimization algorithm. For optimization to work, a criteria needs to be set. The criteria is usually termed the cost function and the optimization process has for task to minimize this cost function. The cost function is a function of all radio parameters and their significance to the cost can be altered by means of the weight coefficients.

The weight coefficients can integrate any of the following aspects into the optimization process:

- system availability;
- traffic capacity;
- spectrum utilization;
- technological limitations;
- economic constraints.

The set of radio parameters values for which the cost function is minimized represents the optimal solution for the limit values.

The role of LEE is very important within the tool. However, since its various elements are still under consideration, it will not be possible to include LEE into the first phase of the implementation.

ANNEX 1

List of input parameters

The following rules are applied:

- a capital letter is used for a distribution function, e.g. P ;
- a small letter is a variable (result of a calculation or a trial), e.g. p ;
- the index refers to a player: wanted transmitter, victim receiver, wanted receiver and interfering transmitter.

Parameters for the wanted transmitter (wt)

$P_{wt}^{supplied}$: sample power level taken from the above distribution (dBm)

$p_{wt}^{supplied}$: power level distribution for various transmitters (dBm)

g_{wt}^{max} : maximum antenna gain (dBi)

$pattern_{wt}$: antenna directivity within operating bandwidth (dB) (supplied as a function or a look-up table)

H_{wt} : antenna height distribution (1/m)

R_{wt}^{max} : radius of the wanted transmitter coverage (km), (not required for point-to-point)

Parameters for the victim receiver (vr)

C/I : protection ratio (dB)

g_{vr}^{max} : maximum antenna gain (dBi)

$pattern_{vr}$: antenna directivity within operating bandwidth (dB) (supplied as a function or a look-up table)

H_{vr} : antenna height distribution (1/m)

$block$: receiver frequency response (dB)

a_{vr} : receiver susceptibility characteristic is expressed as a ratio between desired interfering signal levels producing unacceptable receiver performance and is n as a function of frequency separation between the two signals

$intermod$: receiver intermodulation response (dB)

The intermodulation response is a measure of the capability of the receiver to receive a wanted modulated signal without exceeding a given degradation due to the presence of two unwanted signals with a specific frequency relationship to the wanted signal frequency

f_{vr} : frequency (MHz)

$sens_{vr}$: sensitivity of victim receiver (dBm)

b_{vr} : bandwidth of victim receiver (kHz)

Parameters for the interfering transmitter (it)

- $P_{it}^{supplied}$: power level distribution of various transmitters (dBm)
- $p_{it}^{t_hold}$: power control threshold (dBm)
- $p_{it}^{dyc_rg}$: power control dynamic range (dB)
- $p_{it}^{st_rg}$: power control step range (dB)
- g_{it}^{max} : maximum antenna gain (dBi)
- R_{it}^{max} : radius of the interfering transmitter coverage (km)
- R_{simu} : radius of the area where interferers are spread (km)
- $pattern_{it}$: antenna directivity (dB) (supplied as a function or a look-up table)
- $spur$: unwanted emissions (dBc/(reference bandwidth))
- $spur_0$: unwanted emissions floor (dBm/MHz) (unwanted emissions which would be emitted with the lowest possible power of the transmitter)
- f_{it} : frequency (MHz)
- $dens_{it}$: density (1/km²)
- p_{it}^{tx} : probability of transmission (%), which is a statistical description of the smitter activities averaged over a large number of users and long period of time
- $temp_{it}$: normalized temporal activity variation function of time of the day (1/h)

Parameters for the wanted receiver (wr) belonging to the interfering transmitter

- g_{wr}^{max} : maximum antenna gain (dBi)
- $pattern_{wr}$: antenna directivity (dB) (supplied as a function or a look-up table)
- H_{wr} : antenna height distribution (1/m)
- $sens_{wr}$: dynamic sensitivity of the wanted receiver, taking into account margin for the fast-fading and intra-system interference (dBm)

Environmental and propagation parameters

- f_{propag} : propagation law (median loss + variation) (given in Appendix 1 to Annex 2)
- f_{median} : propagation law (median loss only) (given in Appendix 1 to Annex 2)
- env : environment type (indoor/outdoor, urban/suburban/open area)

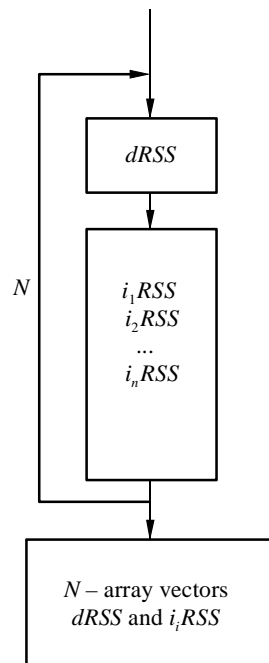
ANNEX 2

Event generation engine

Introduction

This Annex describes how to construct signals that are used in the interfering scenarios: the desired signal and the interfering signals due to unwanted emission, blocking and intermodulation. The calculated signals are stored in an array which serves as input to the DEE as shown in Fig. 4.

FIGURE 4
General flow chart of the EGE



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Inputs

The input parameters are defined in Annex 1. The different players are shown in Fig. 5.

Outputs

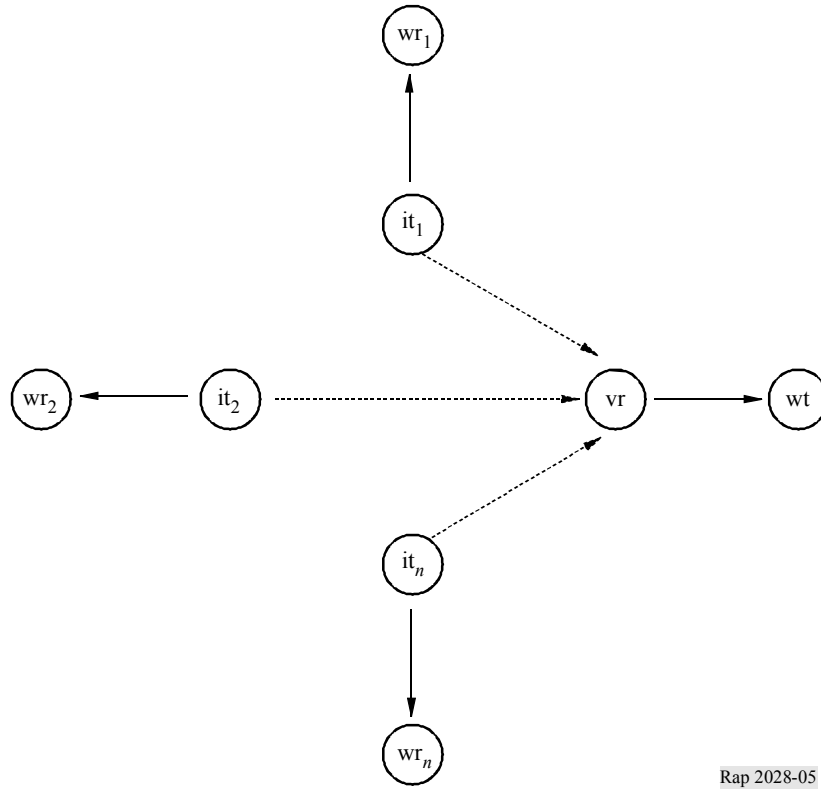
$dRSS$: desired received signal strength (dBm)

$iRSS_{spur}$: interfering received signal strength due to unwanted emissions (dBm)

$iRSS_{blocking}$: interfering received signal strength due to blocking (dBm)

$iRSS_{intermod}$: interfering received signal strength due to intermodulation (dBm)

FIGURE 5
Different players participating in the EGE



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Calculation

In this section.

- T represents a trial from a given distribution (algorithm described in Appendix 4).
- Distributions $U(0,1)$, $G(\sigma)$ and $R(\sigma)$ are defined in Appendix 3.
- Flow chart of $dRSS$ calculation is given in Appendix 5 and flow charts of $iRSS$ calculations are given in Appendices 6 and 8.

NOTE 1 – distances d between transmitters and receivers are applied with the unit in km.

a) $dRSS$ calculation

Case of variable distance:

$$dRSS = f(p_{wt}^{supplied}, g_{wt \rightarrow vr}, pl_{wt \leftrightarrow vr}, g_{vr \leftrightarrow wt}) = p_{wt}^{supplied} + g_{wt \rightarrow vr}(f_{vr}) - pl_{wt \leftrightarrow vr}(f_{vr}) + g_{vr \rightarrow wt}(f_{vr})$$

If the received signal cannot exceed a given value (i.e. if it depends on the power control implemented in the victim system) then:

$$dRSS = \min(dRSS, DRSS_{max}) \quad \text{using } dRSS \text{ as calculated before}$$

where:

$p_{wt}^{supplied}$: maximum power level distribution supplied to the wanted transmitter antenna

$$p_{wt}^{supplied} = T(P_{wt}^{supplied})$$

$pl_{wt \leftrightarrow vr}$: path loss between the wanted transmitter and the victim receiver (propagation loss, slow fading and clutter losses taken into account). Depending on whether the criteria of interference will apply to the instantaneous $dRSS$ (Rayleigh fading excluded) or to the mean $dRSS$

$$pl_{wt \leftrightarrow vr} = f_{propag}(f_{vr}, h_{vr}, h_{wt}, d_{wt \leftrightarrow vr}, env)$$

or

$$pl_{wt \leftrightarrow vr} = f_{median}(f_{vr}, h_{vr}, h_{wt}, d_{wt \leftrightarrow vr}, env)$$

where:

h_{vr} : victim receiver antenna height

$$h_{vr} = T(H_{vr})$$

e.g.: $h_{vr} = T(U(h_{vr}^{min}, h_{vr}^{max})) = h_{vr}^{min} + (h_{vr}^{max} - h_{vr}^{min}) T(U(0, 1))$

h_{wt} : wanted transmitter antenna height

$$h_{wt} = T(H_{wt})$$

e.g.: $h_{wt} = T(U(h_{wt}^{min}, h_{wt}^{max})) = h_{wt}^{min} + (h_{wt}^{max} - h_{wt}^{min}) T(U(0, 1))$

$d_{wt \leftrightarrow vr}$: distance between the victim receiver and the wanted transmitter

$$d_{wt \leftrightarrow vr} = T(R_{max}^{wt})$$

e.g.: $d_{wt \leftrightarrow vr} = R_{max}^{wt} \sqrt{T(U(0, 1))}$

Three different choices for R_{max}^{wt} are considered:

Choice 1: Given distance R_{max}^{wt}

Choice 2: Noise limited network

R_{max}^{wt} is determined by the following equation:

$$f_{median}(f_{vr}, h_{vr}, h_{wt}, d_{wt \leftrightarrow vr}, env) + f_{slowfading}(X\%) = P_{wt}^{supplied} + g_{wt}^{max} + g_{vr}^{max} - sens_{vr}$$

where:

f_{median} : propagation loss not including slow fading

$f_{slowfading}(X\%)$: fading margin to be used for 1-X% coverage loss.

In the case of lognormal fading and a 95% coverage loss at the edge of the coverage, for large distances, the value $f_{slowfading}$ is well known 1.64 times the standard deviation of the propagation loss.

Choice 3: Traffic limited network

$$R_{max}^{wt} = \sqrt{\frac{n_{channels} n_{userperchannel}}{\pi dens_{max} cluster_{frequency}}}$$

$g_{wt \rightarrow vr}$: wanted transmitter antenna gain in the victim receiver direction

$$g_{wt \rightarrow vr} = f(g_{wt}^{max}, pattern_{wt}) = g_{wt}^{max} \times pattern_{wt}(\theta_{wt \rightarrow vr}, \phi_{wt \rightarrow vr}, f_{vr})$$

where:

$(\theta_{wt \rightarrow vr}, \phi_{wt \rightarrow vr})$: azimuth and elevation angles between the top of the wanted transmitter antenna and the top of the victim receiver antenna:

e.g.: $\theta_{wt \rightarrow vr} = T(U(0, 2\pi)) = 2\pi \times T(U(0, 1))$

$$\phi_{wt \rightarrow vr} = T\left(U\left(-\frac{\pi}{2}, \frac{\pi}{2}\right)\right) = \pi \times T(U(0, 1)) - \frac{\pi}{2}$$

$g_{vr \rightarrow wt}$: victim receiver antenna gain in the wanted transmitter direction

$$g_{vr \rightarrow wt} = f(g_{vr}^{max}, pattern_{vr}) = g_{vr}^{max} \times pattern_{vr}(\theta_{wt \rightarrow vr} + \pi, -\phi_{wt \rightarrow vr}, f_{vr})$$

Case of fixed distances:

$P_{wt}^{nominal}$: nominal power distribution

$f_{fading, fixed link}$: fading distribution

$$dRSS = f(P_{wt}^{nominal}, f_{fading, fixed link}) = T(P_{wt}^{nominal}) - T(f_{fading, fixed link})$$

Case of given $dRSS$: distribution to be given by the user.

b) $iRSS_{block}$ calculation

$$iRSS_{block} = \sum_{j=1}^{n_{interferers}} f\left(p_{it}^{supplied}, g_{it}^{PC}, g_{it \rightarrow vr}, pl_{it \leftrightarrow vr}, a_{vr}, g_{vr \rightarrow it}\right)_j = 10 \log \sum_{j=1}^{n_{interferers}} 10^{i_{block}/10}$$

where the j -th interferer signal is given by:

$$i_{block_j} \equiv \left(p_{it}^{supplied} + g_{it}^{PC} + g_{it \rightarrow vr}(f_{it}) - pl_{it \leftrightarrow vr} - a_{vr} + g_{vr \rightarrow it}(f_{it})\right)_j$$

where for each interferer:

f_{it} : interferer transmitting frequency

$$f_{it} = T(f_{it})$$

$p_{it}^{supplied}$: maximum power supplied to the interfering transmitter antenna (before power control)

$$p_{it}^{supplied} = T(p_{it}^{supplied})$$

g_{it}^{PC} : power control gain for the interfering transmitter

$$g_{it}^{PC} = f_{pc}\left(p_{it}^{supplied}, g_{it \rightarrow vr}, pl_{it \leftrightarrow vr}, g_{vr \rightarrow it}, pc_{it}^{t_hold}, pc_{it}^{dyc_rg}, pc_{it}^{st_rg}\right)$$

where:

f_{pc} : power control function (given in Appendix 2)

$pl_{it \leftrightarrow wr}$: path loss between the interfering transmitter and the wanted receiver (propagation loss, slow fading and clutter losses taken into account). Depending on the power control implementation, this can be either mean path loss or instantaneous path loss (Rayleigh fading excluded):

$$pl_{it \leftrightarrow wr} = f_{propag}(f_{it}, h_{wr}, h_{it}, d_{it \leftrightarrow wr}, env) + f_{clutter}(env)$$

or

$$pl_{it \leftrightarrow wr} = f_{mean}(f_{it}, h_{wr}, h_{it}, d_{it \leftrightarrow wr}, env) + f_{clutter}(env)$$

where:

h_{wr} : antenna height of wanted transmitter

$$h_{wr} = T(H_{wr})$$

$$\text{e.g.: } h_{wr} = T(U(h_{wr}^{min}, h_{wr}^{max})) = h_{wr}^{min} + (h_{wr}^{max} - h_{wr}^{min}) T(U(0, 1))$$

h_{it} : interfering transmitter antenna height

$$h_{it} = T(H_{it})$$

$$\text{e.g.: } h_{it} = T(U(h_{it}^{min}, h_{it}^{max})) = h_{it}^{min} + (h_{it}^{max} - h_{it}^{min}) T(U(0, 1))$$

$d_{it \leftrightarrow wr}$: distance between the interfering transmitter and the wanted receiver

$$d_{it \leftrightarrow wr} = T(R_{max}^{it})$$

$$\text{e.g.: } d_{it \leftrightarrow wr} = R_{max}^{it} \sqrt{T(U(0, 1))}$$

Three different choices for R_{max}^{it} are made:

Choice 1: Given distance R_{max}^{it}

Choice 2: Noise limited network.

R_{max}^{it} is determined by the following equation:

$$f_{median} = (f_{vr}, h_{wr}, h_{it} R_{max}^{it}, env) + f_{slowfading}(X\%) = P_{it}^{supplied} + g_{it}^{max} + g_{wr}^{max} - sens_{wr}^{max}$$

where:

f_{median} : propagation loss not including slow fading

$f_{slowfading}(X\%)$: fading margin to be used for 1-X% coverage loss.

In the case of lognormal fading and a 95% coverage loss at the edge of the coverage, for large distances, the value for slow fading is the well known 1.64 times the standard deviation of the propagation loss.

Note that in this case, formulas given for $f_{median}(R_{max}^{wt}) + \dots$ have to be inverted.

Note that in SEAMCAT the following simplifications are introduced:

- all antenna gains are set to 0 dBi, and
- the modified Hata model is applied as propagation model.

Choice 3: Traffic limited network

$$R_{max}^{it} = \sqrt{\frac{n_{channels} n_{userperchannel}}{\pi dens_{max} clusterfrequency}}$$

$g_{it \rightarrow wr}$: interfering transmitter antenna gain in the direction of the closest base station

$$g_{wr \rightarrow it} = f(g_{wr}^{max}, pattern_{wr}) = g_{wr}^{max} \times pattern_{wr}(\theta_{it \rightarrow wr} + \pi, -\varphi_{it \rightarrow wr}, f_{it})$$

where:

$(\theta_{it \rightarrow wr}, \varphi_{it \rightarrow wr})$: azimuth and elevation angles between the top of the interfering transmitter antenna and the top of the wanted receiver antenna

e.g.:

$$\theta_{it \rightarrow wr} = T(U(0, 2\pi)) = 2\pi \times T(U(0, 1))$$

$$\varphi_{it \leftrightarrow wr} = T\left(U\left(-\frac{\pi}{2}, \frac{\pi}{2}\right)\right) = \pi T \times (U(0, 1)) - \frac{\pi}{2}$$

$g_{wr \rightarrow it}$: base station antenna gain in the interfering transmitter direction

$$g_{wr \rightarrow it} = f(g_{wr}^{max}, pattern_{wr}) = g_{wr}^{max} \times pattern_{wr}(\theta_{it \rightarrow wr} + \pi, -\varphi_{it \rightarrow wr}, f_{it})$$

$pl_{it \leftrightarrow vr}$: path loss between the interfering transmitter i and the victim receiver (propagation loss, slow fading and clutter losses taken into account).

$$pl_{it \leftrightarrow vr} = f_{propag}(f_{it}, h_{vr}, h_{it}, d_{it \leftrightarrow vr}, env)$$

or

$$pl_{wt \leftrightarrow vr} = f_{median}(f_{vr}, h_{vr}, h_{wt}, d_{wt \leftrightarrow vr}, env)$$

The choice between f_{median} and f_{propag} would depend on the criteria of interference, and is closely related to the choice made for assessment of $dRSS$, e.g. whether ICE will evaluate:

$$\frac{dRSS_{mean}}{iRSS_{mean}}, \frac{dRSS_{propag}}{iRSS_{propag}}, \frac{dRSS_{mean}}{iRSS_{propag}};$$

where:

h_{vr} : victim receiver antenna height (defined in the $dRSS$ calculation)

h_{it} : interfering transmitter antenna height (defined previously)

$d_{it \leftrightarrow vr}$: distance between the victim receiver and the interfering transmitter.

Two different ways to choose $d_{it \leftrightarrow vr}$:

1. The most common case is when there is no spatial correlation between the elements of the victim system and the elements of the interfering system.

Then, $d_{it \leftrightarrow vr}$ is a result of a trial:

$$d_{it \leftrightarrow vr} = R_{simu} \sqrt{T(U(0, 1))}$$

where:

R_{simu} : radius of the area where interferers are spread

$$R_{simu} = \sqrt{\frac{n^{active}}{\pi dens_{it}^{active}}}$$

where:

n^{active} : number of active interferers considered in the simulation

n^{active} : should be sufficiently large so that the $n + 1$ interferer would bring a negligible additional interfering power

$$dens_{it}^{active} = dens_{it} \times p_{it}^{tx} \times temp_{it}(time)$$

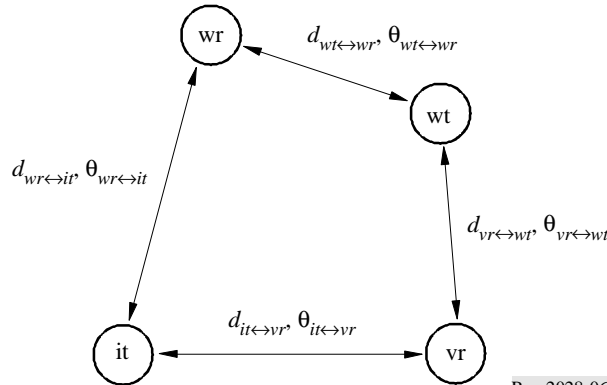
2. This case deals with the situation where the victim system and the interfering system are geographically correlated (e.g. co-located base stations).

This correlation is assumed to be only between one element (victim or wanted transmitter) of the victim system and one element (interferer or wanted receiver) of the interfering system.

A trial (if the distance is not fixed) of the distances and angles between the two correlated elements is made (e.g. $d_{wr \leftrightarrow vr}$, $\theta_{wr \leftrightarrow vr}$). The knowledge of $d_{it \leftrightarrow wr}$, $d_{vr \leftrightarrow wt}$, $\theta_{it \leftrightarrow wr}$, $\theta_{vr \leftrightarrow wt}$ enables to derive the missing co-ordinates (e.g. $d_{it \leftrightarrow vr}$, $\theta_{it \leftrightarrow vr}$).

FIGURE 6

Interfering scenario with a geographical correlation between the victim and the interfering systems



Rap 2028-06

$g_{it \rightarrow vr}(f_{it})$: interfering transmitter antenna gain in the victim receiver direction

$$g_{it \rightarrow vr} = f(g_{it}^{max}, pattern_{it}) = g_{it}^{max} \times pattern_{it}(\theta_{it \rightarrow vr}, \phi_{it \rightarrow vr}, f_{it})$$

where:

$(\theta_{it \rightarrow vr}, \varphi_{it \rightarrow vr})$: azimuth and elevation angles between the top of the closest interfering transmitter antenna and the top of the victim receiver antenna

e.g.: $\theta_{it \rightarrow vr} = T(U(0, 2\pi)) = 2\pi \times T(U(0, 1))$

$$\varphi_{it \leftrightarrow vr} = T\left(U\left(-\frac{\pi}{2}, \frac{\pi}{2}\right)\right) = \pi \times T(U(0, 1)) - \frac{\pi}{2}$$

$a_{vr}(f_{it}, f_{vr})$: attenuation of the victim receiver.

Three possible ways are considered for calculating this attenuation:

1. a_{vr} is given by the user.
2. Blocking is given in terms of blocking attenuation or protection ratio. For a wanted signal 3 dB above the sensitivity, the attenuation a_{vr} can be derived from the following equation (see Appendix 7):

$$a_{vr} = f\left(\frac{C}{N+I}, block_{att}\right) = 3 + \frac{C}{N+I} + block_{att}(f_{it}, f_{vr})$$

3. Blocking is given in terms of absolute level of blocking:

$$a_{vr} = f\left(\frac{C}{N+I}, block_{abs}\right) = \frac{C}{N+I} + block_{abs}(f_{it}, f_{vr}) - sens_{vr}$$

Two cases are envisaged:

Case 1: *block* is a mask which is a function of $\Delta f = (f_{it} - f_{vr})$. It is introduced to enable calculations of interference between systems in adjacent bands;

Case 2: *block* is a fixed value (e.g. 80 dBm). It is used to derive generic limits.

$g_{vr \rightarrow it}(f_{it})$: victim receiver antenna gain in the interfering transmitter direction

$$g_{vr \rightarrow it} = f(g_{vr}^{max}, pattern_{vr}) = g_{vr}^{max} \times pattern_{vr}(\theta_{it \rightarrow vr}, \varphi_{it \rightarrow vr}, f_{it})$$

c) $iRSS_{spur}$ calculation

$$iRSS_{spur} = f(spur, g_{it}^{pc}, g_{it \rightarrow vr}, pl_{it \rightarrow vr}, g_{vr \rightarrow it}) = 10 \log \sum_{j=1}^{n_{interferers}} 10^{i_{spurj}/10}$$

where the j -th interferer signal is defined as:

$$i_{spurj} = (spur(f_{it}, f_{vr}) + g_{it \rightarrow vr}(f_{vr}) - pl_{it \rightarrow vr}(f_{vr}) + g_{vr \rightarrow it}(f_{vr}))_j$$

f_{it} : interferer transmitting frequency (defined in b))

$spur(f_{it}, f_{vr}, g_{pc})$: unwanted emission by the interfering transmitter.

Two cases are envisaged:

Case 1: *spur* is a mask which is a function of $\Delta f = (f_{it} - f_{vr})$. It is introduced to enable calculations of interference between systems in adjacent band.

Case 2: *spur* is a fixed value (e.g. -36 dBm). It is used to derive generic limits.

$spur(f_{it}, f_{vr}, g_{pc})$ generally depends on the effect of the power control. Either:

$$spur(f_{it}, f_{vr}, g_{pc}) = \max(spur_0(f_{it}, f_{vr}), spur(f_{it}, f_{vr}) - g_{pc})$$

or $spur(f_{it}, f_{vr}, g_{pc})$ is defined as a function of $\Delta f = (f_{it} - f_{vr})$ for each possible step of the power control.

g_{it}^{pc} : power control gain for the interfering transmitter (defined in b))

$pl_{it \leftrightarrow vr}$: path loss between the interfering transmitter and the victim receiver (propagation loss, slow fading and clutter losses taken into account)

$$pl_{it \leftrightarrow vr} = f_{propag}(f_{vr}, h_{vr}, h_{it}, d_{it \leftrightarrow vr}, env) + f_{clutter}(env)$$

where:

h_{vr} : victim receiver antenna height (defined in *dRSS* calculation)

h_{it} : interfering transmitter antenna height (defined in b))

$d_{it \leftrightarrow vr}$: distance between the victim receiver and the interfering transmitter (defined in b))

$g_{it \rightarrow vr}(f_{vr})$: interfering transmitter antenna gain in the victim receiver direction:

$$g_{it \rightarrow vr}(f_{vr}) = (g_{it}^{max}, pattern_{it}) = g_{it}^{max} \times pattern_{it}(\theta_{it \rightarrow vr}, \varphi_{it \rightarrow vr}, f_{vr})$$

where:

$(\theta_{it \rightarrow vr}, \varphi_{it \rightarrow vr})$: azimuth and elevation angles between the top of the closest interfering transmitter antenna and the top of the victim receiver antenna (defined in b))

$g_{vr \rightarrow it}(f_{vr})$: victim receiver antenna gain in the interfering transmitter direction

$$g_{vr \rightarrow it}(f_{vr}) = (g_{vr}^{max}, pattern_{vr}) = g_{vr}^{max} \times pattern_{vr}(\theta_{vr \rightarrow it} + \pi, -\varphi_{vr \rightarrow it}, f_{vr})$$

d) *iRSS*_{intermod} calculation

$$iRSS_{intermod} = f(p_{it,k}^{supplied}, g_{it,k}^{pc}, g_{it,k \rightarrow vr}, pl_{it,k \rightarrow vr}, g_{vr \rightarrow it,k}, sens_{vr}, intermod) \quad \text{with } k = i, j$$

$$= 10 \log \sum_{i=1}^n \sum_{j=1, j \neq i}^n 10^{i,j RSS_{intermod}/10}$$

where:

$i, j RSS_{intermod}$: Intermodulation product of third order at the frequency f_0

$$i_{i,j}RSS_{intermod} = 2i_iRSS_{int} + i_jRSS_{int} - 3intermod - 3sens_{vr} - 9 \quad \text{dB}$$

The interferer i transmits at the frequency $f_{it,i} = f_{it}$ and the interferer j at the frequency $f_{it,j}$ (see b)), which defines $\Delta f = f_{it,j} - f_{it}$ and yields $f_0 = f_{it} - \Delta f = 2f_{it} - f_{it,j}$. Assuming an ideal filter (roll off factor 0) the intermodulation product has to be considered only for the bandwidth b :

$$f_{vr} - b/2 \leq f_0 \leq f_{vr} + b/2$$

For all other cases the intermodulation product can be neglected.

i_kRSS_{int} : received power in the victim receiver due to interferer $k = i$ at f_{it} or interferer $k = j$ at $f_{it,j}$

$$i_kRSS_{int} = P_{it,k}^{supplied}, g_{it,k}^{pc}, g_{it,k \rightarrow vr}, pl_{it,k \leftrightarrow vr}, g_{vr \rightarrow it,k}$$

The various parameters are defined in the previous a) to c). For the computation of i_iRSS_{int} the same algorithms as given in Appendix 6 can be used because i_iRSS_{int} corresponds to $i_iRSS_{block} + a_{vr}(f_{it}, f_{vr})$.

intermod: receiver intermodulation response for a wanted signal 3 dB above the sensitivity.

Two cases are envisaged:

Case 1: *intermod* is given by the user, e.g. typical values are 70 dB for base station equipment and 65 dB for mobile and handportable equipment. It is used to derive generic limits.

Case 2: *intermod*(Δf) is measured as a function of Δf referred to f_{vr} (see Appendix 9)

sens_{vr}: sensitivity of victim receiver.

APPENDIX 1

TO ANNEX 2

[Knuth, 1969]

Propagation model

A number of propagation models are provided in the tool. They are depending on the environment chosen for the scenarios:

- general environment: open area, suburban or urban area;
- environment for the interferers: indoor or outdoor;
- environment for the victim receiver: indoor or outdoor.

The domain of validity for the models is described in Table 1.

TABLE 1

Below 30 MHz	No model available. Curves of Recommendation ITU-R P.368 are suited for high power transmitters and large distances and are therefore not adapted to interference calculations
Between 30 MHz and 3 GHz	Modified Hata model available for outdoor-outdoor path loss calculations. Care should be taken when propagation distances are expected to be above 20 km. Indoor-indoor and indoor-outdoor models also suitable
Above 3 GHz	Modified Hata model not advised. Spherical diffraction model is suitable for open area environment. No model available for suburban and urban environment. Indoor-indoor and indoor-outdoor models also suitable

To improve the flexibility of the tool, a “generic” model, e.g. $L = A + B \log(d) + C d$, both for the wanted signal path and the interfering path d can also be entered by the user. The user of the tool is then to enter the parameters A , B , C of the median attenuation formula and the distribution of the variation in path loss D_v . As a default distribution, a lognormal distribution is to be proposed with a standard deviation to be entered by the user. Then we have:

$$f_{propag}(d) = L + T(D_v)$$

Also, more elaborate models can be implemented by the user using a simple script. An example of the free space path loss and for the script used is given in the following:

$$L = 32.5 + 20 \log(f \text{ [MHz]}) + 20 \log(d \text{ [km]})$$

and in SEAMCAT software by

```
L1 = 92,5;
L2 = 20 * log10(freq()/1000);
L3 = 20 * log10(dist());
L = L1 + L2 + L3;
eval L.
```

1 Modified Hata model

$$f_{propag}(f, h_1, h_2, d, env) = L + T(G(\sigma))$$

where:

- L : median propagation loss (dB)
- σ : standard deviation of the slow fading distribution (dB)
- f : frequency (MHz)
- H_m : $\min\{h_1, h_2\}$
- H_b : $\max\{h_1, h_2\}$
- d : distance (km), preferably less than 100 km
- env : (outdoor/outdoor), (rural, urban or suburban), (propagation above or below roof).

If H_m and/or H_b are below 1 m, a value of 1 m should be used instead. Antenna heights above 200 m might also lead to significant errors. Propagation below roof means that both H_m and H_b are above the height of roofs. Propagation is above roof in other cases (H_b above the height of roofs).

1.1 Calculation of the median path loss L

Case 1: $d \leq 0.04$ km

$$L = 32.4 + 20 \log(f) + 10 \log(d^2 + (H_b - H_m)^2 / 10^6)$$

Case 2: $d \geq 0.1$ km

$$a(H_m) = (1.1 \log(f) - 0.7) \min\{10, H_m\} - (1.56 \log(f) - 0.8) + \max\{0, 20 \log(H_m / 10)\}$$

$$b(H_b) = \min\{0, 20 \log(H_b / 30)\}$$

$$\alpha = \begin{cases} 1 & \text{for } d \leq 20 \text{ km} \\ 1 + \left(0.14 + 1.87 \times 10^{-4} f + 1.07 \times 10^{-3} H_b\right) \left(\log \frac{d}{20}\right)^{0.8} & \text{for } 20 \text{ km} < d \leq 100 \text{ km} \end{cases}$$

Sub-case 1: Urban

30 MHz $< f \leq$ 150 MHz

$$L = 69.6 + 26.2 \log(150) - 20 \log(150 / f) - 13.82 \log(\max\{30, H_b\}) + \alpha[44.9 - 6.55 \log(\max\{30, H_b\})] \log(d) - a(H_m) - b(H_b)$$

150 MHz $< f \leq$ 1 500 MHz

$$L = 69.6 + 26.2 \log(f) - 13.82 \log(\max\{30, H_b\}) + \alpha[44.9 - 6.55 \log(\max\{30, H_b\})] \log(d) - a(H_m) - b(H_b)$$

1 500 MHz $< f \leq$ 2 000 MHz

$$L = 46.3 + 33.9 \log(f) - 13.82 \log(\max\{30, H_b\}) + \alpha[44.9 - 6.55 \log(\max\{30, H_b\})] \log(d) - a(H_m) - b(H_b)$$

2 000 MHz $< f \leq$ 3 000 MHz

$$L = 46.3 + 33.9 \log(2000) + 10 \log(f / 2000) - 13.82 \log(\max\{30, H_b\}) + \alpha[44.9 - 6.55 \log(\max\{30, H_b\})] \log(d) - a(H_m) - b(H_b)$$

Sub-case 2: Suburban

$$L = L(\text{urban}) - 2\{\log[(\min\{\max\{150, f\}, 2000\}) / 28]\}^2 - 5.4$$

Sub-case 3: Open area

$$L = L(\text{urban}) - 4.78 \{\log[\min\{\max\{150, f\}, 2000\}]\}^2 + 18.33 \log[\min\{\max\{150, f\}, 2000\}] - 40.94$$

Case 3: 0.04 km $< d <$ 0.1 km

$$L = L(0.04) + \frac{[\log(d) - \log(0.04)]}{[\log(0.1) - \log(0.04)]} [L(0.1) - L(0.04)]$$

When L is below the free space attenuation for the same distance, the free space attenuation should be used instead.

1.2 Assessment of the standard deviation for the lognormal distribution

Case 1: $d \leq 0.04$ km
 $\sigma = 3.5$ dB

Case 2: $0.04 \text{ km} < d \leq 0.1 \text{ km}$

$$\sigma = 3.5 + \frac{(12 - 3.5)}{(0.1 - 0.04)} (d - 0.04) \quad \text{dB} \quad \text{for propagation above the roofs}$$

$$\sigma = 3.5 + \frac{(17 - 3.5)}{(0.1 - 0.04)} (d - 0.04) \quad \text{dB} \quad \text{for propagation below the roofs}$$

Case 3: $0.1 \text{ km} < d \leq 0.2 \text{ km}$

$$\sigma = 12 \text{ dB} \quad \text{for propagation above the roofs}$$

$$\sigma = 17 \text{ dB} \quad \text{for propagation below the roofs}$$

Case 4: $0.2 \text{ km} < d \leq 0.6 \text{ km}$

$$\sigma = 12 + \frac{(9 - 12)}{(0.6 - 0.2)} (d - 0.2) \quad \text{dB} \quad \text{for propagation above the roofs}$$

$$\sigma = 17 + \frac{(9 - 17)}{(0.6 - 0.2)} (d - 0.2) \quad \text{dB} \quad \text{for propagation below the roofs}$$

Case 5: $0.6 \text{ km} < d$

$$\sigma = 9 \text{ dB}$$

2 Spherical diffraction model

The spherical propagation model is based on Recommendations ITU-R P.452, ITU-R P.676 and ITU-R P.526¹.

According to Recommendation ITU-R P.452 the median loss between transmitter and receiver is given by the following equation:

$$L_{bd}(p) = 92.5 + 20 \log f + 20 \log d + L_d(p) + A_g$$

where:

$L_{bd}(p)$: basic loss (dB) as function of the time percentage, p (%)

f : frequency (GHz)

d : distance (km)

$L_d(p)$: diffraction loss (dB) as function of the time percentage, p (%)

A_g : attenuation due to atmospheric gas and water (dB)

¹ The used documentation is based on documents published in 1990-1994. In the meantime newer Recommendations are available. Unfortunately some of the useful information were shifted to Reports or other Recommendations.

The attenuation due to atmosphere is given by:

$$A_g = [\gamma_o(f) + \gamma_w(\rho, f)]d$$

where:

$\gamma_o(f)$: linear attenuation due to dry air (oxygen) (dB/km)

$\gamma_w(\rho, f)$: linear attenuation (dB/km) due to water as function of the water concentration ρ (g/m³), default value: 3 g/m³.

Both terms can be approximated by the following equations according to Recommendation ITU-R P.676:

– Attenuation due to water:

$$\gamma_w(\rho, f) = \left[0.050 + 0.0021\rho + \frac{3.6}{(f - 22.2)^2 + 8.5} + \frac{10.6}{(f - 183.3)^2 + 9} + \frac{8.9}{(f - 325.4)^2 + 26.3} \right] f^2 \rho \times 10^{-4} \quad \text{for } f < 350 \text{ GHz}$$

– Attenuation due to oxygen:

$$\gamma_o(f) = \left[7.19 \times 10^{-3} + \frac{6.09}{f^2 + 0.227} + \frac{4.81}{(f - 57)^2 + 1.50} \right] f^2 \times 10^{-3} \quad \text{for } f \leq 57 \text{ GHz}$$

$$\gamma_o(f) = 10.5 + 1.5(f - 57) \quad \text{for } 57 < f \leq 60 \text{ GHz}$$

$$\gamma_o(f) = 15 - 1.2(f - 60) \quad \text{for } 60 < f \leq 63 \text{ GHz}$$

$$\gamma_o(f) = \left[3.79 \times 10^{-7} f + \frac{0.265}{(f - 63)^2 + 1.59} + \frac{0.028}{(f - 118)^2 + 1.47} \right] (f + 198)^2 \times 10^{-3} \quad \text{for } f > 63 \text{ GHz}$$

Note that for simplification a linear interpolation between 57 and 63 GHz is used. The maximum is 15 dB/km for 60 GHz.

According to Recommendation ITU-R P.526, the diffraction loss $L_d(p)$ can be derived by the received field strength E referred to the free space E_0 :

$$-L_d(p) = 20 \log \frac{E}{E_0} = F(X) + G(Y_1) + G(Y_2)$$

where:

X : normalized radio path between transmitter and receiver

Y_1 : normalized antenna height of the transmitter

Y_2 : normalized antenna height of the receiver

$$X = 2.2 \beta f^{1/3} a_e^{-2/3} d$$

$$Y = 9.6 \times 10^{-3} \beta f^{2/3} a_e^{-1/3} h_i$$

where:

β : parameter derived from the earth admittance factor K : $\beta = 1$ for $f > 20$ MHz

f : frequency (MHz)

a_e : equivalent earth radius in (km) (definition see below)

d : distance (km)

h_i : antenna height above ground (m) with $i = 1$ or 2 for the transmitter or receiver, respectively.

The distance-dependent term $F(X)$ is given by the semi-empirical formula:

$$F(X) = 11 + 10 \log(X) - 17.6X$$

The antenna height gain $G(Y)$ is given by the formula set:

$$\begin{aligned} G(Y) &= 17.6(Y - 1.1)^{1/2} - 5 \log(Y - 1.1) - 8 && \text{for } Y > 2 \\ G(Y) &= 20 \log(Y + 0.1Y^3) && \text{for } 10 K < Y < 2 \\ G(Y) &= 2 + 20 \log K + 9 \log(Y/K) [\log(Y/K) + 1] && \text{for } K/10 < Y < 10 K \\ G(Y) &= 2 + 20 \log K && \text{for } Y < K/10 \end{aligned}$$

where:

K : normalized earth surface admittance factor (see Recommendation ITU-R P.526), default value: 10^{-5} .

Note that different units for the frequency are used.

This variation in path loss is provided through the variability of the equivalent earth radius a_e (km) which is considered to be dependent on the time percentage, p :

$$a_e(p) = 6375 k(p)$$

with the earth radius factor $k(p)$ expressed as:

$$k(p) = k_{50} + (5 - k_{50}) \frac{(1.7 - \log p)}{(1.7 - \log \beta_0)} \quad \text{for } p < 50\%$$

$$k(p) = k_{50} \quad \text{for } p > 50\%$$

and

$$k_{50} = \frac{157}{157 - \Delta N}$$

where:

ΔN : mean gradient of the radio refraction profile over a 1 km layer of the atmosphere from the surface. The default value is 40 units/km for Europe (standard atmosphere). This value yields to $k_{50} \approx 4/3$ y $a_e = 8500$ km.

NOTE 1 – The mean gradient is positive.

β_0 : existence probability (%) of the super-refractive layer ($\Delta N > 100$ units/km) in the low atmosphere. Default value: 1% for Europe.

Note that the probabilities p and β_0 are denoted in %, i.e. a range of variety: 0 ... 100%.

Note that in SEAMCAT, $p = 50\%$; is currently implemented, i.e. the median with respect to time is computed. Later versions may allow to choose a certain time percentage.

The following restrictions of application of this model are to be considered:

- The frequency range should be larger than 3 GHz, with caution lower frequencies may be used but not below 300 MHz due to the surface admittance and polarization effects.
- The model was developed for open (rural) area. Therefore, the additional attenuation due to obstacles like buildings found in suburban or urban environment is not included.
- The loss due to rain is not covered.
- This model is applicable only for terrestrial radio paths.

3 Combined indoor-outdoor propagation models

Most of the published propagation models are derived either for outdoor or indoor application. But in the “real world” a combination of both types is required.

In SEAMCAT, the classical outdoor models, Hata (SE21 version) and spherical diffraction model (Recommendations ITU-R P.452, ITU-R P.526 and ITU-R P.676), are combined with an indoor model. An illustrative description is given in the following.

The path loss p_L consists of median path loss L and the Gaussian variation $T(G(\sigma))$ where σ is the standard deviation:

$$p_L(f, h_1, h_2, d, env) = L + T(G(\sigma))$$

where:

- f : frequency (MHz)
- h_i : antenna height of the transmitter antenna (m)
- h_2 : antenna height of the receiver antenna (m)
- d : distancia (km)
- env : parameter for the environments of the transmitter and receiver.

For outdoor-outdoor holds:

- Scenario: transmitter and receiver are both outdoor.
- Modified Hata model:
Median: $L(outdoor-outdoor) = L_{Hata}(outdoor-outdoor)$
Variation: intrinsic variation, $\sigma(outdoor-outdoor) = \sigma_{Hata}$
- Spherical diffraction model
Median: $L(outdoor-outdoor) = L_{spherical}$
Variation: no variation possible, $\sigma(outdoor-outdoor) = 0$

Case 1: Indoor-outdoor or outdoor-indoor

- Scenario: transmitter is indoor and receiver is outdoor, or vice versa
- Modified Hata model:

$$\text{Median: } L(indoor-outdoor) = L_{Hata}(outdoor-outdoor) + L_{we}$$

where L_{we} is the attenuation due to external walls (default value = 10 dB).

$$\text{Variation: } \sigma(\text{indoor-outdoor}) = \sqrt{\sigma_{\text{Hata}}^2 + \sigma_{\text{add}}^2}$$

where σ_{add} is the additional standard deviation of the signal (default value: 5 dB).

The standard deviation of the lognormal distribution is increased, compared to the outdoor-outdoor scenario due to additional uncertainty on materials and relative location in the building.

– Spherical diffraction model

$$\text{Median: } L(\text{indoor-outdoor}) = L_{\text{spherical}} + L_{\text{we}}$$

$$\text{Variation: } \sigma(\text{indoor-outdoor}) = \sigma_{\text{add}}$$

The lognormal distribution is determined by the additional variation due to the variation in building materials, for the spherical diffraction model no variation is considered.

Case 2: Indoor-indoor

There are two different scenarios possible: The transmitter and receiver are in the same or in different buildings. The scenario used by SEAMCAT is randomly selected.

a) Selection of the scenario

The first step is to determine whether the indoor-indoor scenario corresponds to the transmitter and receiver in the same building or not. This is done by the calculation of the random variable in the same building (SB).

Trial of SB condition:

– $d < 0.020$ km (20 m): SB = Yes $\Rightarrow P(\text{Yes}) = 1$

– 0.020 km $< d < 0.050$ km (50 m):

$$\text{SB} = \text{Yes} \quad P(\text{Yes}) = (0.050 - d)/0.030$$

$$\text{SB} = \text{No} \quad P(\text{No}) = 1 - P(\text{Yes}) = (d - 0.020)/0.030$$

– $d > 0.050$ km (50 m): SB = Yes $\Rightarrow P(\text{Yes}) = 0$

b) Indoor-indoor, different buildings

– Scenario: transmitter and receiver in different buildings: $P(\text{Yes}) = 0$ or $P(\text{No}) = 1$

– Modified Hata model:

$$\text{Median: } L(\text{indoor-indoor}) = L_{\text{Hata}}(\text{outdoor-outdoor}) + 2L_{\text{we}}$$

It is to be noted that the loss due to two external walls should be added.

$$\text{Variation: } \sigma(\text{indoor-indoor}) = \sqrt{\sigma_{\text{Hata}}^2 + 2\sigma_{\text{add}}^2}$$

– Spherical diffraction model

$$\text{Median: } L(\text{indoor-indoor}) = L_{\text{spherical}} + 2L_{\text{we}}$$

$$\text{Variation: } \sigma(\text{indoor-indoor}) = \sqrt{2}\sigma_{\text{add}}$$

The lognormal distribution is determined by the additional variation due to the variation in building materials, for the spherical diffraction model no variation is considered. The variation is increased for the second external wall.

c) Indoor-indoor, same building

- Scenario: transmitter and receiver in the same building: $P(\text{Yes}) = 1$ or $P(\text{No}) = 0$
- Indoor propagation model:

Median:

$$L(\text{indoor-indoor}) = -27.6 + 20 \log(1000d) + 20 \log(f) + \text{fix}\left(\frac{1000d}{d_{\text{room}}}\right) L_{wi} + k_f \left[\frac{k_f + 2}{k_f + 1} - b \right] L_f$$

with: $k_f = \text{fix}\left(\frac{|h_2 - h_1|}{h_{\text{floor}}}\right)$

L_{wi} : loss of internal wall (dB) (default value = 5 dB)

L_f : loss between adjacent floor (dB) (default value = 18.3 dB)

b : empirical parameter (default value = 0.46)

d_{room} : size of the room (m) (default value = 4 m)

h_{floor} : height of each floor (m) (default value = 3 m)

Note that the path length d uses the unit km and the frequency the unit MHz

Variation: $\sigma(\text{indoor-indoor}) = \sigma_{in}$

The lognormal distribution trial is made using a standard deviation entered by the user and covering the variation, internal in the building, due to building design, in furniture of the rooms, etc. The default value is $\sigma_{in} = 10$ dB.

APPENDIX 2

TO ANNEX 2

Power control function

$$g_{it}^{PC} = f_{pc}(p_{it}^{supplied}, g_{it \rightarrow wr}, pl_{it \leftrightarrow wr}, g_{wr \rightarrow it}, pc_{it}^{t_hold}, pc_{it}^{dyc_rg}, pc_{it}^{st_rg})$$

$$P = f(p_{it}^{supplied}, g_{it \rightarrow wr}, pl_{it \leftrightarrow wr}, g_{wr \rightarrow it}) = p_{it}^{supplied} + g_{it \rightarrow wr} - pl_{it \leftrightarrow wr} + g_{wr \rightarrow it}$$

P : power received by the wanted receiver, e.g. closest base station of the interfering system

where $p_{it}^{supplied}$, $g_{it \rightarrow wr}$, $g_{wr \rightarrow it}$ and $pl_{it \leftrightarrow wr}$ are defined in the $iRSS$ calculation sections.

Case 1: $P \leq pc_{it}^{t_hold}$

$$p_{it}^{supplied_PC} = p_{it}^{supplied}$$

$$g_{it}^{PC} = 0$$

Case (i + 1): $pc_{it}^{t_hold} + (i - 1) \cdot pc_{it}^{st_rg} \leq P < pc_{it}^{t_hold} + i \cdot pc_{it}^{st_rg}$

$$p_{it}^{supplied_PC} = p_{it}^{supplied} - (i - 1) \cdot pc_{it}^{st_rg}$$

$$g_{it}^{PC} = -(i - 1) \cdot pc_{it}^{st_rg}$$

where i is an integer ranging from 1 to $n_steps = \frac{pc_{it}^{dyc_rg}}{pc_{it}^{st_rg}}$

Case ($n_steps + 2$): $P > pc_{it}^{t_hold} + pc_{it}^{dyc_rg}$

$$p_{it}^{supplied_PC} = p_{it}^{supplied} - pc_{it}^{dyc_rg}$$

$$g_{it}^{PC} = -pc_{it}^{dyc_rg}$$

APPENDIX 3

TO ANNEX 2

Distribution definitions

- Uniform distribution: $U(0, 1) = \begin{cases} 1 & \text{if } 0 \leq x \leq 1 \\ 0 & \text{otherwise} \end{cases}$
- Gaussian distribution: $G(\sigma) = \frac{1}{\sqrt{2\pi}\sigma} \exp\left(-\frac{x^2}{2\sigma^2}\right)$
- Rayleigh distribution: $R(\sigma) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right)$
- User defined distribution: The option to include an user-defined distribution in the tool should be considered.

APPENDIX 4

TO ANNEX 2

Pseudo-random number generation

[Knuth, 1969; Rubinstein, 1981]

- From a uniform distribution $U(0, 1)$

$$u_{i+1} = T(U(0, 1)) = \frac{x_i + 1}{m}$$

where:

$$x_{i+1} = (a \cdot x_i) \pmod{m}$$

a : multiplier, e.g. $a = 16\,807$ or $396\,204\,094$ or $950\,706\,376$

m : modulus, e.g. $m = 2^{31} - 1 = 2\,147\,483\,647$

x_0 : seed, integer variable taking a value between 1 and $(m - 1)$

- From a Gaussian distribution $G(\sigma)$

$$T(G(\sigma)) = v_1 \sqrt{\frac{-2 \ln(s)}{s}}$$

where:

$$\text{while } s \geq 1, d_0 \begin{cases} v_1 = 2 \cdot T_{seed1}(U(0, 1)) - 1 \\ v_2 = 2 \cdot T_{seed2}(U(0, 1)) - 1 \\ s = v_1^2 + v_2^2 \end{cases}$$

v_1 and v_2 are two independent random variables (using two different seeds) uniformly distributed between -1 and $+1$.

- From a Rayleigh distribution $R(\sigma)$

$$T(R(\sigma)) = \sqrt{(v_1^2 + v_2^2)} \times \frac{-2 \ln(s)}{s}$$

where:

$$\text{while } s \geq 1, d_0 \begin{cases} v_1 = 2 \cdot T_{seed1}(U(0, 1)) - 1 \\ v_2 = 2 \cdot T_{seed2}(U(0, 1)) - 1 \\ s = v_1^2 + v_2^2 \end{cases}$$

v_1 and v_2 are two independent random variables (using two different seeds) uniformly distributed between -1 and $+1$.

From any type of distribution with a given cumulative distribution function, cdf .

Some trials may be performed according to a user-defined distribution F .

Trial is based on the use of the reciprocal cumulative distribution function, cdf^{-1} , relative to the user-defined distribution, F , applied to the result of a uniform sample between 0 and 1.

$$T(F) = cdf^{-1}(p) \quad \text{where} \quad p = T(U(0, 1)) \quad (\text{uniform trial between 0 and 1})$$

FIGURE 7

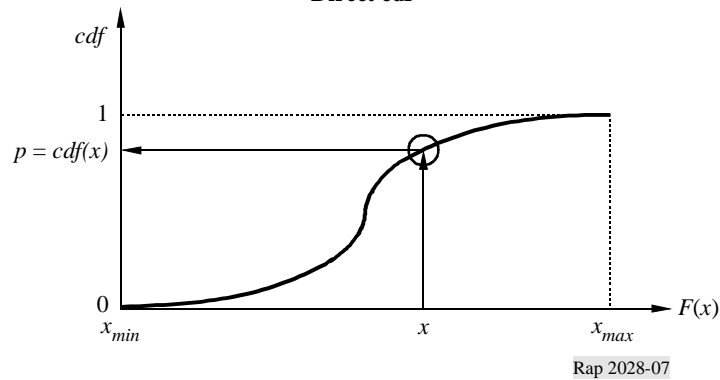
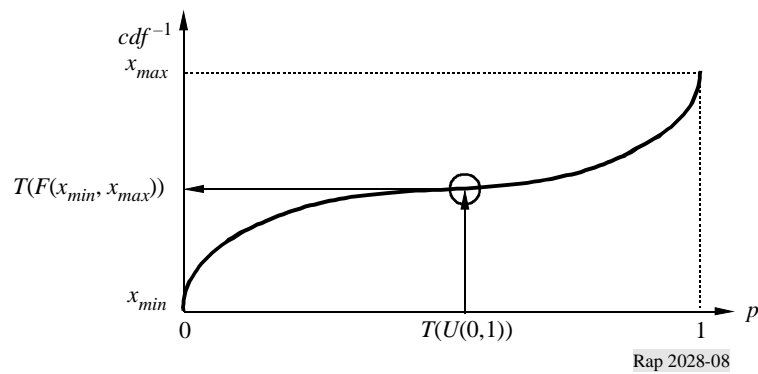
Direct cdf

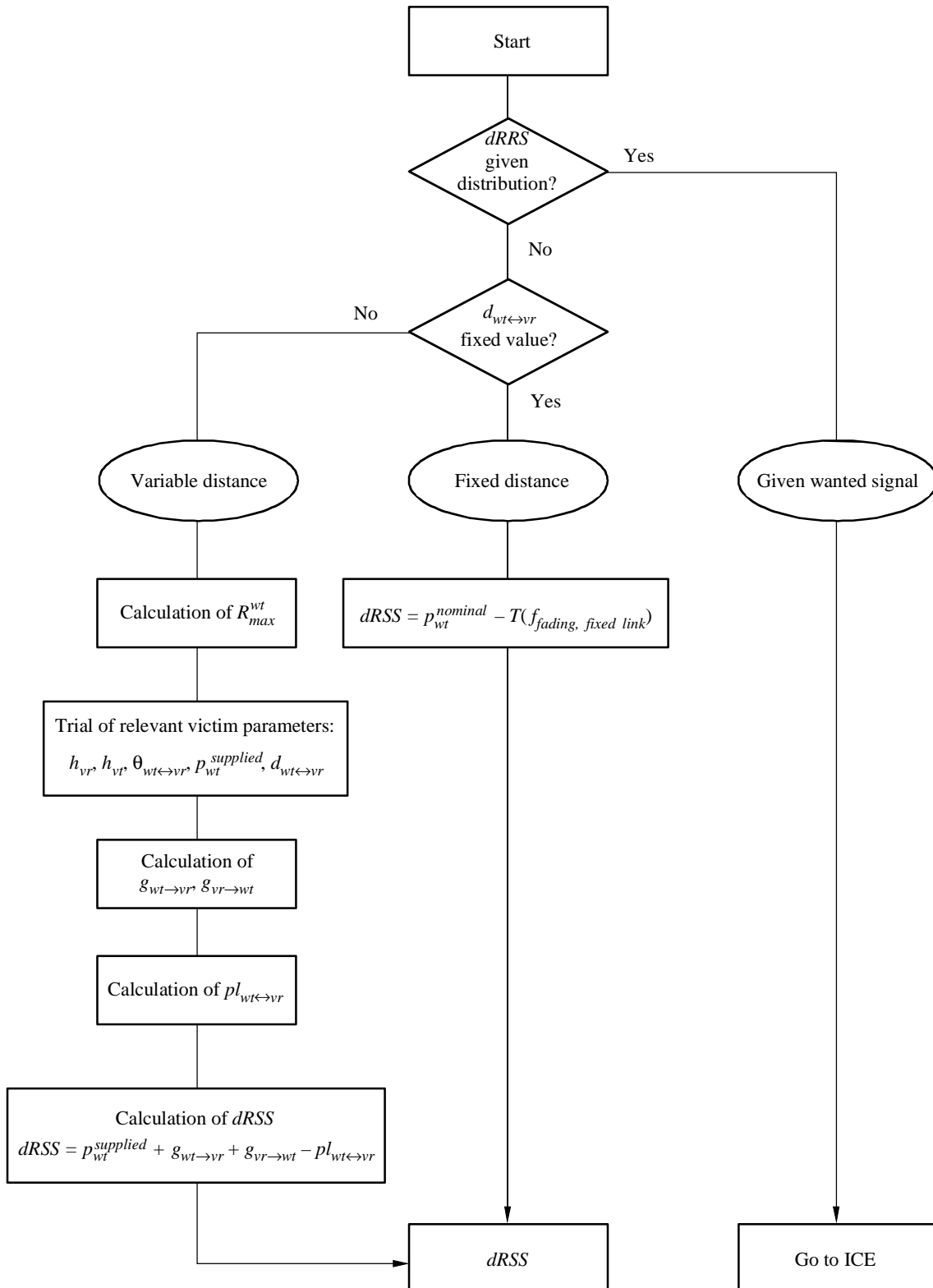
FIGURE 8

Inverse cdf

APPENDIX 5

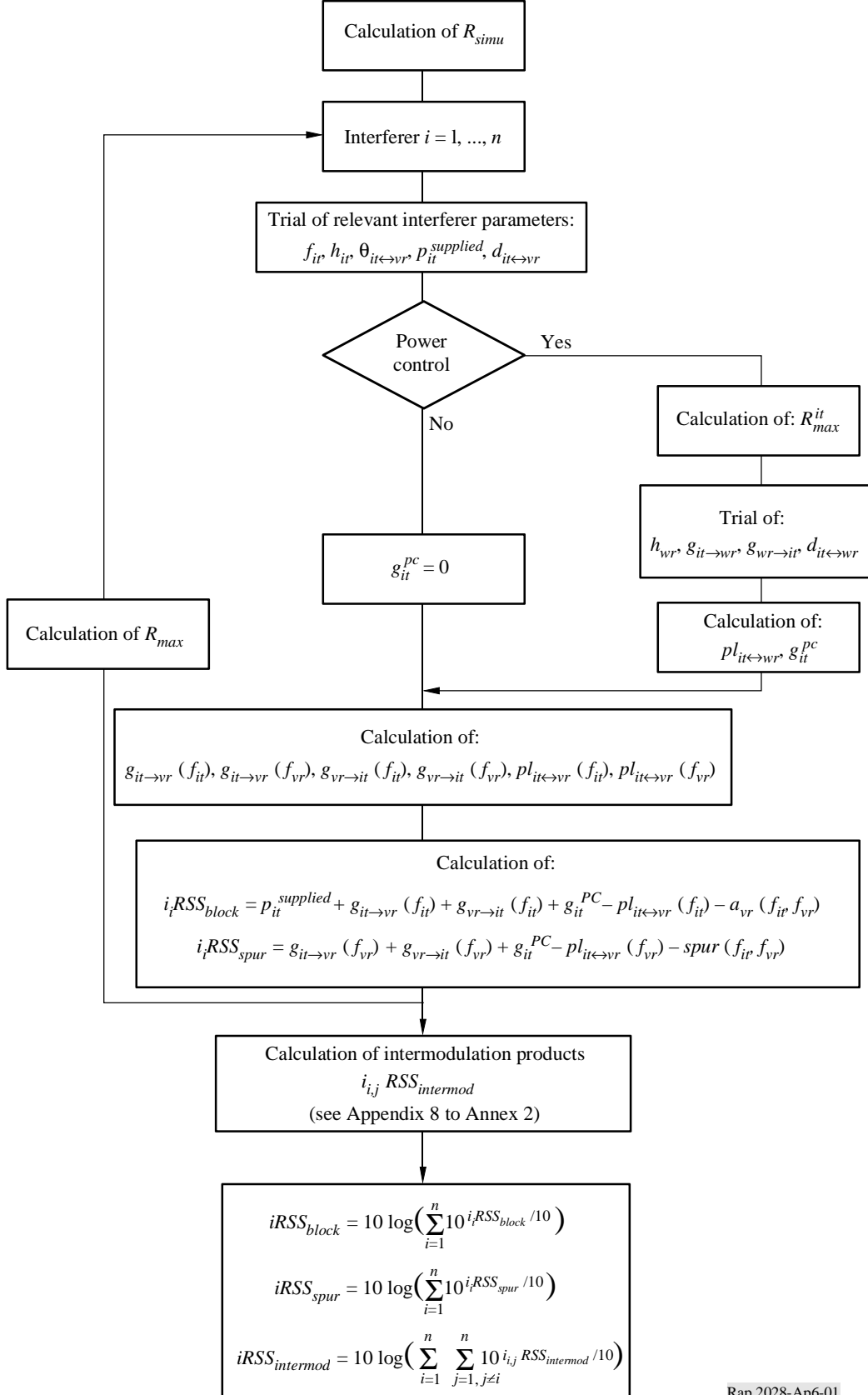
TO ANNEX 2

***dRSS* calculation flow chart**



APPENDIX 6

TO ANNEX 2

***i*RSS due to unwanted and blocking calculation**

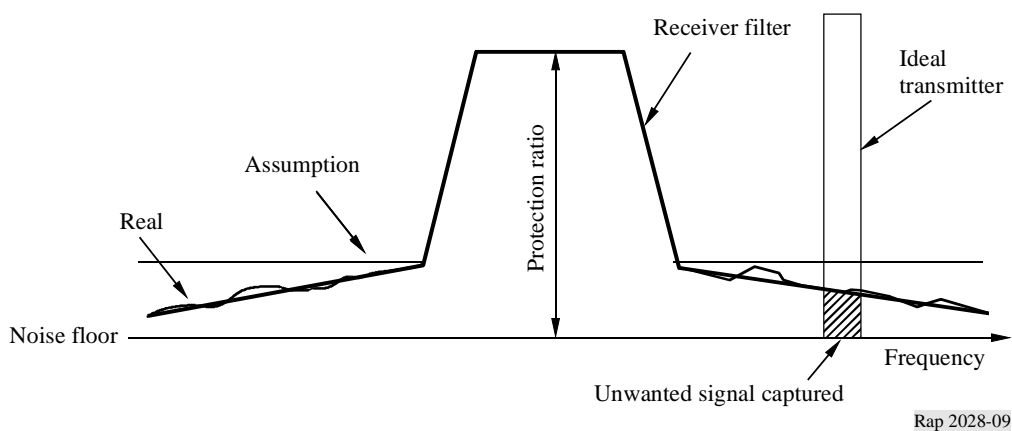
APPENDIX 7

TO ANNEX 2

Receiver Blocking**1 Basic concept**

The receiver is capturing some unwanted signal because its filter is not ideal.

FIGURE 9
Basic concept



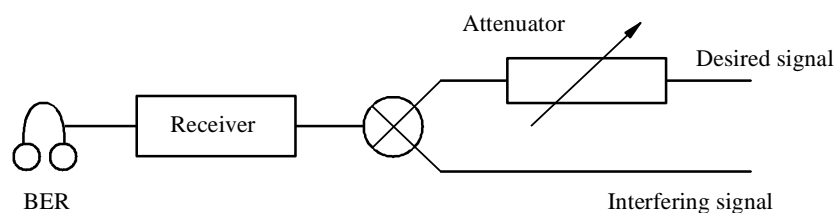
Rap 2028-09

Definition: Blocking is a measure of the capability of the receiver to receive a modulated wanted input signal in the presence of an unwanted input signal on frequencies other than those of the spurious responses or the adjacent channels, without these unwanted input signals causing a degradation of the performance of the receiver beyond a specified limit (Document I-ETS 300 113:1992).

2 Blocking level measurements

- Adjust the desired signal at the bit error ratio (BER) limit level.
- Increase this desired signal by 3 dB and add the interfering signal which is increased until the same BER is obtained.
- The ratio (interfering signal/desired signal) is the value of the receiver blocking.

FIGURE 10
Measurement procedure



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3 Attenuation of the receiver

During the measurement procedure, the three following equations are valid:

- Noise floor + Protection ratio + 3 dB = Desired signal level,
- Desired signal level + Blocking = Interfering signal level,
- Interfering signal level – Attenuation = Noise floor.

Hence:

$$\text{Attenuation} = 3 \text{ dB} + \text{Protection ratio} + \text{Blocking}$$

FIGURE 11

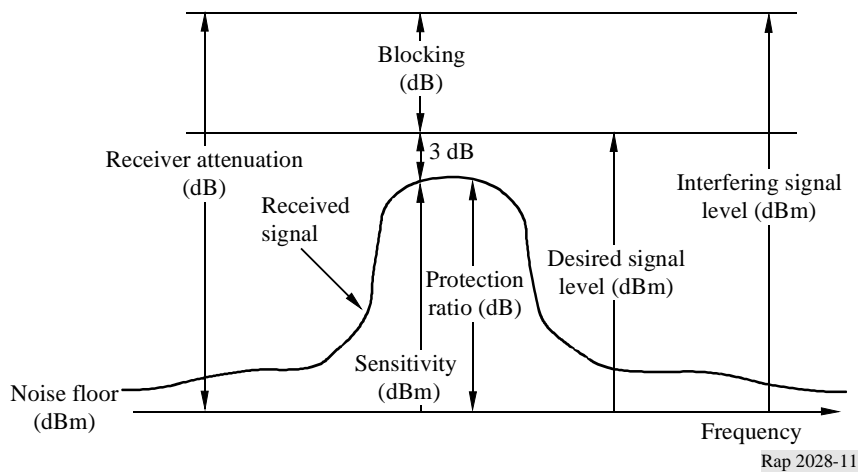
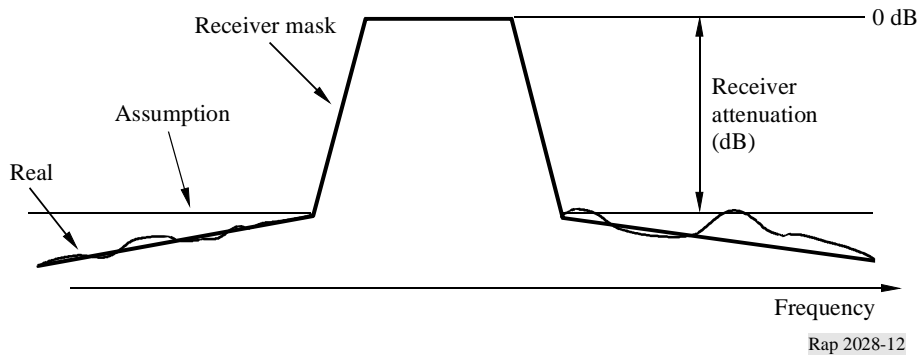


FIGURE 12

Receiver mask

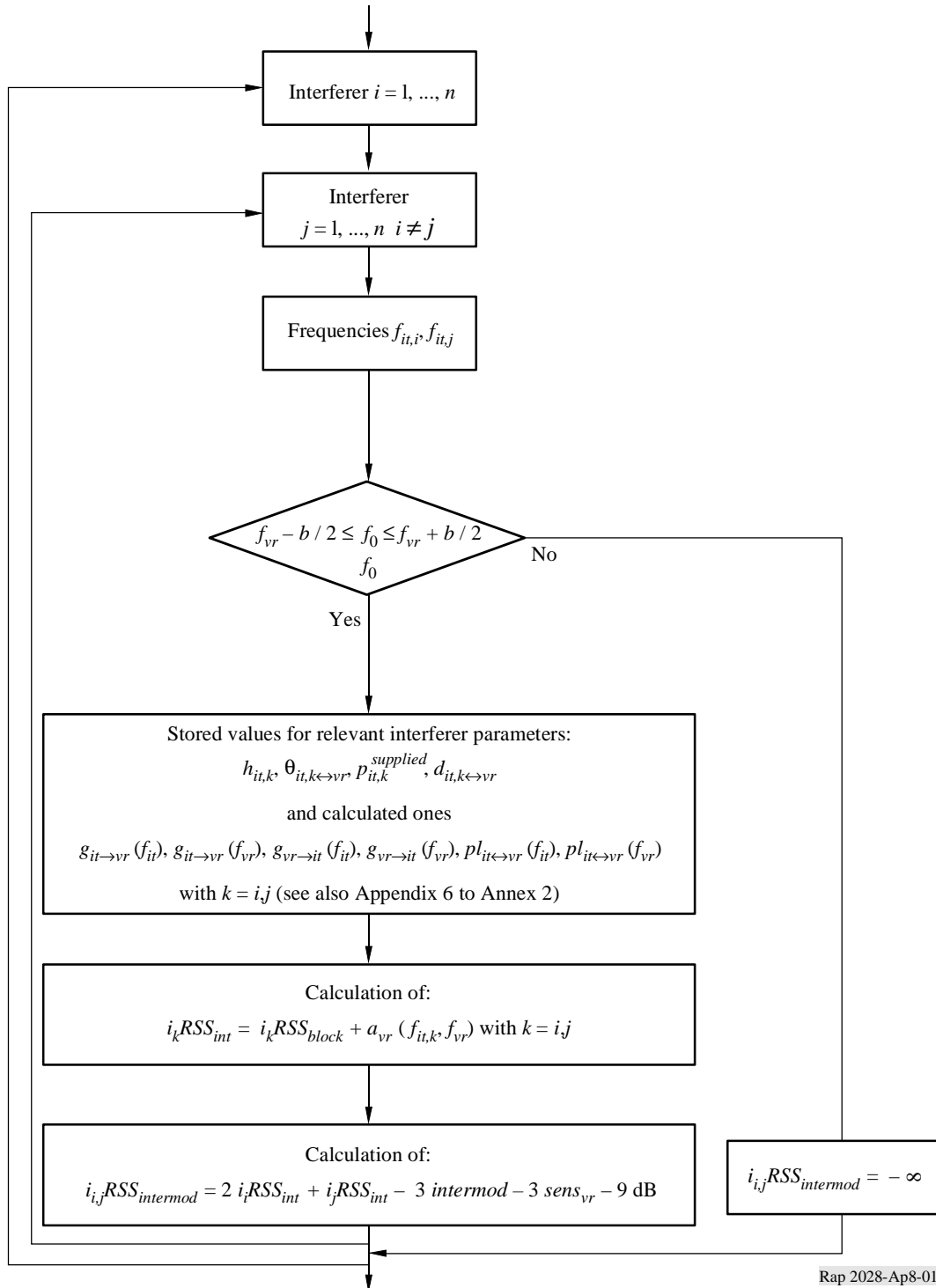


APPENDIX 8

TO ANNEX 2

***i*RSS due to intermodulation**

This flow chart is part of the flow chart given in Appendix 6.



APPENDIX 9

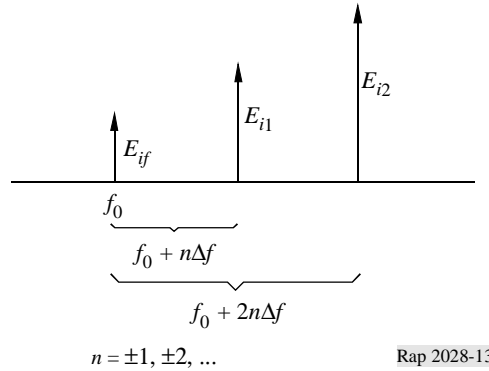
TO ANNEX 2

Intermodulation in the receiver

The main contribution to intermodulation interference originates from interfering signals in neighbouring channels due to the frequency selectivity of the antennas and the receiver equipment. We consider a service with a desired signal at frequency f_0 , a channel separation Δf and interfering signals E_{i1} and E_{i2} at frequencies $f_0 + n\Delta f$ and $f_0 + 2n\Delta f$, respectively. The receiver non-linearities produce an intermodulation product E_{if} of third order at the frequency (see Fig. 13).

$$f_0 = 2(f_0 + n\Delta f) - (f_0 + 2n\Delta f) \quad n = \pm 1, \pm 2, \dots \quad (1)$$

FIGURE 13



The signal strength E_{if} of the intermodulation product is given by:

$$E_{if} = kE_{i1}^2 E_{i2} \quad (2)$$

with some constant k to be determined. For signal levels (measured in dB) equation (2) reads

$$L_{if} = 2L_{i1} + L_{i2} + 20 \log k \quad (3)$$

The constant $20 \log k$ in equation (3) can be found from the measurement procedure which is described in the European Telecommunications Standards Institute (ETSI) Standard ETS 300-113, § 8.8. The method is similar to the contribution in Appendix 7 for blocking interference.

ETS 300-113 defines via the intermodulation response L_{imr} , the interfering signal levels $L_{i1} = L_{i2}$ at which bit errors due to intermodulation just start to be recorded (see Fig. 14).

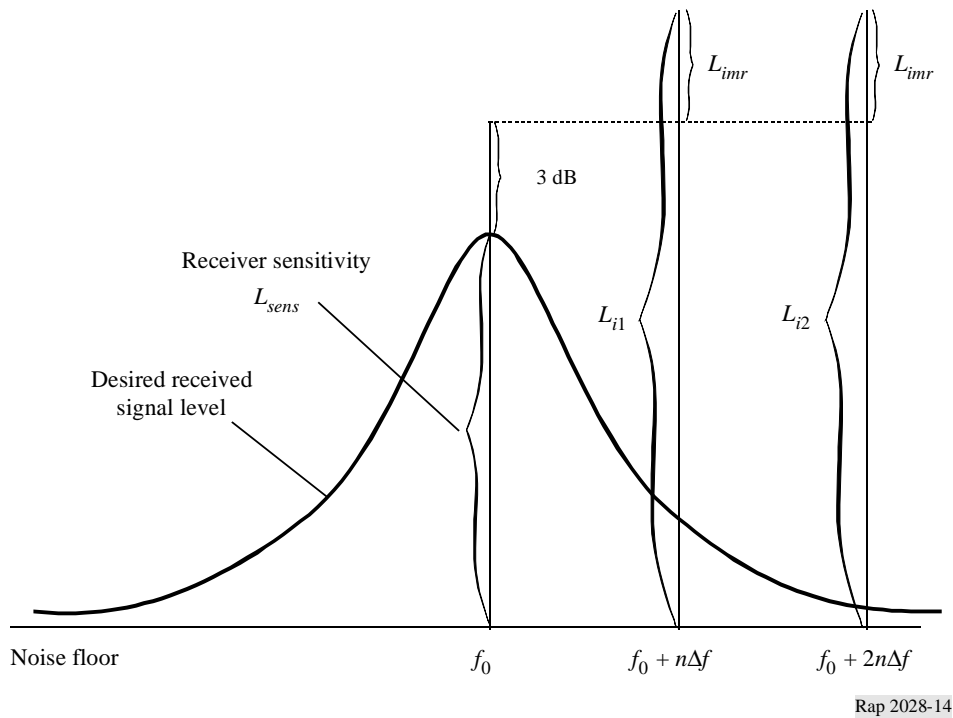
This means, for L_{i1} and L_{i2} as in Fig. 14, we have an intermodulation product L_{if} just at the noise floor (0 dB). Introducing L_{i1} and L_{i2} from Fig. 14 into equation (3) we obtain:

$$0 = 2(L_{imr} + 3 \text{ dB} + L_{sens}) + (L_{imr} + 3 \text{ dB} + L_{sens}) + 20 \log k \quad (4)$$

With the value of k from equation (4), equation (3) becomes:

$$L_{if} = 2L_{i1} + L_{i2} - 3L_{imr} - 3L_{sens} - 9 \quad \text{dB} \quad (5)$$

FIGURE 14



APPENDIX 10

TO ANNEX 2

Influence of different bandwidths**a) Wanted path**

The wanted transmitter transmits its power p_{wt} (dBm) at the frequency f_{vr} within a given bandwidth b_{vr} . This bandwidth is also used for the determination of the intermodulation products (see Appendix 8).

b) Interfering transmitter

For the interfering transmitter, an unwanted transmission mask p_{mi} as function of $\Delta f = f - f_{it}$ should be defined as maximum power levels $p_{mi}(\Delta f)$ in reference bandwidth $b_s(\Delta f)$ as specified by the user.

The interfering transmitter power p_{it} (dBm) at f_{it} is used for evaluating the link budget with the wanted receiver (i.e. power control).

If no mask is defined, the unwanted emission bandwidth b_{it} should be defined and the transmitted power should be assumed to spread uniformly over b_{it} with no emission outside b_{it} .

NOTE 1 – This approach provides a clear splitting between the case where no mask is needed (i.e. co-channel) and the case where a mask is given, which assumes that the mask is also specified within the transmitter bandwidth (no systematic assumption can be done here since we need to envisage non-flat spectrum). b_{it} is not used by the tool in the case of the user-defined mask.

c) Principle of determination of interfering power

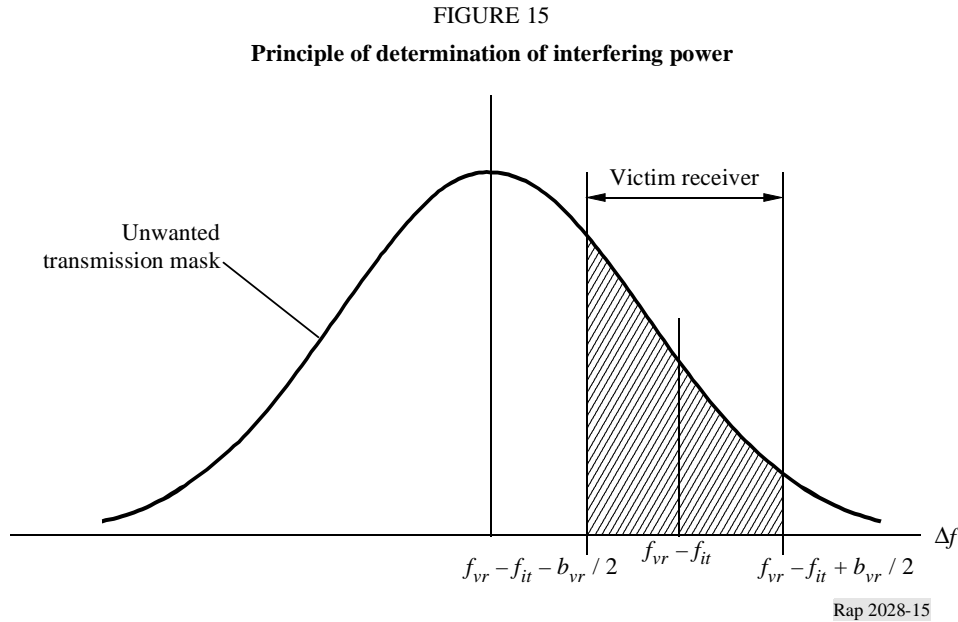


Figure 15 shows the principle of determination of the interfering power, $spur(f_{it}, f_{vr})$. If $f_{it} = f_{vr}$, then the interfering frequencies falls exactly in the receiving band of the victim receiver (co-channel interference).

For simplification within the algorithms, the mask function p_{mi} is normalized to a 1 Hz reference bandwidth:

$$p_{ni} = p_{mi}(\Delta f) - 10 \log \frac{b}{1 \text{ Hz}}$$

The bandwidth b is the bandwidth used for deriving p_{mi} . If no mask was provided, $b = b_{it}$ and $p_{mi} = p_{it}$ (assuming flat spectrum). If a mask has been specified, $b = b_s$ and $p_{mi} = spur(\Delta f)$.

The total received interfering power $spur_{tot}$ can easily be calculated by integration over the receiver bandwidth from $a = f_{vr} - f_{it} - b_{vr}/2$ to $b = f_{vr} - f_{it} + b_{vr}/2$

$$spur_{tot} = 10 \log \left\{ \int_a^b 10^{(p_{ni}(\Delta f)/10)} d\Delta f \right\}$$

with p_{ni} denoting the normalized mask (dBm/Hz). Using a 1 Hz reference bandwidth the integral can be replaced by a summation:

$$spur_{tot} = 10 \log \left\{ \sum_{i=a}^b 10^{(p_{ni}(\Delta f_i)/10)} \right\}$$

where $spur_{tot}$ is given in dBm.

NOTE 1 – The interfering power of a radio system having a different bandwidth can be estimated by the aforementioned algorithms. This calculation is only required for the interference due to unwanted emissions but not for blocking and intermodulation.

Note that it is recommended at all times that the mode of a user-defined mask be applied even if the mask is flat.

d) Implementation in SEAMCAT

User-defined mask

In c) the principle is explained. However, this algorithm is very slow in terms of computation time. Therefore the following approach is used:

The total interfering power relative to carrier, $spur_rel$, can be calculated by integration over the receiver bandwidth from $a = f_{vr} - f_{it} - b_{vr}/2$ to $b = f_{vr} - f_{it} + b_{vr}/2$

$$spur_rel = 10 \log \left\{ \int_a^b P_{rel}^{linear}(\Delta f) d\Delta f \right\} = 10 \log \left\{ \int_a^b 10^{\frac{P_{rel}^{dBc}(\Delta f)}{10}} d\Delta f \right\}$$

With P_{rel}^{dBc} denoting the normalized user-defined mask (dBc/Hz).

This mask is expressed as an array of $N + 1$ points $(\Delta f_i, P_i)$ and assumed linear between these points.

$$P_{rel}(\Delta f) = P_i + \frac{\Delta f - \Delta f_i}{\Delta f_{i+1} - \Delta f_i} (P_{i+1} - P_i)$$

This leads to:

$$spur_rel = 10 \log \left\{ \sum_{i=0}^{N-1} \int_{\Delta f_i}^{\Delta f_{i+1}} 10^{\frac{P_{rel}^{dBc}(\Delta f)}{10}} d\Delta f \right\}$$

where:

$$\Delta f_0 = a = f_{vr} - f_{it} - B_{vr}/2$$

$$\Delta f_N = b = f_{vr} - f_{it} + B_{vr}/2$$

Intermediate calculation:

$$\begin{aligned}
 spur_i^{dBc} &= \int_{\Delta f_i}^{\Delta f_{i+1}} 10^{\frac{P_{rel}^{dBc}(\Delta f)}{10}} d\Delta f \\
 spur_i^{dBc} &= 10^{\frac{P_i}{10}} \int_{\Delta f_i}^{\Delta f_{i+1}} \left[10^{\frac{P_{i+1}-P_i}{10(\Delta f_{i+1}-\Delta f_i)}} \right]^{(\Delta f-\Delta f_i)} d\Delta f \\
 spur_i^{dBc} &= \frac{10^{\frac{P_i}{10}}}{K^{\Delta f_i}} \int_{\Delta f_i}^{\Delta f_{i+1}} K^{(\Delta f-\Delta f_i)} d\Delta f, \quad K = 10^{\frac{P_{i+1}-P_i}{10(\Delta f_{i+1}-\Delta f_i)}} \\
 spur_i^{dBc} &= \frac{10^{\frac{P_i}{10}}}{K^{\Delta f_i}} \left[e^{\ln K} \right]_{\Delta f_i}^{\Delta f_{i+1}} = \frac{10^{\frac{P_i}{10}}}{\ln K} \left[K^{\Delta f_{i+1}-\Delta f_i} - 1 \right], \quad \ln K = \frac{\ln 10}{10} \cdot \frac{P_{i+1}-P_i}{\Delta f_{i+1}-\Delta f_i} \\
 spur_i^{dBc} &= \frac{10}{\ln 10} \frac{10^{P_{i+1}} - 10^{P_i}}{P_{i+1} - P_i} (\Delta f_{i+1} - \Delta f_i)
 \end{aligned}$$

Eventually:

$$spur_{rel} = 10 \log \left\{ \frac{10}{\ln 10} \sum_{i=0}^{N-1} \frac{(P_{i+1}^{linear} - P_i^{linear}) (\Delta f_{i+1} - \Delta f_i)}{(P_{i+1}^{dBc} - P_i^{dBc})} \right\}$$

To get the total unwanted emission, one has then to add the nominal power of the interfering transmitter:

$$spur_{tot} = spur_{rel} + P_{it}$$

Unwanted emission floor

The aforementioned formulas are also applicable to the unwanted emission floor, $spur_{floor}$ (dBm), except that in the end no power is added. If the unwanted emission floor is selected then the mechanism is the same whether one of the two masks is constant or not.

$$P_{it}(\text{dBm}) + spur_{rel}(\text{dBc}) + g_{PC} > spur_{floor}(\text{dBm})$$

Note that the comparison involves the power control gain if power control is selected.

Note that the unwanted emission floor is referred to 1 MHz in SEAMCAT.

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ANNEX 3

Distribution evaluation engine

The flow chart for the DEE is shown in Fig. 16. A fit-of-goodness test can be performed either by the chi-squared test or by the Kolmogorov-Smirnov algorithm (used in SEAMCAT).

This algorithm basically tests if a random sample of observations conform to a pre-specified cumulative distribution. The pre-defined distribution can be continuous, discrete or hybrid. Thus, the chi-squared method is very versatile and a single algorithm is proposed for use within DEE for testing all possible types of probability distribution functions.

An array of samples on RSS random variable is passed to the DEE. Firstly the DEE tests if the array length, N (number of samples), is long enough to produce a stable distribution. This is accomplished by using $N-dN$ samples to establish an initial discrete distribution function and calculate the corresponding cdf. This cdf is then used as a reference in the chi-squared test performed now on the complete population of N samples. Should the test show that two discrete distributions differ more than an acceptable and pre-specified value, a message is sent back to the EGE to generate some extra samples. On the contrary, if the chi-squared criteria is satisfied the DEE proceeds with testing whether or not a continuous probability density function can be used.

The flow-chart in Fig. 16 is an example of a Gaussian distribution test. The chi-squared algorithm is equally applicable to any other continuous distribution that might be representative of RSS random variable. A continuous distribution function enables a closed form expression for probability calculation in ICE, this in turn warrants a numerically efficient calculation. If no continuous pdf fits the sample population with the adequate accuracy, discrete pdf representation and a numerical probability calculation is the only way forward.

Notation used:

$\langle RSS \rangle$: random variable population

N : sample population size

I : internal counter to give stability testing

dN : portion of population size (e.g. $dB = 0.1N$)

Y : chi-squared test criteria (see Appendix 1 to Annex 3)

$\chi_{1-\alpha}$: quantile – reference level for chi-squared test

n : total counter sample

$\langle C \rangle$: discrete cdf coefficient array

The flow chart in Fig. 17 presents one of many different possibilities to form the discrete pdf for a random variable.

FIGURE 16
DEE flow chart

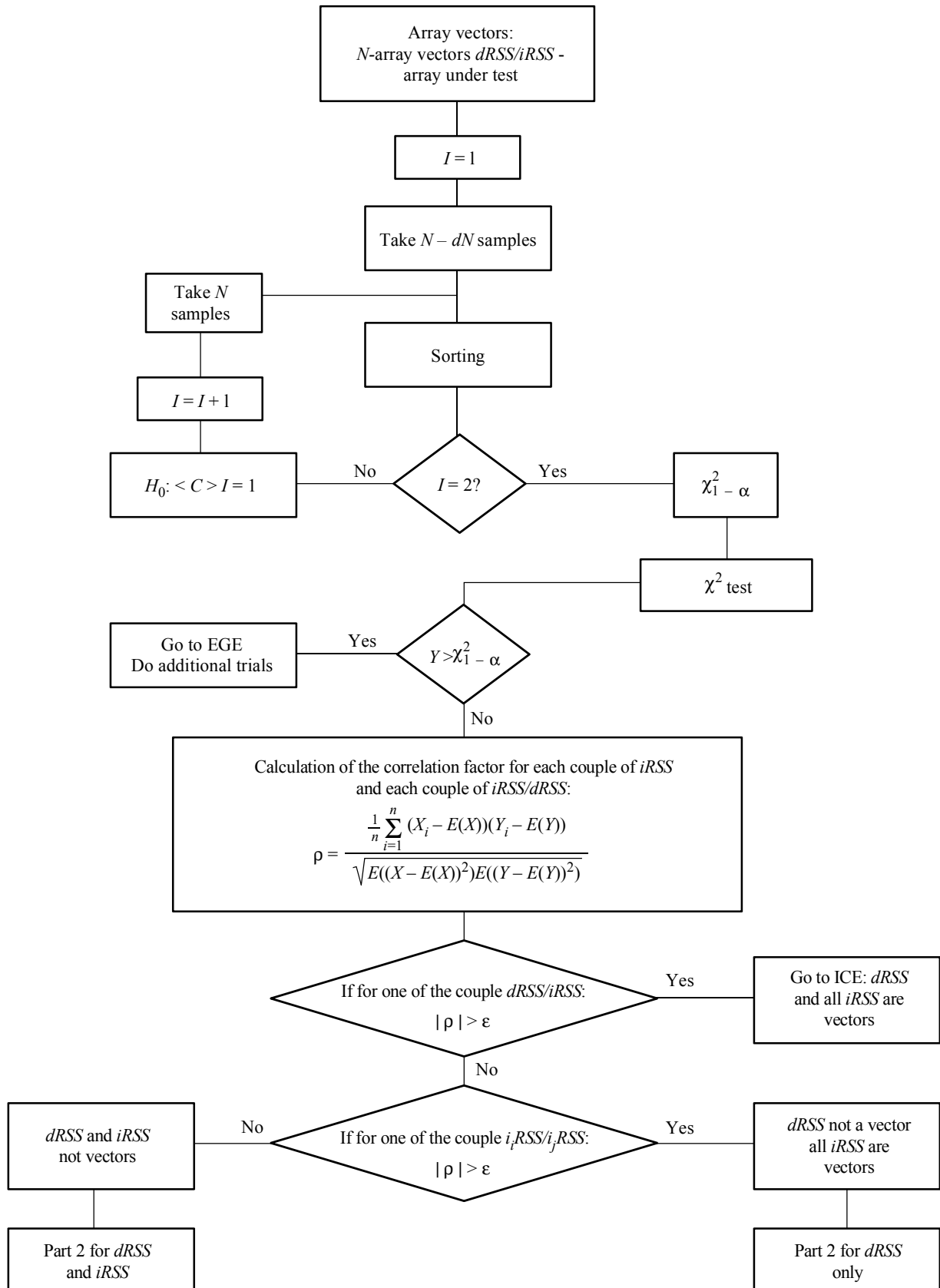
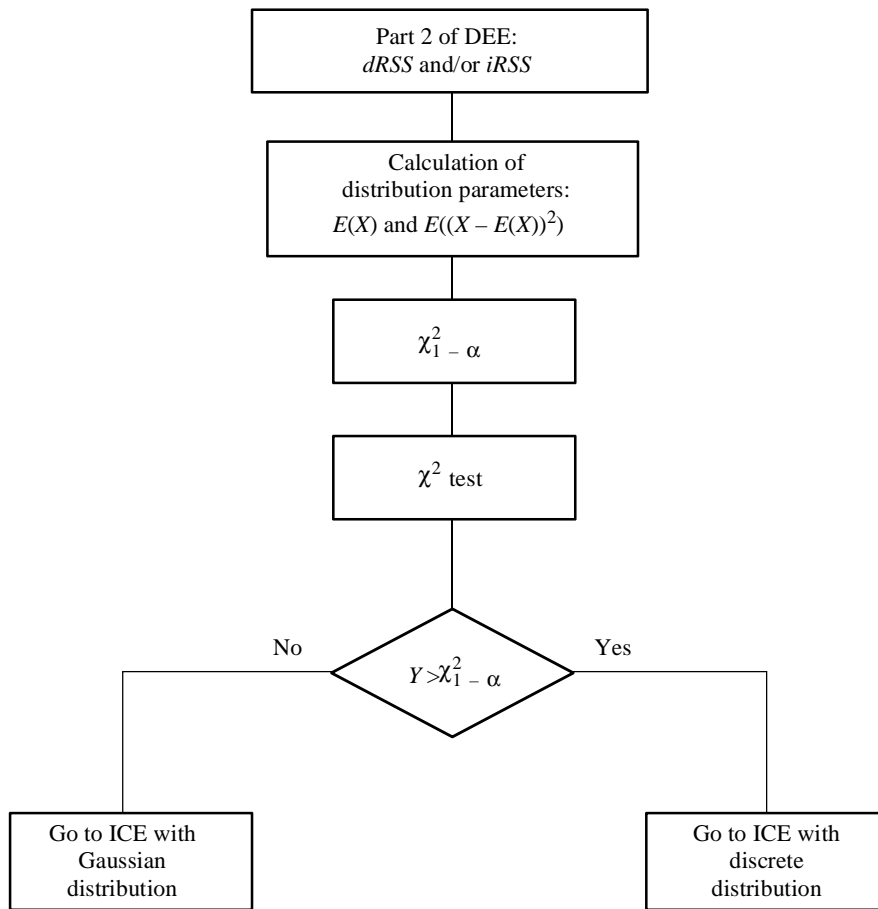


FIGURE 17



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APPENDIX 1

TO ANNEX 3

Chi-squared goodness-of-fit test

The chi-squared goodness-of-fit test is one the oldest and best known statistical tests.

Lets assume X_1, X_2, \dots, X_N be a sample set drawn from a population with unknown cdf, $F_x(x)$. The chi-squared test is based on testing the null hypothesis:

$$H_0: F_x(x) = F_0(x) \text{ for all } x \quad \text{against the alternative } H_1: F_x(x) \neq F_0(x) \text{ for some } x$$

Assume that N observations are grouped into K mutually exclusive categories. Let's denote by N_j the observed number of trials in j -th category ($j = 1, 2, \dots, K$). In addition, denote N_j^0 the number of trials expected to fall into j -th category according to the known cdf, $F_0(x)$.

The actual test employs the following criteria:

$$Y = \sum_{j=1}^K \frac{(N_j - N_j^0)^2}{N_j^0}, \quad \sum_{j=1}^K N_j = N$$

which tends to be small when H_0 is true and large when H_0 is false. The Y is also the random variable which obeys chi-square distribution for large N .

In practice, for the hypothesis H_0 to prevail we expect:

$$P(Y > \chi_{1-\alpha}^2) = \alpha$$

where α is the significant level, say 0.05 or 0.1; the quantile $\chi_{1-\alpha}^2$ corresponds to probability of $1-\alpha$ is given in the Tables for chi-squared distribution (see Table 2).

The chi-squared goodness-of-fit test is equally applicable to discrete and continuous probability density functions.

TABLE 2

Quantile $\chi_{1-\alpha}^2$ for chi-squared distribution

$K \backslash 1-\alpha$	0.975	0.95	0.90	0.75
10	3.25	3.94	4.86	6.74
20	9.59	10.85	12.44	15.45
30	16.79	18.49	20.60	24.48
40	24.43	28.51	29.05	33.66
50	32.36	34.76	37.69	42.94
60	40.48	43.19	46.46	52.29
70	48.76	51.74	55.33	61.70
80	57.15	60.39	64.28	71.14
90	65.65	69.13	73.29	80.62
100	74.22	77.93	82.36	90.13

APPENDIX 2

TO ANNEX 3

Kolmogorov-Smirnov test of stability

Purpose of this evaluation stage is to estimate whether the number of generated events is enough to consider the results as stable from a statistical point of view. The stability evaluation is performed by a goodness-of-fit test with the Kolmogorov-Smirnov test in order to check if the distribution obtained with $N - dN$ samples and the one obtained with N samples do not differ by more than a specified value:

First, two cumulative distribution functions have to be derived from the input array vector:

- distribution derived from the first $N - dN$ samples of the array vector,
- distribution derived from the complete array vector (N samples).

This is done by means of a simple array sort. The test then simply consists in performing the chi-squared test with following input:

- specified stability threshold (between 0 and 1),
- reference distribution: distribution derived from the N -array,
- tested distribution: distribution derived from the $N - dN$ array.

According to the result of the Kolmogorov-Smirnov test, if the result is greater than the stability threshold, stability evaluation is considered successful.

ANNEX 4

Interference calculation engine

The ICE has two different functions:

- Process different interfering signals in order to calculate the probability for interference. Three types of interfering signals are considered: spurious emission, out-of-band emission, and blocking and intermodulation.
- Derive generic limits. The output of the ICE is then a multidimensional surface giving the probability of interference versus radio parameters. The general ICE flow chart is shown in Fig. 18.

The interfering signal distributions are calculated with respect to reference levels or functions of unwanted (emission mask), blocking (receiver mask) or intermodulation attenuation. The translation law for the cdf from reference ref_{i-init} to reference ref_i is given by the following formula:

$$P(iRSS_i(ref_i) < X) = P(iRSS_i(ref_{i-init}) < X - t(ref_i - ref_{i-init})); \quad t = \begin{cases} 1; & i = spur \\ -1; & i = block \\ -3; & i = intermod \end{cases} \quad (6)$$

The complete and quick (approximate) flow charts for the ICE are shown in Figs. 19 and 20 respectively. For sake of simplicity, the case of $t = 1$ (equation (6), spurious case) appears in flow charts of Figs. 19 and 20.

Quick calculation algorithm

In the ICE quick calculation algorithm we make the following two assumptions:

- The i_iRSS are independent variables, where the index i corresponds to the i -th type of interfering scenario.
- One of the i_iRSS is dominant with respect to all the other interfering signals.

The overall probability P_D for not being interfered by the composite interfering signal reads:

$$P_D = P \left(\frac{dRSS}{iRSS_{composite}} > \frac{C}{I} \mid dRSS > sens_{vr} \right) \quad (7)$$

Using the second assumption, we can approximate equation (7) by the following equation:

$$P_D = P \left(\bigcap_{i=1}^n \left(\frac{dRSS}{i_iRSS} > \frac{C}{I} \mid dRSS > sens \right) \right) \quad (8)$$

and since the i_iRSS are independent variables, we can write equation (8) as:

$$P_D \approx \prod_{i=1}^n P \left(\frac{dRSS}{i_iRSS} > \frac{C}{I} \mid dRSS > sens \right) \equiv \prod_{i=1}^n P_i(C/I) \quad (9)$$

For each interfering scenario corresponds a set of references, ref_i , e.g. *spur*, a_{vr} , etc. The user can choose the set of references that will be used in the calculation of P_D . We incorporate ref_i in equation (9) and get the following approximation:

$$P_D \approx \prod_{i=1}^n P_i(C/I, ref_i) \quad (10)$$

which is used in the quick calculation algorithm. It can be easily shown that $1 - P_D$ gives the probability of being disturbed by at least one of the n interferers.

Complete ICE flow chart

Three cases are considered:

- The desired and/or the interfering signals are correlated. In this case the probability P_D is calculated by processing directly the data vectors. For each interfering scenario, the interfering signals of all interferers are summed up to get $iRSS_{composite}$. Then, from the two vectors $iRSS$ and $iRSS_{composite}$ we calculate the probability P_D :

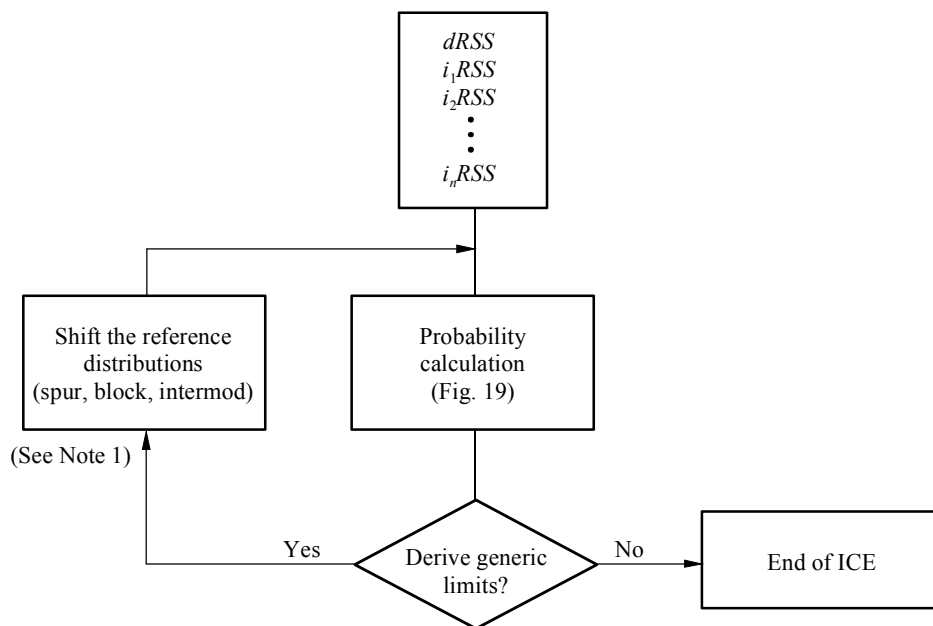
$$P_D = P \left(\frac{dRSS}{iRSS_{composite}} > \frac{C}{I} \mid dRSS > sens \right) \quad (11)$$

by summing up all the terms satisfying $dRSS > sens$. Similarly to the quick calculation case, when we sum up elements from the data vectors to calculate equation (10), we should update the data so that it corresponds to a desired set of references.

- All signals are uncorrelated and their distributions (calculated by the DEE) are given in closed form. First, the cumulative distribution function of the composite interfering signal is calculated by integrating the i_iRSS distribution functions. Note that the ref_i cause linear shifts of the i_iRSS distributions with respect to one another. In the calculation of the $i_iRSS_{composite}$ composite, the i_iRSS distributions should be shifted so that they all refer to the same set of references. Finally, equation (10) is calculated by using the conditional probability formula which integrates the distributions $dRSS$ and $iRSS_{composite}$.
- The third case is similar to the second one, with the exception that the $iRSS_{composite}$ distribution function is determined by the Monte Carlo technique.

FIGURE 18

General ICE flow chart

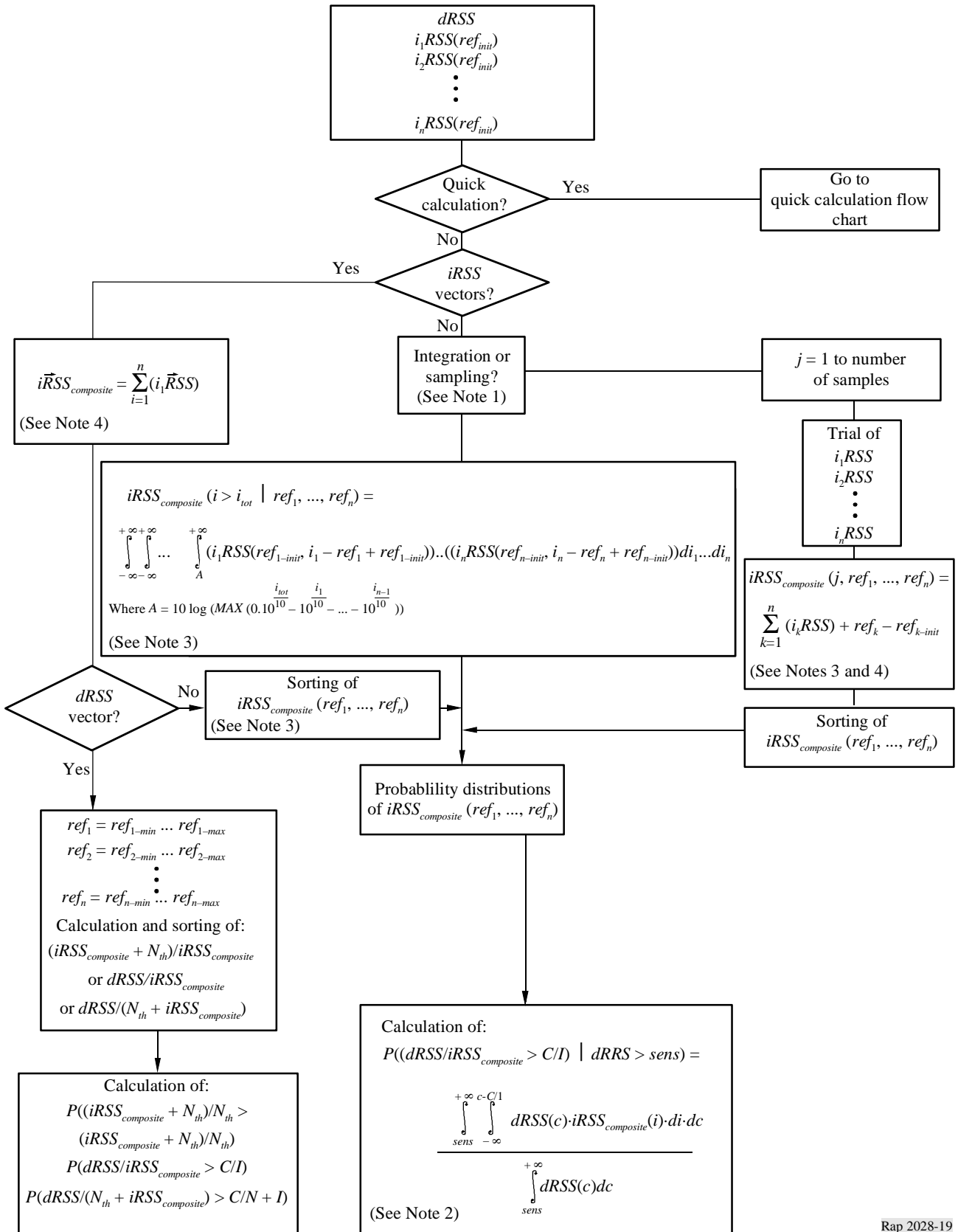


Note 1 – This loop is repeated for each value of spur, block, and intermod in order to get an N -dimensional curve.

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The flow chart in Fig. 18 is describing the logical process of ICE, which is well suited in the case of the full integration for the calculation of $iRSS_{composite}$ (see flow chart in Fig. 19). However, in the case of input vector data or Monte Carlo sampling process, the calculation of the summation of vectors for determining $iRSS_{composite}$ and the trials of i_iRSS , respectively, which are time and resource consuming, can be made only once as shown in Fig. 19.

FIGURE 19
ICE flow chart



Notes relative to Fig. 19:

Note 1 – Computing time is the criteria to choose between sampling or integrating.

Note 2 – This formula is detailed in Document SE21(96)/20(Add.1). ($dRSS/I$) is the criteria used in this example. Other criteria may be used.

Note 3 – ref_1, \dots, ref_n are the values of the relevant parameters (spur, a_{av} , ...) for which the calculation of the probability of interference is needed.

Note 4 – The meaning of this sum is symbolic since the addition is to be made on the linear values and that i_iRSS is expressed in dB.

FIGURE 20
ICE quick calculation flow chart

