TECHNICAL REPORT

CISPR 16-3

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AMENDMENT 2
2006-11

INTERNATIONAL SPECIAL COMMITTEE ON RADIO INTERFERENCE

Amendment 2

Specification for radio disturbance and immunity measuring apparatus and methods –

Part 3: CISPR technical reports

International Electrotechnical Commission, 3, rue de Varembé, PO Box 131, CH-1211 Geneva 20, Switzerland
Telephone: +41 22 919 02 11 Telefax: +41 22 919 03 00 E-mail: inmail@iec.ch Web: www.iec.ch

Commission Electrotechnique Internationale
International Electrotechnical Commission
Международная Электротехническая Комиссия

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FOREWORD

This amendment has been prepared by CISPR subcommittee A: Radio interference measurements and statistical methods.

The text of this amendment is based on the following documents:

DTR	Report on voting
CISPR/A/659/DTR	CISPR/A/681/RVC
CISPR/A/662/DTR	CISPR/A/678/RVC

Full information on the voting for the approval of this amendment can be found in the report on voting indicated in the above table.

The committee has decided that the contents of this amendment and the base publication will remain unchanged until the maintenance result date indicated on the IEC web site under "http://webstore.iec.ch" in the data related to the specific publication. At this date, the publication will be

- reconfirmed,
- withdrawn,
- replaced by a revised edition, or
- amended.

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3 Definitions

Add, on page 9, after 3.11, the following new definitions:

3.12
weighting (e.g. of impulsive disturbance)
the pulse-repetition-frequency (PRF) dependent conversion (mostly reduction) of a peak-detected impulse voltage level to an indication which corresponds to the interference effect on radio reception

NOTE 1 For the analog receiver, the interference effect is the psychophysical annoyance, i.e. a subjective quantity (audible or visual, usually not a certain number of misunderstandings of a spoken text).

For the digital receiver, the interference effect may be defined by the critical Bit Error Ratio (BER) (or Bit Error Probability (BEP)), for which perfect error correction can still occur, or by another objective and reproducible parameter.

3.13
weighting characteristic
the peak voltage level as a function of PRF for a constant effect on a specific radio-communication system, i.e., the disturbance is weighted by the radio communication system itself

3.14
weighting function
weighting curve
the relationship between input peak voltage level and PRF for constant level indication of a measuring receiver with a weighting detector, i.e. the curve of response of a measuring receiver to repeated pulses

3.15
weighting factor
the value in dB of the weighting function relative to a reference PRF or relative to the peak value

3.16
weighting detector
detector which provides an agreed weighting function

3.17
weighted disturbance measurement
measurement of disturbance using a weighting detector

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4 Technical Reports

Add, after the existing subclause 4.7 published in Amendment 1, the following new subclauses 4.8 and 4.9:

4.8 Background material on the definition of the r.m.s.-average weighting detector for measuring receivers

4.8.1 Introduction – purpose of weighted measurement of disturbance

Generally, a weighted measurement of impulsive disturbance serves the purpose of minimizing the cost of disturbance suppression, while keeping an agreed level of radio protection. The weighting of a disturbance for its effect on modern digital radiocommunication services is important for the definition of emission limits that will protect these services. Amendment 1 of CISPR 16-1-1 defines a detector that is a combination of an r.m.s. and an average detector. The selection of the type of detector and of the transition between these detector functions is based on measurements and theoretical investigations.

4.8.2 General principle of weighting – the CISPR quasi-peak detector

The effect on radiocommunication services depends on the type of interference (e.g. broadband or narrowband, pulse rate etc.) and on the type of service itself. The effect of the pulse rate was recognized a short time after the CISPR was founded in 1933. As a result, the quasi-peak weighting receiver for the frequency range of 150 kHz to 1 605 kHz was defined as shown for band B in Figure 4.8.1. However in CISPR 1 [1] it was already accepted that “Subsequent experience has shown that the r.m.s. voltmeter might give a more accurate assessment” but the quasi-peak type of voltmeter has been retained for certain reasons – mainly for continuity.

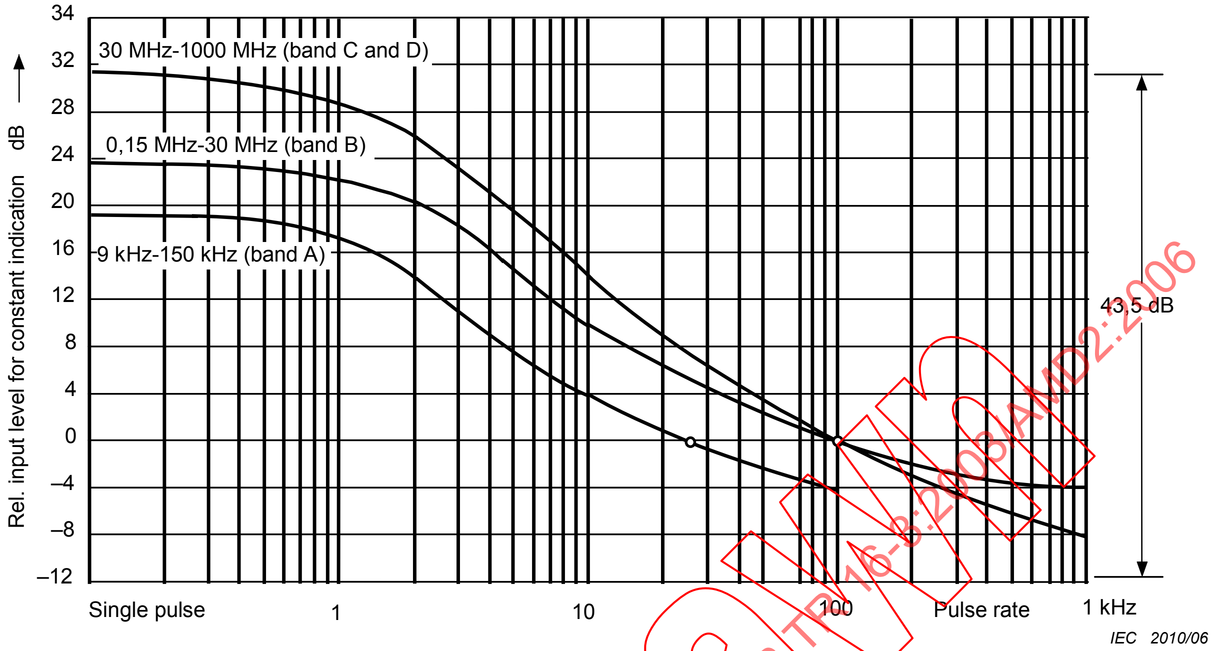


Figure 4.8.1 – Weighting curves of quasi-peak measuring receivers for the different frequency ranges as defined in CISPR 16-1-1. The weighting factor is shown relative to a reference pulse rate (25 Hz or 100 Hz)

4.8.3 Other detectors defined in CISPR 16-1-1

- **Peak detector**

The peak detector follows the signal at the output of the IF envelope detector and holds the maximum value during the measurement time (also called dwell time) until its discharge is forced. This indication is independent of the pulse repetition frequency (PRF).

- **Average detector**

The average detector determines the linear average of the signal at the output of the IF envelope detector. It should be kept in mind that for low PRFs, CISPR 16-1-1 specifies the average detector measurement result as the maximum scale deflection of a meter with a time constant specified for the quasi-peak detector. This is necessary to avoid reduced level indication for a pulse modulated disturbance by using long measurement times. The weighting function varies with 20 dB per decade of the PRF (see Figure 4.8.2).

- **RMS detector**

The r.m.s. detector determines the r.m.s. value of the signal at the output of the IF envelope detector. Despite being mentioned in [1] and being described in CISPR 16-1-1, at the time of writing of this report it has not been put to practical use in CISPR product standards. The weighting function varies with 10 dB per decade of the PRF (see Figure 4.8.2). Up to now, no meter time constant applies for the r.m.s. detector for intermittent, unsteady and drifting narrowband disturbances.

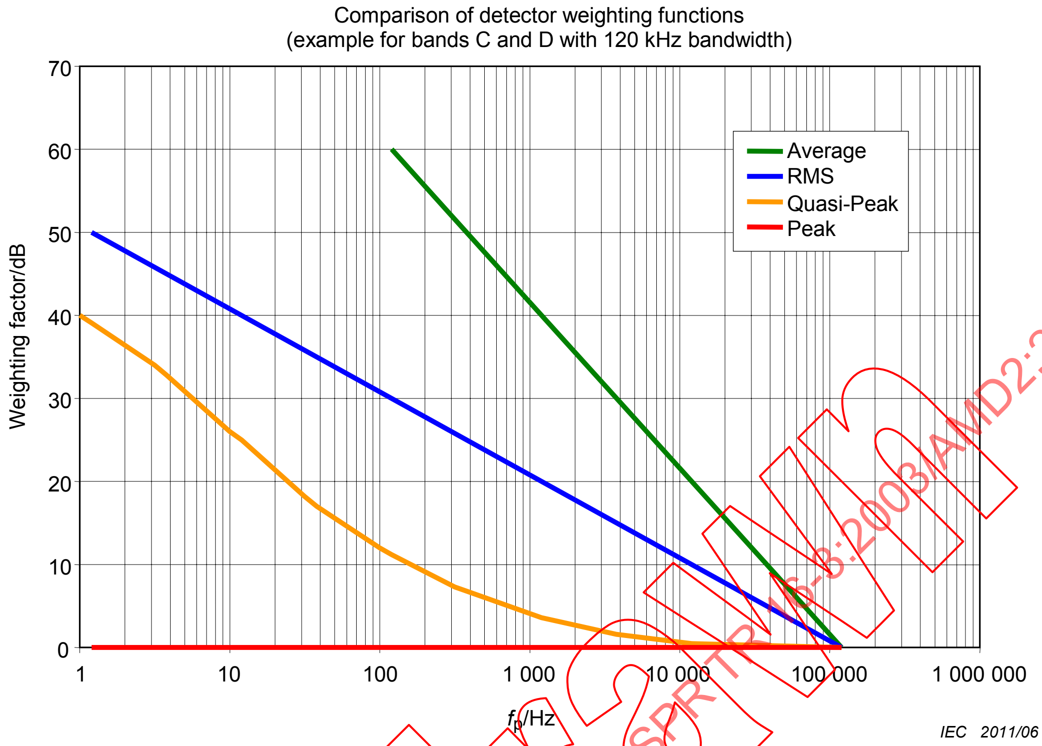


Figure 4.8.2 – Weighting curves for peak, quasi-peak, r.m.s. and linear average detectors for CISPR bands C and D

4.8.4 Procedures for measuring pulse weighting characteristics of digital radiocommunications services

All modern radio services use digital modulation schemes. This is not only true for mobile radio but also for audio and TV. Procedures for data compression and processing of analog signals (voice and picture) are used together with data redundancy for error correction. Usually, up to a certain critical bit-error ratio (BER) the system can correct errors so that perfect reception occurs.

Whereas analog radio systems require signal-to-noise ratios of as much as 50 dB for satisfactory operation, in general, digital radio communication systems allow error-free operation down to signal-to-noise ratios of approximately 10 dB. However the transition region from error-free operation to malfunction is small. Therefore planning guidelines for digital radio are based on almost 100 % coverage. When a digital radio receiver operates at low input levels, the susceptibility to radio disturbance is important. In mobile radio reception, the susceptibility to radio disturbance is combined with the problem of multi-path propagation.

4.8.4.1 Principles of measurement

The significance of the weighting curve for band B in Figure 4.8.1 is as follows: to a listener the degradation of reception quality, caused by a 100-Hz pulse, is equivalent to the degradation from a 10-Hz pulse, if the pulse level is increased by an amount of 10 dB. In analogy to the above, an interference source with certain characteristics will produce a certain BER, e.g. 10^{-3} in a digital radiocommunication system, when the interfering signal is received in addition to the radio signal. The BER will depend e.g. on the pulse repetition frequency (PRF) and the level of the interfering signal. In order to keep the BER constant, the level of the interfering signal will have to be readjusted while the PRF is varied. This level variation vs. PRF determines the weighting characteristics. Measurement systems with BER indication are needed to determine the required level of the interfering signal for a constant BER as e.g. shown in Figure 4.8.3.

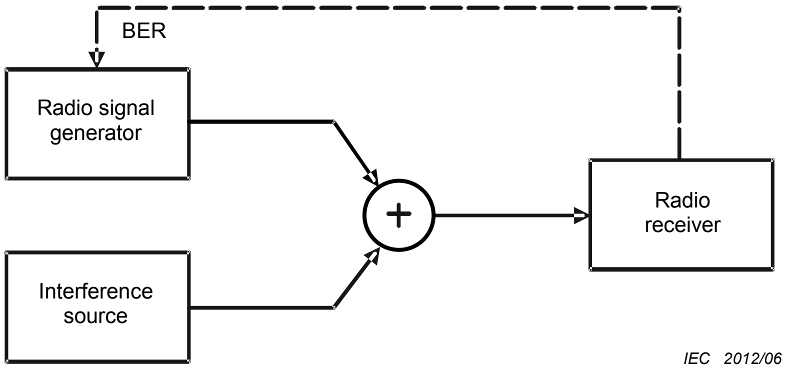


Figure 4.8.3 – Test setup for the measurement of the pulse weighting characteristics of a digital radiocommunication system

The test setup shown in Figure 4.8.3 consists of a radio signal generator that transmits the wanted radio signal to the receiver. For the determination of the BER, the radio receiver either has to know the original bit sequence for comparison with the detected bit sequence or the latter must be looped back to the radio signal generator for comparison with the original. Both systems are available and have been used for tests. Mobile radio testers, e.g., apply the loop-back principle.

4.8.4.2 Generation of the interference signal

A signal generator with pulse-modulation capability can be used to generate the interference signal. For correct measurements, the pulse modulator requires a high ON/OFF ratio of more than 60 dB. Using the appropriate pulse width, the interference spectrum can be broadband or narrowband, where the definition of broadband and narrowband is relative to the communication channel bandwidth. Figure 4.8.4 gives an example of an interference spectrum used for the determination of weighting characteristics.

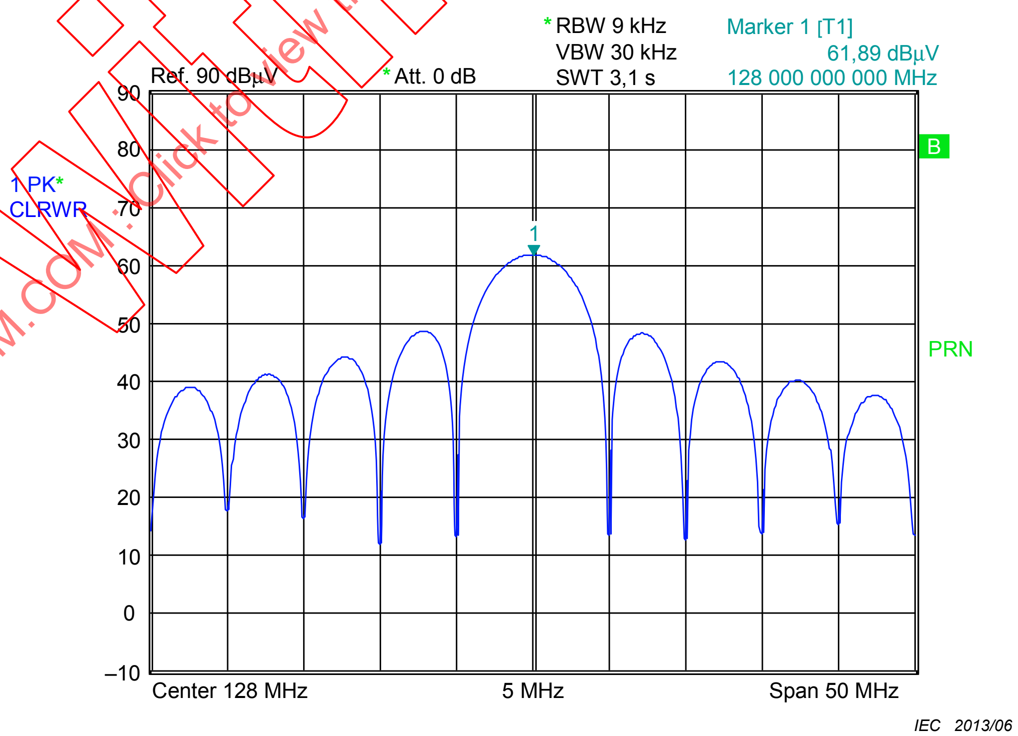


Figure 4.8.4 – Example of an interference spectrum: pulse modulated carrier with a pulse duration of 0,2 µs and a PRF < 10 kHz

With increasing pulse duration, the main lobe of the spectrum becomes narrower. This is also used to study the effect of narrowband pulses on radiocommunication systems. The advantage of using a band-limited pulse spectrum instead of a broadband pulse generator is to avoid overloading the receiver under test. Otherwise non-linearity effects could cause deterioration of the weighting characteristics. In addition to pulse-modulated carriers, unmodulated carriers can be used to determine the sensitivity of different systems to narrowband (CW signal) EMI.

Extensive measurements have also been presented in [2] with on/off-keying of a QPSK-modulated signal, thus keeping the spectrum width wider than the system bandwidth even with longer pulse durations. Since actual receivers do not provide BER indication, the method described in the ITU Recommendation 1368 was used as the failure criteria: DVB-T reception was regarded as distorted when more than one visible erroneous block was shown on the screen within an observation period of 20 s. Alternatively, any picture-freeze, also for short periods, was regarded as a failure. For DRM, the reception was considered as distorted when the system showed more than one dropout in a 20 s observation time.

Further measurements have been made with spread-spectrum modulated carriers in order to study the effect of spread-spectrum clock interference on wideband radiocommunication services (see [3] and [4]).

Table 4.8.1 – Overview of types of interference used in the experimental study of weighting characteristics

Interference signals	Pulse-modulated	On/Off-keyed QPSK-modulated	Spread-spectrum modulated
Pulse width in relation to signal bandwidth	T < 1/B to 100/B	T < 1/B to 100/B	Continuous
T = pulse width, B = radio signal bandwidth			

4.8.4.3 Other principles of measurement

The receiver under test should receive a signal that is just sufficient to give quasi error-free reception (e.g. a BER = 10^{-7} or a factor of 10^{-3} lower than the critical BER). Thus the receiver operates like a receiver at the rim of a coverage area, where a disturbance above the emission limit can easily cause interference.

For radio telephone systems, where the downlink (to the mobile) and uplink (to the base station) frequencies are in different bands, the use of a pulse modulated carrier helps to concentrate the interference on the mobile receiver and thus avoids interference with the loop-back connection.

4.8.5 Theoretical studies

The work of developing measurement procedures considering a digital radio receiver as a disturbance victim, is a very complex problem since there are many different modulation and coding schemes to consider as digital communication services are undergoing rapid development. The results of theoretical studies for radio systems using error correction have been presented in [5] and [6]. These studies are based on the same fundamental assumptions that are explained above:

- the BER is the performance parameter of interest for the digital communication system;
- the repetitive pulsed disturbance is the waveform of particular interest;
- the disturbance pulses have a pulse duration that is short compared to the digital symbols transmitted.

Results for some selected convolutional codes (for more details, see [5]):

A convolutional code is generated by passing the information sequence through a linear finite-state shift register. In general, the shift register consists of *K* stages and *n* algebraic function generators. The input data to the channel encoder is shifted into and along the shift register *k* bits at a time. The number of output bits for each *k*-input sequence is *n* bits. The rate *R* of the code is defined as *n/k*. The parameter *K* is called the constraint length of the convolutional code. In Figures 4.8.5 a) and b) as well as 4.8.6 a) and b) the r.m.s. and peak values corresponding to a constant BER of 10^{-3} are shown for different convolutional codes and *binary phase shift keying* (BPSK) modulation. These results have been simulated with ACOLADE© [1] (Advanced Communication Link Analysis and Design Environment). In the graphs, the pulse repetition frequency of the disturbance is presented as related (normalized) to the gross-bit rate (or symbol rate) R_s of the communication system. The simulation is done in the band-pass domain. This means that the results can be transformed to an arbitrary carrier frequency. The disturbance pulse width is 10 % of the bit duration time. For the lowest rate *R* = ¼, the r.m.s. value is approximately constant down to the critical point where it increases rapidly. Thus, for a well-protected system, the r.m.s. value corresponding to a constant BER is constant with respect to the pulse repetition frequency of the repetitive pulsed disturbance.

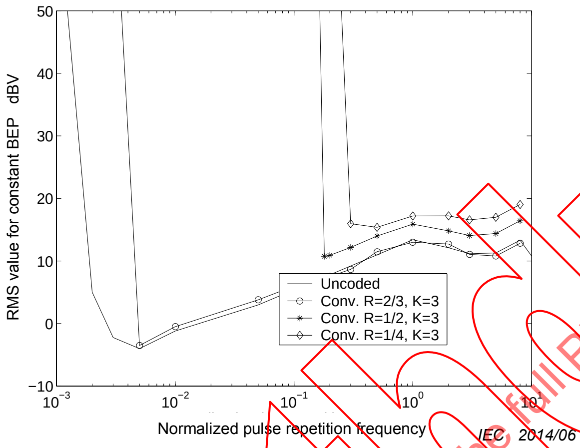


Figure 4.8.5 a) – The r.m.s. level for constant BEP for three *K*=3, convolutional codes of different rate

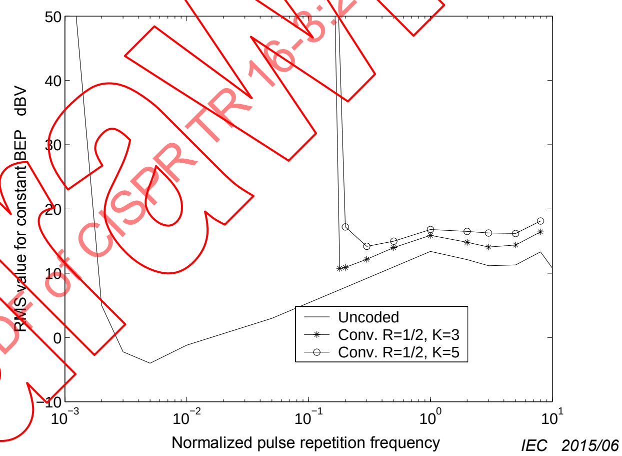


Figure 4.8.6 a) –The r.m.s. level for constant BEP for two rate ½, convolutional codes

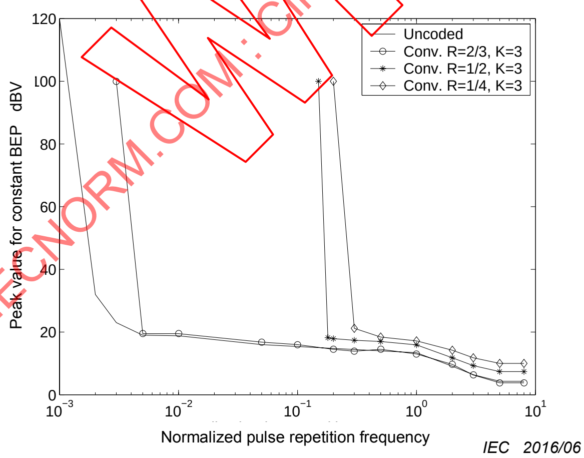


Figure 4.8.5 b) – The peak level for constant BEP for three *K*=3, convolutional codes of different rate

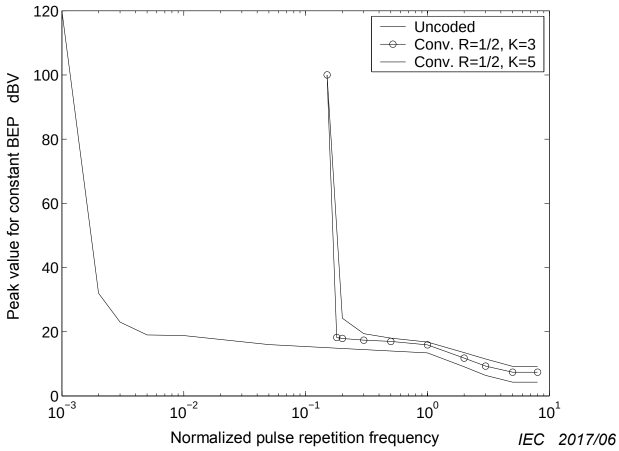


Figure 4.8.6 b) –The peak level for constant BEP for two rate ½, convolutional code

[1] ACOLADE© is an example of a suitable product available commercially. This information is given for the convenience of users of this Technical Report and does not constitute an endorsement by IEC of this product.

The results in Figure 4.8.5 show the following: above the symbol rate R_s, the weighting characteristic follows the r.m.s. value of the impulsive signal that causes the interference. Below R_s, the weighting characteristic depends on the amount of coding: for the uncoded signal, the peak value increases with less than 10 dB per decade as the PRF decreases. With better coding, the part of the weighting characteristic with flat response becomes shorter. Therefore, it is important to characterize real radiocommunication systems in order to obtain meaningful results.

4.8.6 Experimental results

The methods described in 4.8.4 have been used for the measurement results in this part. The test signals are described where necessary.

4.8.6.1 Weighting in band A

For band A, i.e. below 150 kHz, no measurement results of digital radiocommunication systems are available.

NOTE Weighting of radio disturbance generally requires a consideration of intermittent, unsteady and drifting narrowband disturbances. Therefore the concept of defining a corner frequency, below which the average detector becomes effective has been applied to band A as well, using the corner frequency proposed for band B, since the original CISPR specification of the r.m.s. detector does not apply a meter time constant.

4.8.6.2 Weighting in band B

Weighting of interference to the Digital Radio Mondial (DRM) Broadcast System

At the World Radio Conference (WRC) in June 2003, the new Digital Radio Mondial was officially started. During the four week duration of the conference, a great number of special DRM transmissions became available from many radio stations. The measurements reported below, were taken on 8 July, 2003, when a great number of transmissions were still available.

DRM uses OFDM (Orthogonal Frequency Division Multiplex) with 200 carriers. The occupied bandwidth of each transmission is 10 kHz. In addition to the digitized audio signal, a certain amount of data (radio station information etc.) is transmitted. A conventional AM receiver can be used to downconvert the signal to an IF of 12 kHz, which is then decoded using a digital signal processor and a special DRM software radio.

During the time of measurement, the radio stations in table 4.8.2 were received at the station near Munich, with amateur dipole antennas mounted on the roof with a higher receive input voltage (50 to 60 dBuV) than required for the experiment.

Table 4.8.2 – DRM radio stations received for the measurement of the weighting characteristics

Frequency kHz	**Beam**	**Target**	**Av. DRM power** kW	**Program**	**Transmit site**
5975	060	W Europe	40	T-Systems Media Broadcast	Jülich
6095	ND	Europe	35	RTL/music and short announcements	Junglinster, Luxembourg
6140	ND	W & C Europe	40	DW English	Jülich
7320	105	W & C Europe	33	BBCWS	Rampisham
13605	037	C Europe	6	IBB/R. Sawa	Morocco
15440	040	W & C Europe	80	DW English	Sines
"W & C" means West and Central (Europe)					

The various transmissions were available for 1h or 2 h. The measurement results (weighting characteristics) were essentially the same for all frequencies, even if the amount of data transmitted in addition to the audio signal was not the same. Time dependent fading of the input signal had to be compensated for manually using a step attenuator that was inserted in the antenna connection, see Figure 4.8.7.

Principally the same type of interference signal was generated as in Figure 4.8.4. However, for a signal with an occupied bandwidth of 10 kHz, it is possible to use a longer pulse duration (10 µs or more).

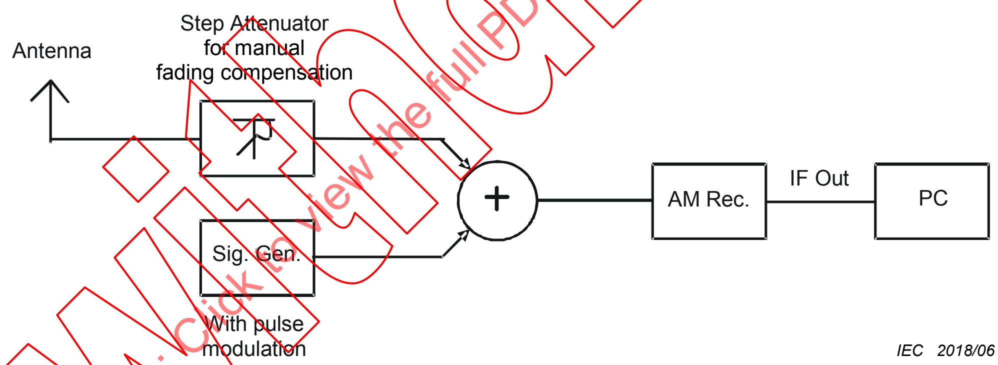


Figure 4.8.7 – Test setup for the measurement of weighting curves for Digital Radio Mondial (DRM). The received signal was downconverted to an IF of 12 kHz for decoding by special hard and software in a personal computer (PC)

Since no indication of BER was available, the "Audio" status indication on the PC (DRM software radio display) was used as a criterion. As soon as the interference becomes too high, the "Audio" status indication will turn from green to red.

As explained earlier, the signal level is attenuated so that the reception quality is just enough. The weighting characteristic (see Figure 4.8.8) shows a 10 dB/decade increase of the interference signal for PRFs between 1 kHz and 5 Hz. The nonlinearities are mainly due to uncompensated fading of the input signal. A detailed weighting curve is shown for a pulse width of 10 µs. For higher pulse widths, the weighting curve was measured only at three (resp. four) points to verify the 10 dB/decade behaviour. Below a PRF of 5 Hz, the weighting curve rises suddenly. And below about 2 Hz, the signal cannot be disturbed by the pulse width of 500 µs. However lightning strokes are reported to generate longer dropouts, which indicates that longer clicks might cause such dropouts as well.

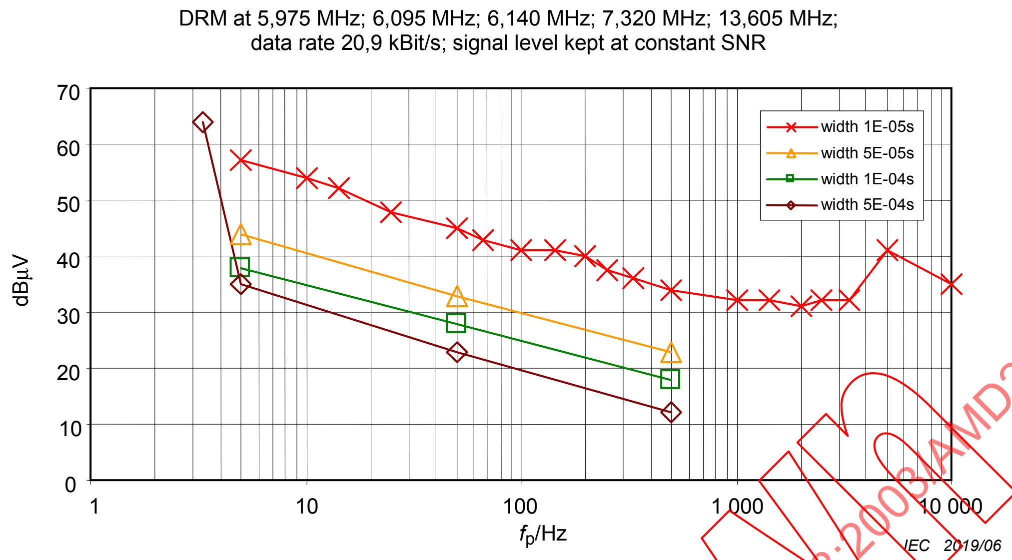


Figure 4.8.8 – Weighting characteristics for DRM signals for various pulse widths of the pulse-modulated carrier. Since the DRM signals are actual radio signals, the exact modulation scheme is not known

The report [2] describes the following DRM signals and two receiver types for the measurements:

- Mode B, Modulation 16/64QAM, Interleave 2 sec, protection level 1 / 0
- Mode B, Modulation 16/64QAM, Interleave 2 sec, protection level 0 / 0

The interference signal for Figures 4.8.9 and 4.9.10 is a pulse-modulated carrier with additional QPSK modulation in order to generate a wide bandwidth of the interference spectrum as explained in 4.8.4.2.

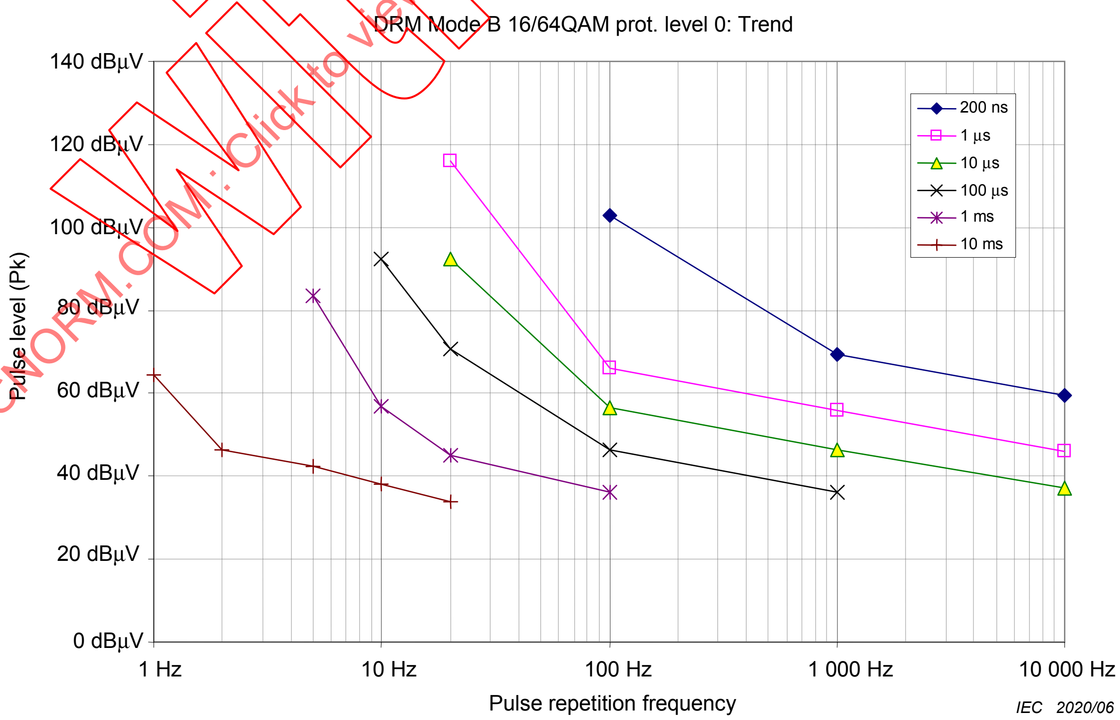


Figure 4.8.9 – Weighting characteristics for DRM protection level 0: average of results for two receivers

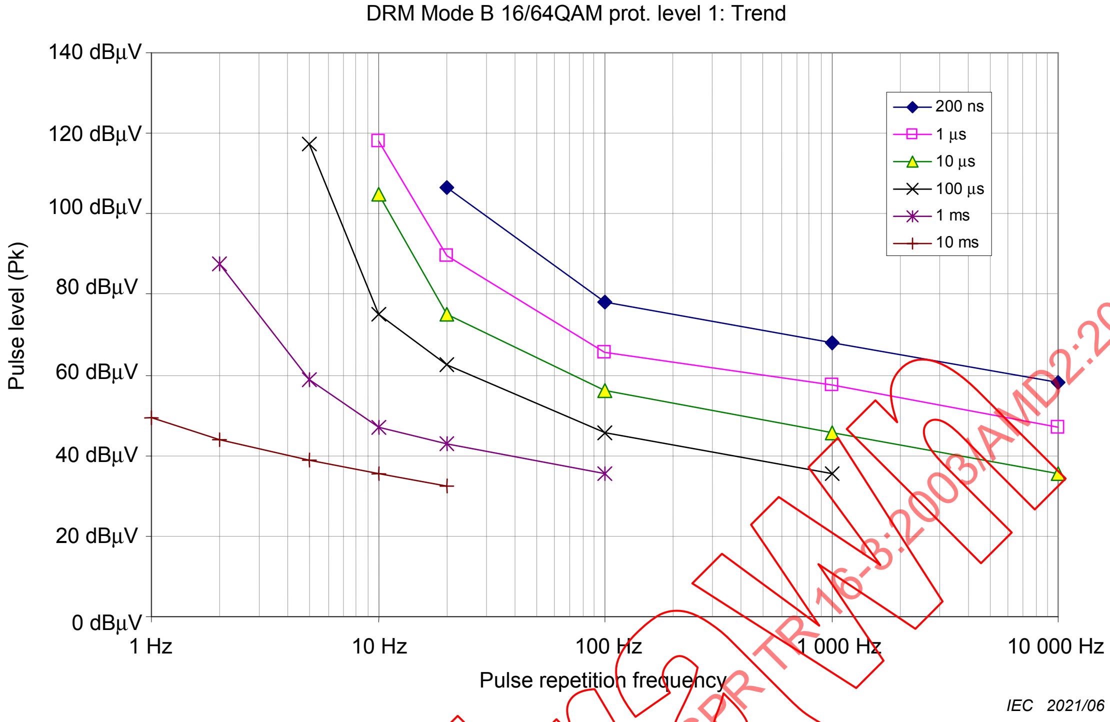


Figure 4.8.10 – Weighting characteristics for DRM protection level 1: average of results for two receivers

The weighting characteristics in Figures 4.8.9 and 10 show a 10 dB/decade slope down to approx. 100 Hz. Since there is no other digital radio system in band B, the corner frequency of the proposed RMS/AV detector between r.m.s. and linear average detection for this frequency band can only be based on the results of DRM (see 4.8.7). A corner frequency of 10 Hz is therefore proposed for band B as a compromise between the two results.

4.8.6.3 Weighting in bands C/D

4.8.6.3.1 Weighting of impulsive interference to Digital Video Broadcast Terrestrial (DVB-T)

- **Test setup**

One test setup for DVB-T consists of a DVB-T signal generator and a DVB-T measuring receiver. The components are connected via coaxial cables. The interference signal (a pulse-modulated carrier, see Figure 4.8.4 for an example of the spectrum) is fed into the signalling connection via a combiner.

The parameters used are the following.

DVB-T uses COFDM (Coded Orthogonal Frequency Division Multiplex) with 6817 (8k) or 1705 (2k) carriers. The OFDM carriers may be modulated either with QPSK (Quadrature Phase Shift Keying) or with 64 QAM (Quadrature Amplitude Modulation), resp. 16 QAM. QAM is preferred to QPSK as QAM allows higher data transfer rates. The transmission code rate *CR* is defined by *CR* = number of information bits/(number of information bits + error protection bits). Values of *CR* = 2/3 and 3/4 are used in actual systems. Each COFDM symbol is followed by a guard interval *GI* which is *GI* = 1/8 in actual systems. The DVB-T modulation and coding system allows many combinations, of which only a few are relevant. Therefore the parameters used in systems operating in some European countries have been selected. These allow transmission rates between 14,745 Mbit/s and 24,88 Mbit/s (see Table 4.8.4) depending on modulation and code rate. Different coders and decoders are used in the system. The bit-error ratio (BER) reading can be taken before the Viterbi decoder as well as before and after the Reed Solomon decoder of the measuring receiver. A comparison is given in Table 4.8.3. The transmission level is set so that the BER after the Reed Solomon decoder without interference is just below 10^{-8}. This results in different signal levels depending on the system parameters. The interference levels have then been adjusted to a critical value of BER = $2{,}0*10^{-4}$ before the Reed Solomon decoder.

For the BER measurement, the modulator generates a Pseudo Random Binary Sequence (PRBS) as data stream. The evaluation of the data stream is done in the receiver in two different procedures. The BER before Viterbi and before Reed-Solomon is evaluated by correlation. Flags in the bit stream are used to determine the BER after Reed-Solomon. If the decoder does not recognize a flag as correct, the following bit combination is determined to be false.

The relationship in Table 4.8.3 was found experimentally between the bit error ratios before and after the Viterbi and Reed Solomon decoders for two pulse rates.

Table 4.8.3 – Comparison of BER values for the same interference level

Pulse rate Hz	10 k	500 k
BER before Viterbi decoder	$1{,}5*10^{-2}$	$4{,}4*10^{-3}$
BER before Reed Solomon	$2{,}0*10^{-4}$	$2{,}0*10^{-4}$
BER after Reed Solomon	$1{,}0*10^{-6}$	$1{,}0*10^{-8}$

So, the results with BER measured before Reed Solomon (with $2{,}0*10^{-4}$) and after Reed Solomon (with $1{,}0*10^{-6}$) are roughly comparable.

Table 4.8.4 – Transmission parameters of DVB-T systems used in various countries

Country	Modulation	Code rate	Guard interval	Transfer rate
France/UK	64QAM 2k	3/4	1/8	24,88 Mbit/s
Spain	64QAM 8k	3/4	1/8	24,88 Mbit/s
Germany	16QAM 8k	2/3	1/8	14,745 Mbit/s

The measurement results are presented in Figures 4.8.11, 4.8.12 and 4.8.13. In all tests, the interference signal leading to these results are pulse-modulated carriers.

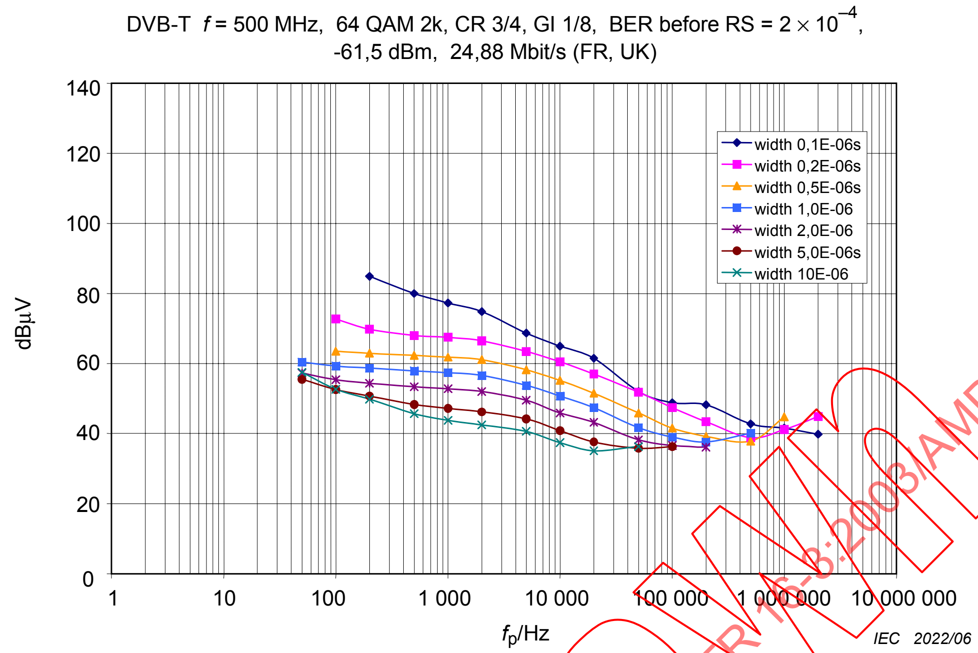


Figure 4.8.11 – Weighting characteristics for DVB-T with 64QAM 2k, *CR* 3/4 (as used in France and United Kingdom)

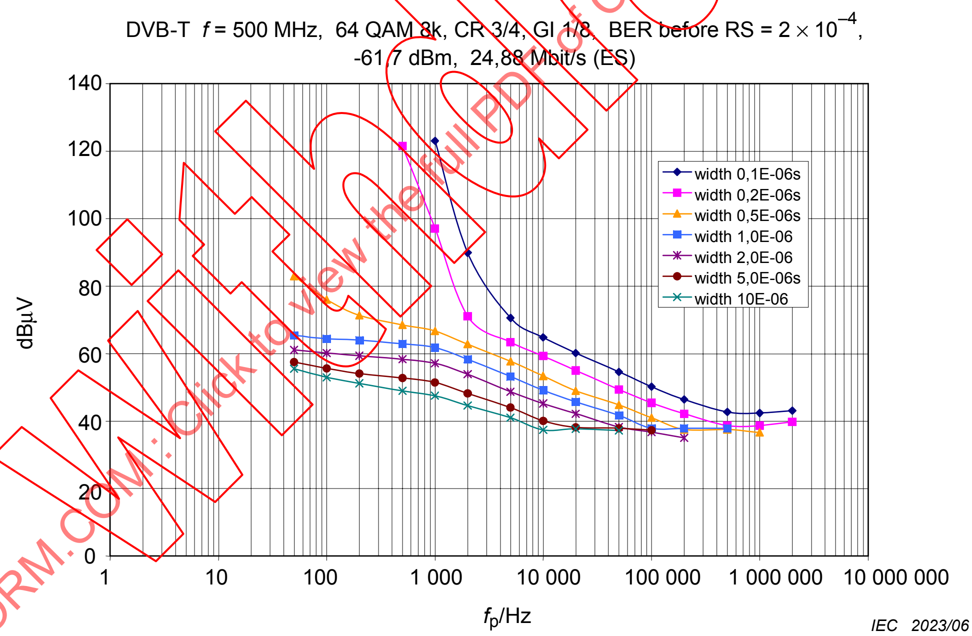


Figure 4.8.12 – Weighting characteristics for DVB-T with 64QAM 8k, *CR* 3/4 (as used in Spain)

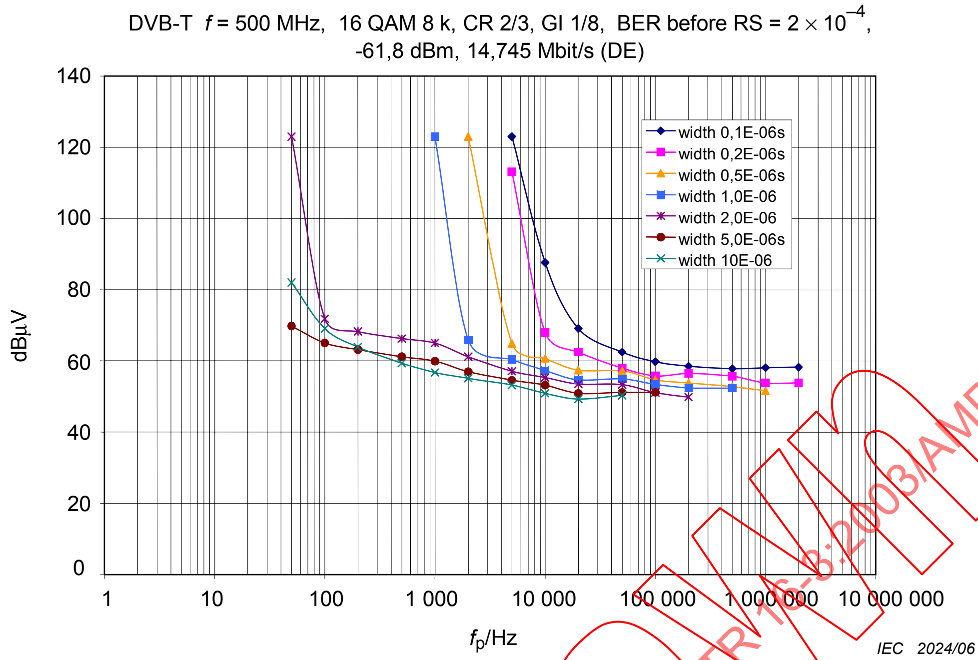


Figure 4.8.13 – Weighting characteristics for DVB-T with 16QAM 8k, *CR* 2/3 (as used in Germany)

A number of 6 different receiver types were tested in report [2] for DVB-T with 16QAM 8k, *CR* 2/3 and for DVB-T with 64QAM 8k, *CR* 2/3. To get receiver independent results, the individual characteristics were combined using average values inside the range where all receivers offered a result. Excluded were two receivers in PRF ranges, where they showed a non-typical behavior. These combined results are shown in the "trend" characteristics in Figures 4.8.14 and 4.8.15. The interference signal for both figures is a pulse-modulated carrier with additional QPSK modulation in order to generate bandwidth of the interference spectrum at least as wide as the DVB-T signal spectrum as explained in 4.8.4.2.

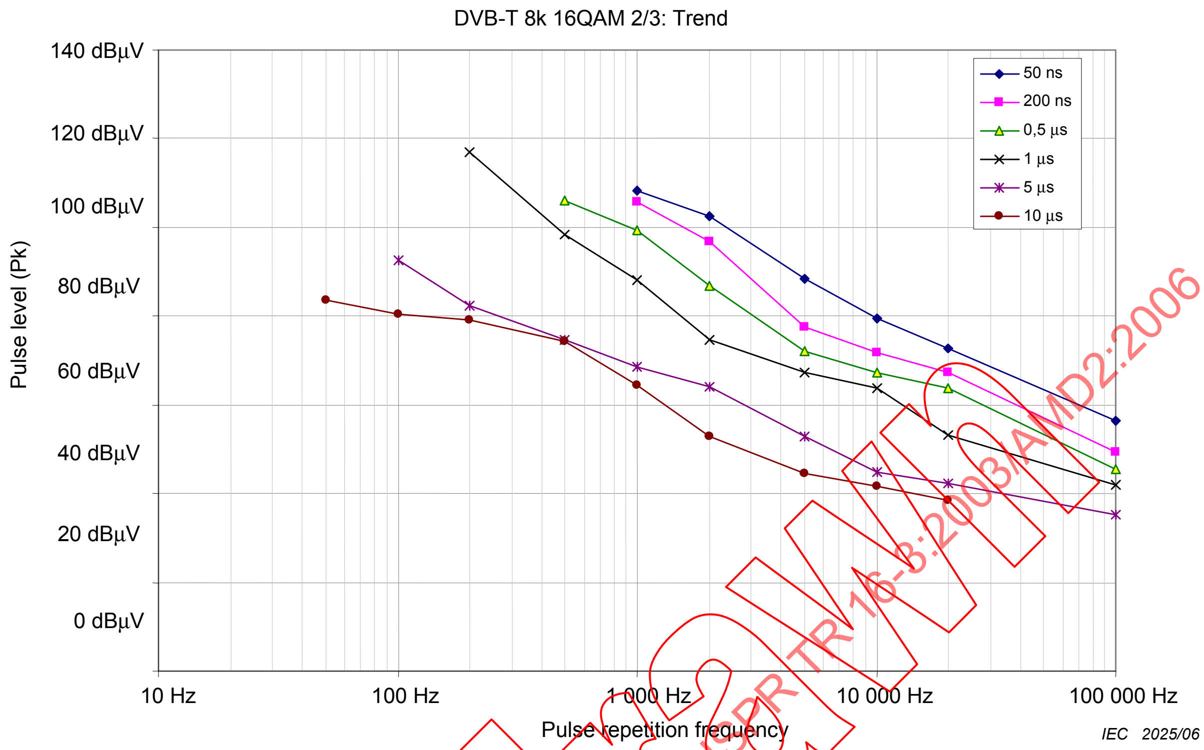


Figure 4.8.14 – Average weighting characteristics of 6 receiver types for DVB-T with 16QAM

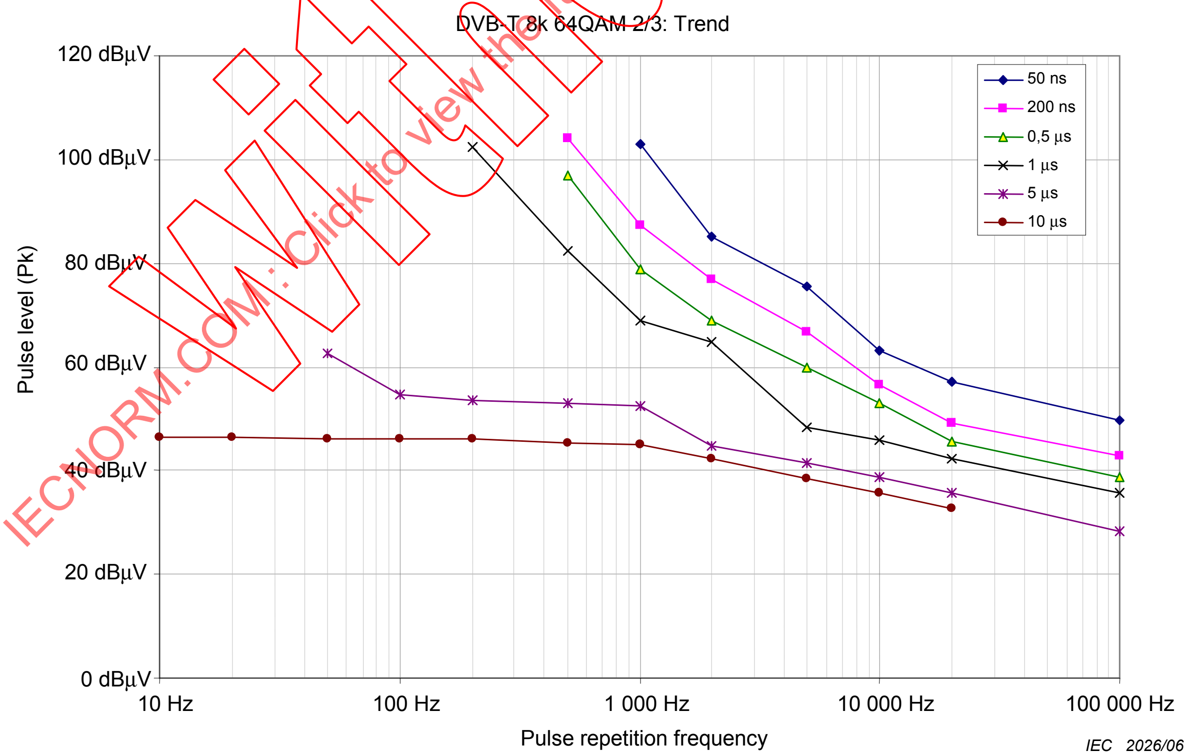


Figure 4.8.15 – Average weighting characteristics of 6 receiver types for DVB-T with 64QAM

- **Interpretation of the results**

In Figure 4.8.11, the corner frequency can only be assumed to be approx. 100 Hz, whereas in Figures 4.8.12 and 4.8.13, the corner frequencies can clearly be seen. They however depend on the interference pulse width as in Figures 4.8.11 and 4.8.13. Since all weighting curves are given for the shortest pulse (see Figure 4.8.1), also for the corner frequency, the shortest pulse is always relevant. The system used in Germany shows the most robust performance against impulsive interference due to its lower code rate and 16QAM 8k modulation.

4.8.6.3.2 Weighting of impulsive interference to other digital radiocommunication systems operating in CISPR bands C and D

- **Digital Audio Broadcasting (DAB)**

DAB operates in the VHF (174 MHz to 230 MHz) and the L (1 452 MHz to 1 492 MHz) bands with a bandwidth of 1,5 MHz per channel using Coded Orthogonal Frequency Division Multiplex (COFDM) to minimise multipath fading. The audio signal data rate is reduced by MUSICAM (a masking pattern adapted for Universal Coding and Multiplexing), which is a part of the MPEG-2 (Moving Picture Expert Group) standard. The total transmitted bit rate is 2,4 Mbit/s. The 1500 subcarriers are modulated using Differential QPSK (DQPSK). The weighting characteristics in Figure 4.8.16 were measured using a test version of a DAB receiver. Weighting characteristics of commercial DAB receivers have been presented in report [2].

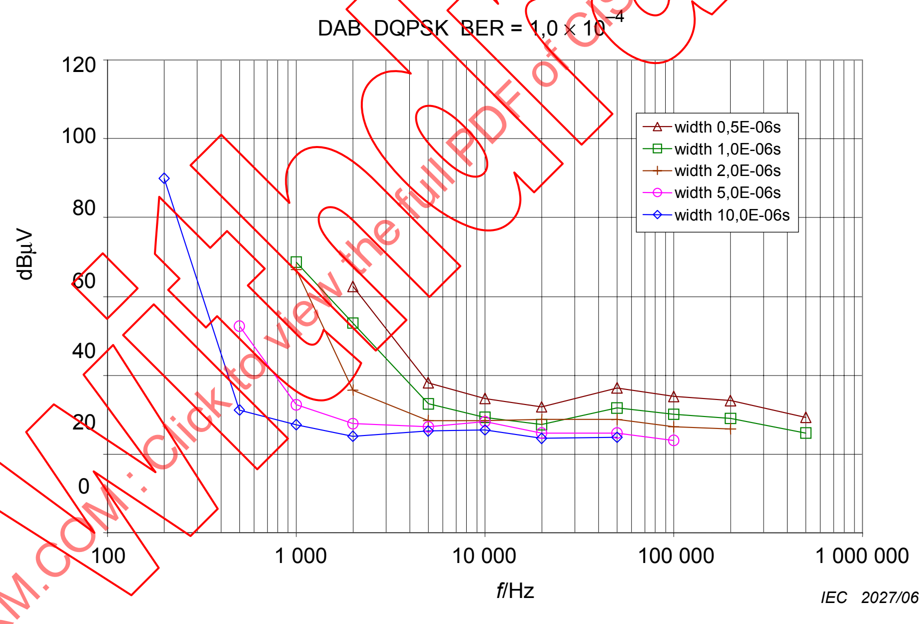


Figure 4.8.16 – Weighting characteristics for DAB (signal level -71 dBm) with a flat response down to approximately 1 kHz

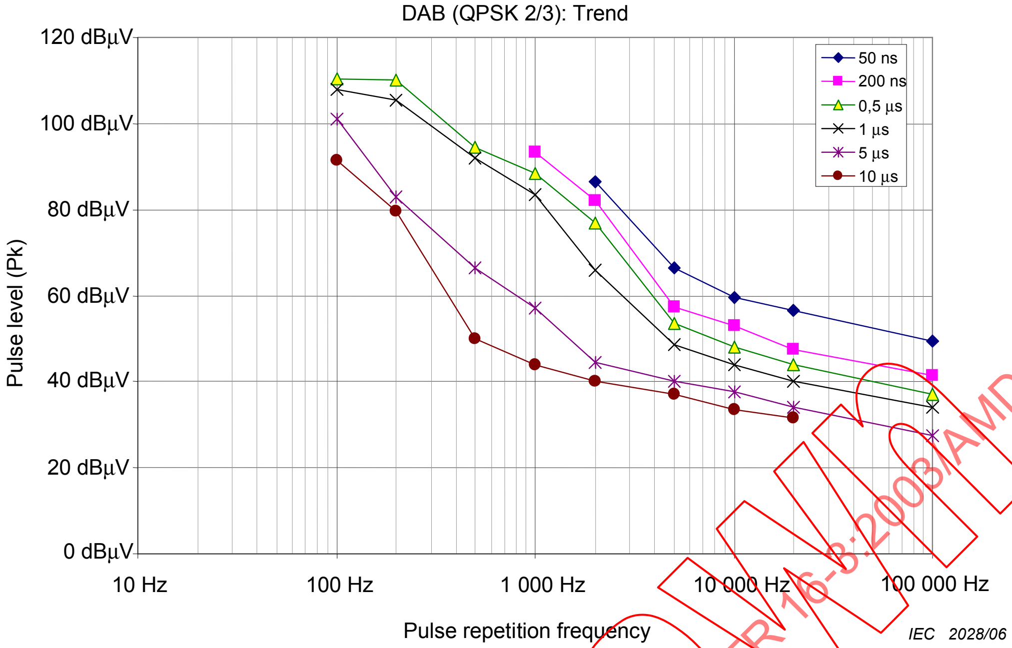


Figure 4.8.17 – Weighting characteristics for DAB: average of two different commercial receiver types

The differences between the results in Figures 4.8.16 and 4.8.17 are possibly due to the different types of the impulsive signal: for Figure 4.8.16 a simple pulse-modulated carrier was used, whereas for Figure 4.8.17 an on/off-keyed QPSK-modulated signal was used.

- **Terrestrial Trunked Radio (TETRA) system**

TETRA is used in workshops, the building and construction industries, airports, transportation/trucking and safety services. It operates in the frequency range 380 MHz to 520 MHz (in some areas also in 870 MHz to 990 MHz) with a data rate of 36 kbit/s per carrier, an occupied bandwidth of ≈ 25 kHz and channel separations of 12,5, 20 or 25 kHz. Speech data reduction is done using Algebraic Code Excited Linear Prediction (ACELP) to 4,8 kbit/s per traffic channel. Up to four traffic channels are normally transmitted on one carrier. The error protection may be high or low, depending on the code rate. The modulation procedure is π/4-DQPSK. Figure 4.8.18 shows the measured weighting characteristics for a high code rate = 1 (low error protection).

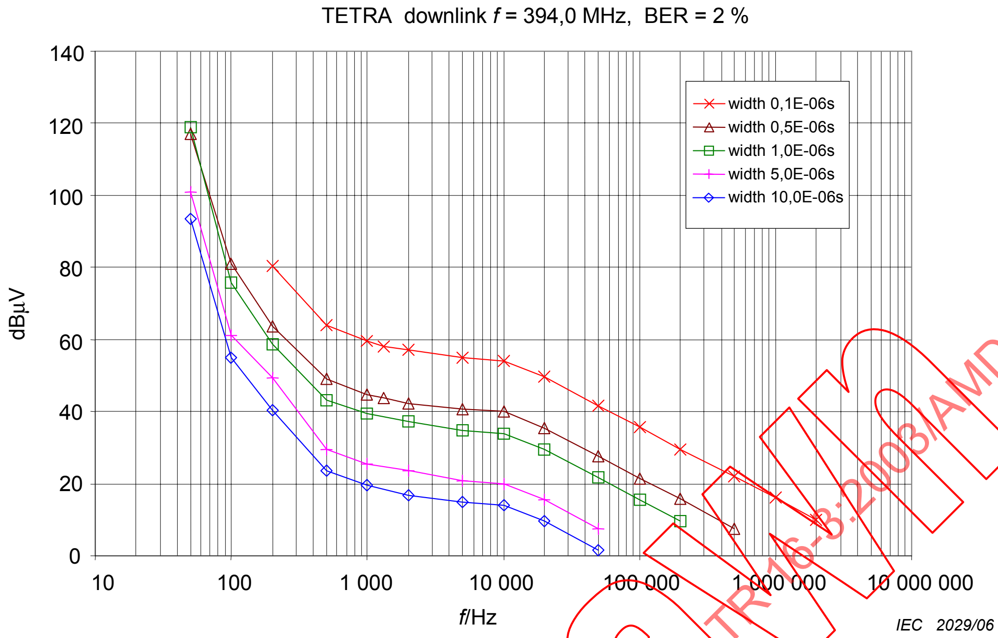


Figure 4.8.18 – Weighting characteristics for TETRA (signal level – 80 dBm) for a code rate of 1

Since the pulse spectrum is much wider than the channel bandwidth, all weighting characteristics are separated by the PRF ratio in dB. Above a PRF of 10 kHz, the slope of curves is 20 dB/decade, corresponding to the increase of the voltage of the center line of the interference spectrum. Therefore the weighting characteristics below 10 kHz PRF should be regarded as relevant.

- **Global System for Mobile Communication (GSM)**

This digital cellular telecommunication system operates in the 900 MHz (GSM 900) and 1 800 MHz (GSM 1800) frequency bands. The offset between uplink (mobile to base station) and downlink is 45 MHz (GSM 900) and 95 MHz (GSM 1800) respectively. The occupied bandwidth is 300 kHz and channel spacing is 200 kHz. Modulation for constant spectrum envelope is achieved with Gaussian Minimum Shift Keying (GMSK). The error correction mechanisms applied are different for traffic channels (1b bits) and other bits (Class 2 bits). Therefore different bit error rates apply: BER, RBER 1b and 2 (residual BER) and FER (Frame error rates). The test setup and signals of Figures 4.8.3 and 4.8.4 have been used, with a mobile communication tester as a signal source.

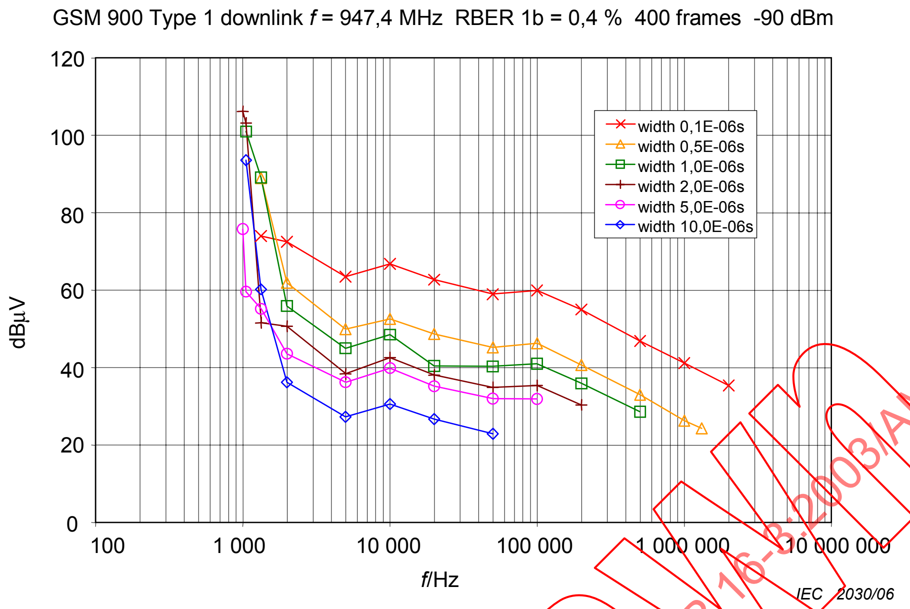


Figure 4.8.19 – Weighting characteristics for RBER 1b of GSM (signal level –90 dBm)

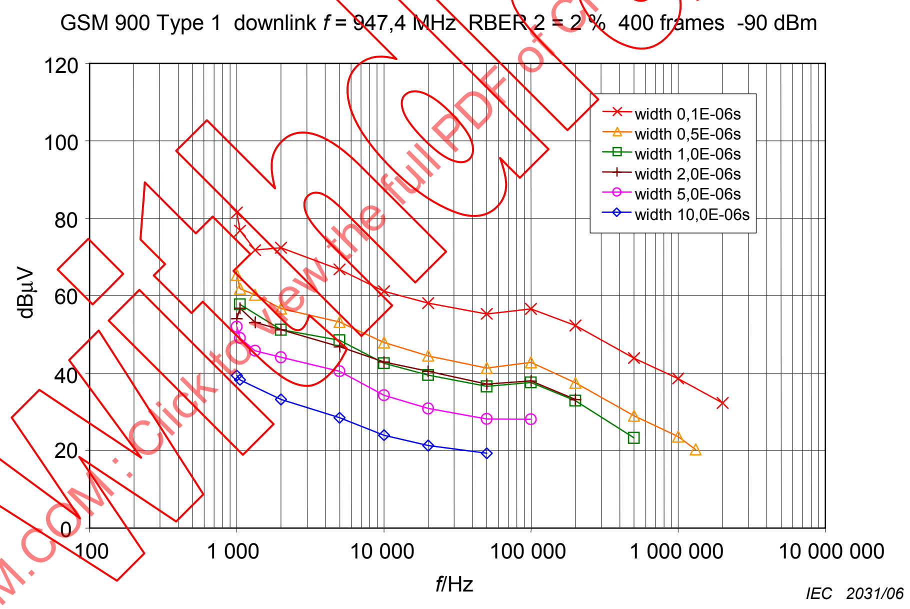


Figure 4.8.20 – Weighting characteristics for RBER 2 of GSM

The characteristics typically rise at 10 dB/decade between 100 kHz and 2 kHz with a steeper slope below about 2 kHz PRF. Unfortunately measurements below a PRF of 1 kHz were not possible due to instability of the test system. The results shown in Figures 4.8.19 and 4.8.20 are very similar to the BER and RBER 1b curves of Figure 4.8.21 similar to those published in [7] and [8] using the simulation software COSSAP. The values obtained in Figure 4.8.21 have been calculated assuming a pulse-modulated carrier with a pulse duration of 2 µs as the interference signal.

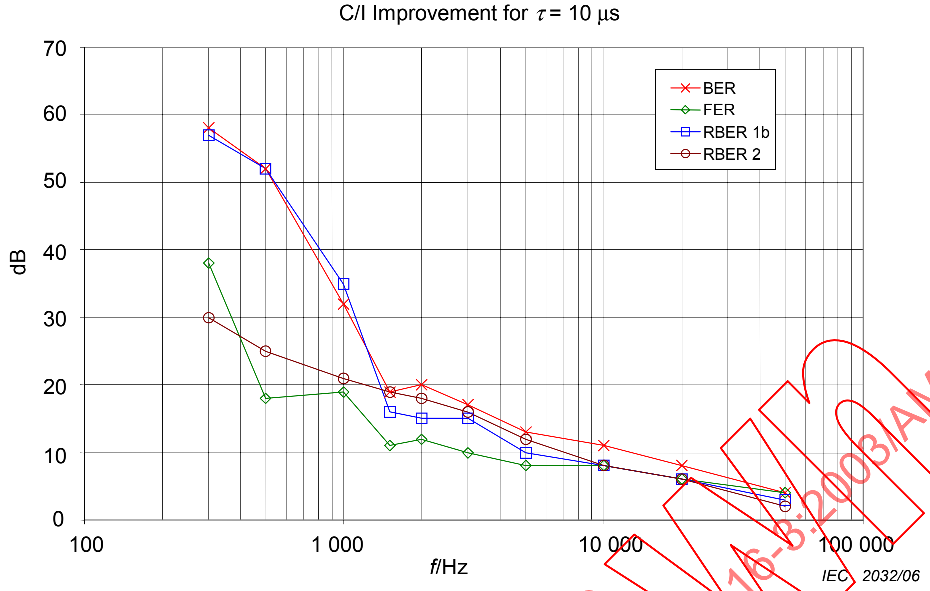


Figure 4.8.21 – Carrier-to-interference improvements with decreasing PRF in dB computed for GSM using COSSAP

- **Frequency Modulation (FM) Broadcast System**

Based on the assumption that FM broadcast will survive past the transition from analog to digital radio systems for some time, measurements have been made based on the methods of report [2] resulting in Figure 4.8.22. The FM signal contained a pilot carrier only; the increase of noise due to the interference was measured in the demodulated signal. The interference signal is a pulse-modulated carrier with additional QPSK modulation in order to generate bandwidth of the interference spectrum at least as wide as the FM signal spectrum as explained in 4.8.4.2.

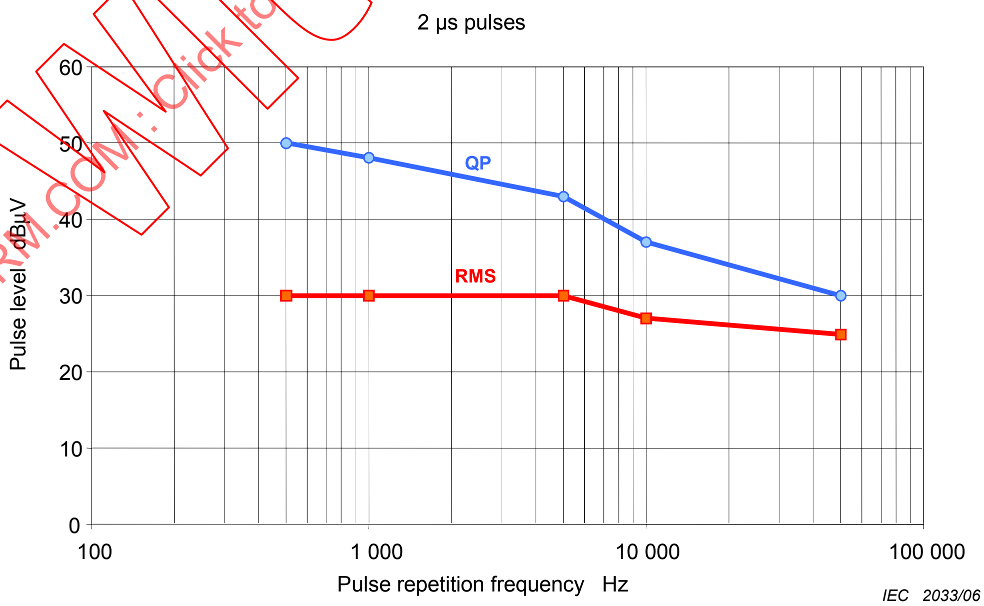


Figure 4.8.22 – RMS and quasi-peak values of pulse level for constant effect on FM radio reception

Figure 4.8.22 is not a weighting characteristic! It shows that the r.m.s. value of the pulse level with 2 µs width is closer to being constant than the quasi-peak value. This has been shown for other pulse widths as well but is not presented here for reasons of space.

4.8.6.4 Weighting for Band E (1 through 18 GHz)

- **GSM system**

The weighting characteristics found for a mobile operating in the 1 800 MHz (GSM 1800) frequency band is very similar to the system operating in the 900 MHz (GSM 900) frequency band (compare Figure 4.8.23 with Figures 4.8.19 through 4.8.21). The offset between uplink (mobile to base station) and downlink is 95 MHz for GSM 1800. As in Figures 4.8.19 through 4.8.21, the curves are rising below 2 kHz PRF with a slope of more than 20 dB/decade.

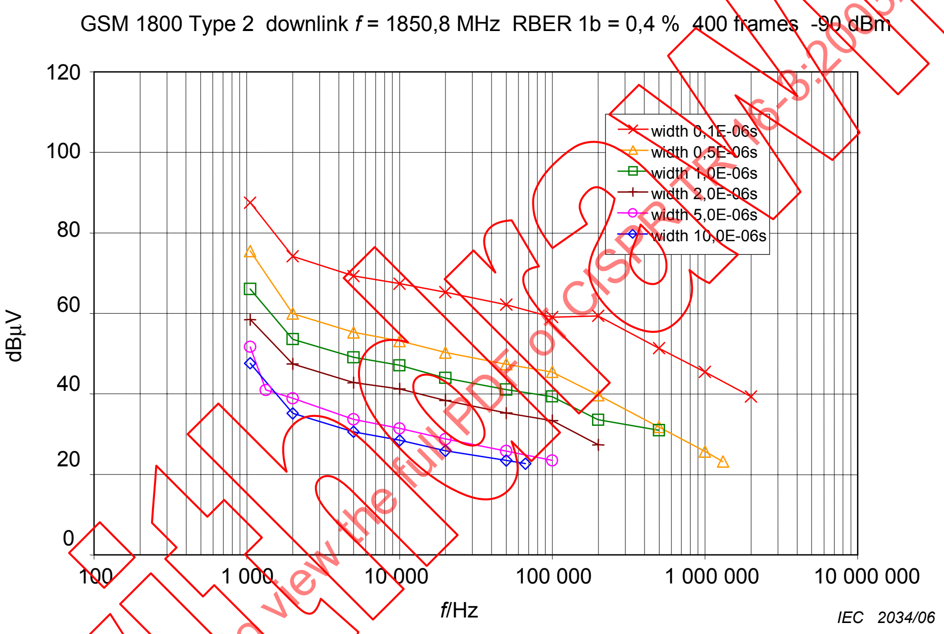


Figure 4.8.23 – Weighting characteristics for RBER 1b of GSM (signal level –90 dBm)

- **Digitally Enhanced Cordless Telephone (DECT) system**

DECT is used in homes and offices for distances up to 300 m (in picocells). It provides 10 channels spaced 1,728 MHz apart in the frequency range 1,88 to 1,90 GHz. The occupied bandwidth is ≈ 1,5 MHz. For speech data reduction Adaptive Differential Pulse Code Modulation (ADPCM) is used. Modulation is done with Gaussian Mean Shift Keying (GMSK). The data stream for testing is Pseudo Random Binary Sequence (PRBS).

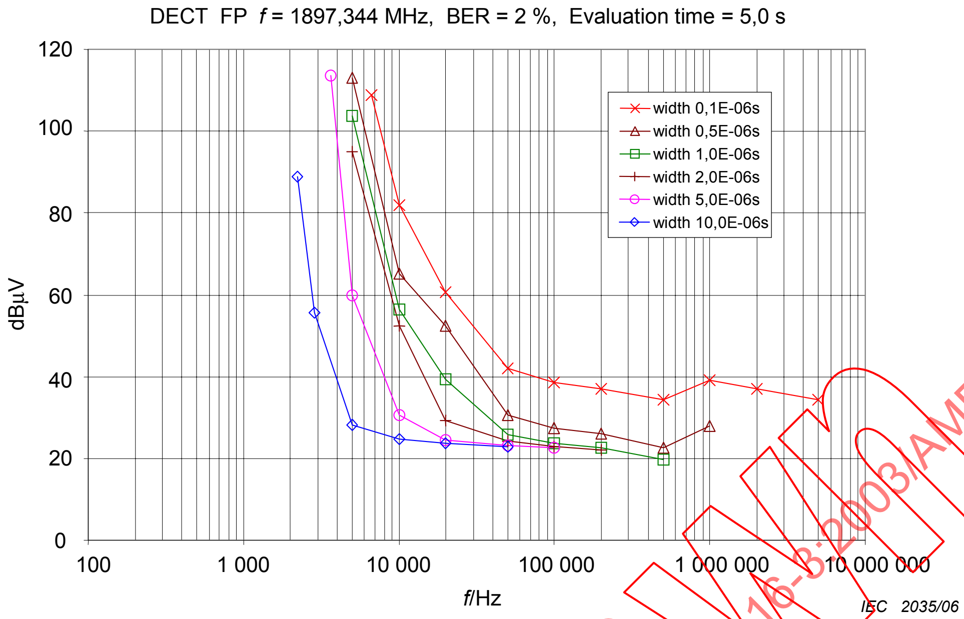


Figure 4.8.24 – Weighting characteristics for DECT (signal level –83 dBm)

The weighting characteristics for DECT show a response near 10 dB/decade in the range between 50 kHz and 500 kHz PRF in the upper PRF areas for narrow pulses and a steep slope below about 10 kHz PRF. Only for longer pulse widths, the weighting characteristic is flat.

- **Code Division Multiple Access (CDMA) systems IS-95 and J-STD 008**

IS-95/J-STD 008 have been specified by TIA (US Telecommunications Industry Association) and are used in the frequency ranges 825 MHz to 900 MHz (IS-95) and 1,8 GHz to 2,0 GHz. The occupied bandwidth is ≈ 1,4 MHz (3 dB: 1,23 MHz). The modulation is done with Quadrature Phase Shift Keying (QPSK). For the uplink (mobile to base station) the optimum setting of the receive power at the base station is controlled via power control bits.

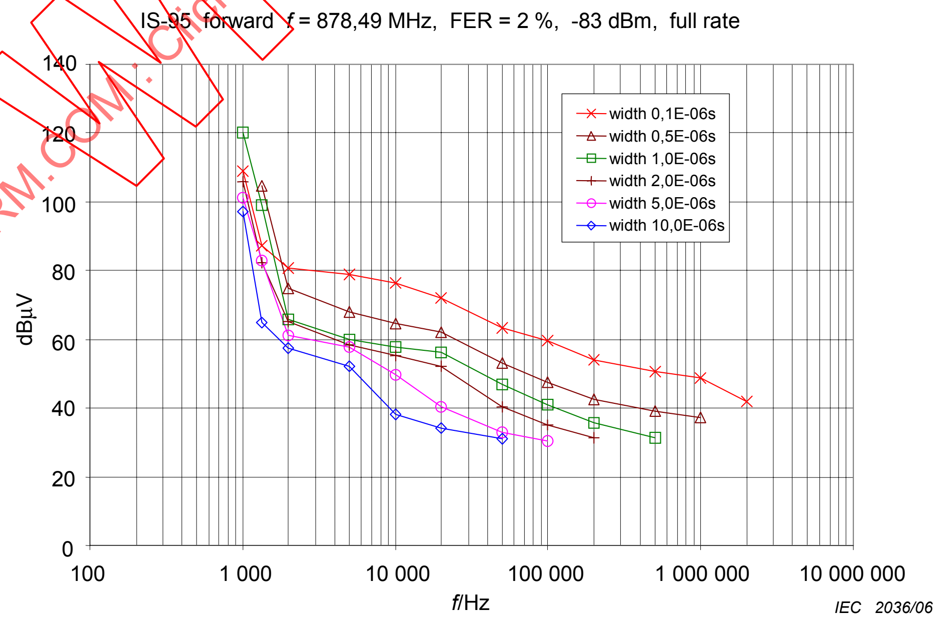


Figure 4.8.25 – Weighting characteristics for IS-95 (signal level -97 dBm) with comparatively high immunity to interference

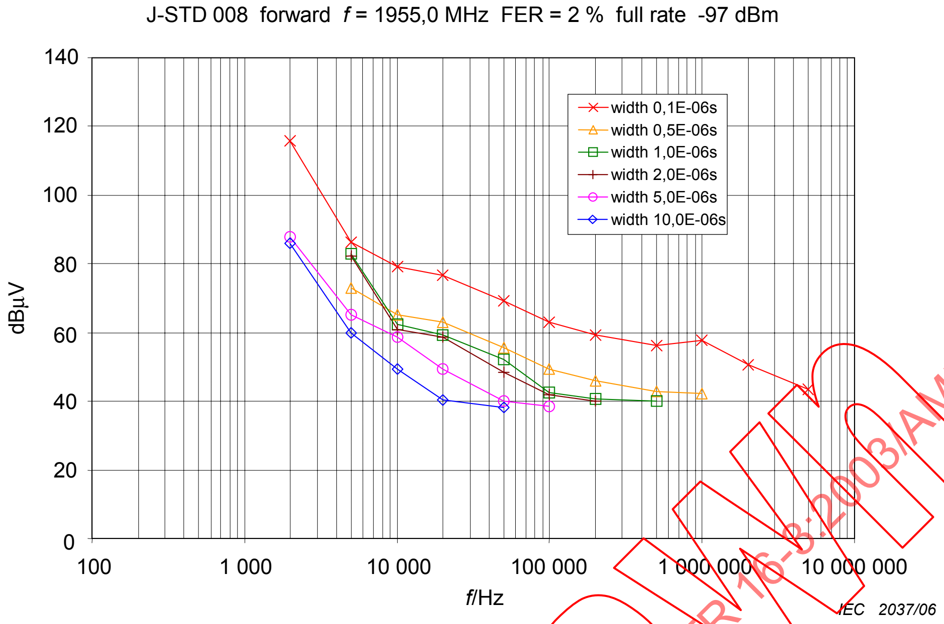


Figure 4.8.26 – Weighting characteristics for J-STD 008 (signal level –97 dBm)

- **3rd Generation Digital Radiocommunication Systems**

Two different systems have been investigated:

– Wideband CDMA (W-CDMA), which is going to be deployed in Europe, and

– CDMA2000, which is mainly going to be applied in North America and some other areas.

Tests have been made on both systems. However at the time of testing, available mobile phones for W-CDMA did not give stable BER results in the test setup (loop back) with the mobile testers. So, only results for cdma2000 are available now. Results for W-CDMA will certainly become available at a later date. They have been used later with success for evaluating the interference effect to spread-spectrum clock signals (see [3] and [4]).

CDMA2000 as described by Third Generation Partnership 2 (3GPP2) is an access method intended for use in the IMT-2000 proposal for Third Generation (3G) cellular telephone systems. The system is based on spread-spectrum codes and provides high and variable data rates. It is an evolutionary development of IS-95 (cdmaOne) which is also based on Code Domain Multiple Access (CDMA) to the air interface. This means that the individual channels are separated from each other by individual codes. The basic chip rate is 1,2288 MHz. All IMT-2000 compatible systems feature transmitted bit rates of up to 384 kb/s up to a cruising speed of 500 km/h, in urban areas up to 120 km/h.

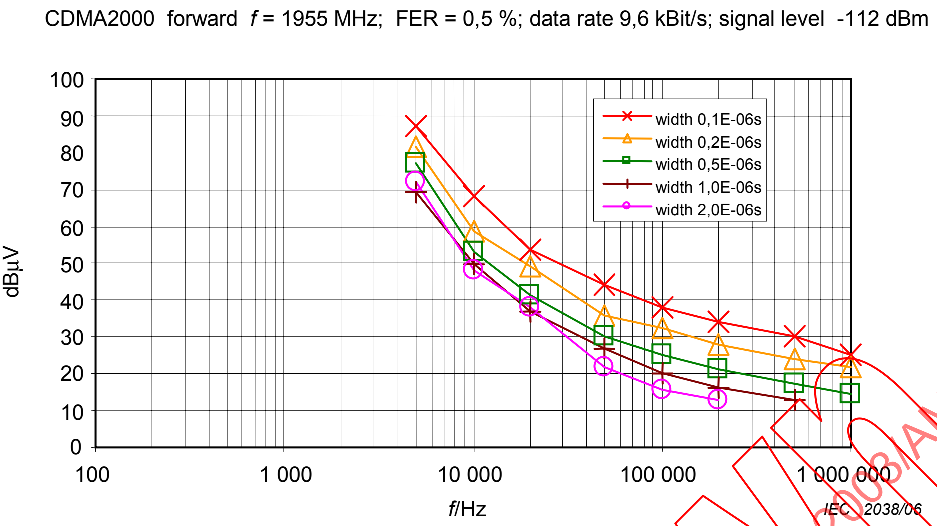


Figure 4.8.27 – Weighting characteristics for the Frame Error Ratio (FER) of cdma2000 (measured at a receive signal level of –112 dBm) for a low data rate of 9,6 kb/s. The curves are rising increasingly fast at lower PRFs

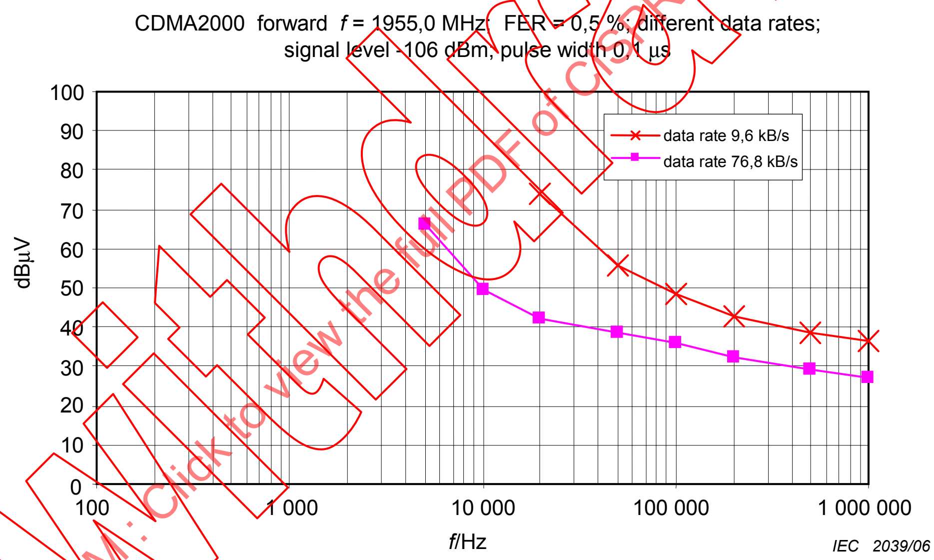


Figure 4.8.28 – Weighting characteristics for the Frame Error Ratio (FER) of cdma2000 (measured at a receive signal level of – 106 dBm) for two different data rates (9,6 and 76,8 kb/s)

For higher data rates (e.g. 384 kb/s, which was not available for the test), the system would still be more susceptible to impulsive disturbance. For the higher data rates, the faster rising knee (corner frequency) is shifted to lower PRFs but remains high compared to 1 kHz.

4.8.7 Effects of spread-spectrum clock interference on wideband radiocommunication signal reception

It was argued that the classical impulsive (i.e. broadband) and unmodulated or pulse-modulated narrowband interference may not completely reflect today's sources of interference. In this context, the question on the effect of spread-spectrum-clock signals came up. This type of emission has in the recent past raised the concern of spectrum regulators and some studies have been published. Spread-spectrum clocking reduces the measured emission level of computer clocks, but what is the effect of spread-spectrum clocking on victim radiocommunication systems? Spread-spectrum clock interference was expected to have especially a severe effect on wideband radiocommunication services. Therefore DVB-T and W-CDMA as well as CDMA2000 were selected for tests and presented in [3] and [4].

The application of frequency modulation or other spread spectrum modulation to the clock signal distributes the clock power over a frequency band wider than the EMI measurement bandwidth and thus reduces the emission level, when measured with a bandwidth as narrow as e.g. 120 kHz. In table 4.8.5 measurements are shown for a frequency-modulated clock signal spectrum and of the corresponding unmodulated clock signal with f_{centre} = 500 MHz, f_{mod} = 30 kHz (sinewave), spread amount δ = 3,5 MHz (i.e. the spectrum width due to modulation) and a peak level reduction Δ = 5,0 dB.

Table 4.8.5 – Example of measurement results in dB(µV) of unmodulated and FM modulated carriers for various detectors (bandwidth 120 kHz)

Detector	Unmodulated carrier	FM modulated carrier for highest peak	FM modulated carrier for centre frequency
PK in dB(µV)	55,6	50,39	44,3
QPK in dB(µV)	55,4	49,30	43,16
AV in dB(µV)	55,38	38,38	37,12
RMS in dB(µV)	55,38	42,50	38,87

Using the measurement bandwidth of 1 MHz at 2 GHz (with a proportionally higher spread amount) reduces the differences between unmodulated and FM modulated carriers to 1,2 dB for the peak detector, 17 dB for the average detector and to 11 dB for the r.m.s. detector.

The measurement results in [3] and [4] show that the high immunity of the DVB-T and W-CDMA systems to unmodulated carriers (i.e. clock signals) is lost to frequency modulation. Considering

– that due to a lower measured emission level, the interference level may be increased by the amount of Δ (e.g. 6 dB) to reach the same emission limit and

– that frequency modulation causes a reduction of up to 25 dB of the original immunity of the system to clock signals,

frequency modulation causes a total increase of possible interference effect of up to 25 + 6 dB = 31 dB.

It is agreed that some digital modulation systems have been especially designed to suppress unmodulated interference. In general, however it must be pointed out that the EMI measurement bandwidth should match the bandwidth of possible victim radiocommunication receivers. It might therefore be an advantage to describe the detector function for various measurement bandwidths. If an r.m.s. detector is used for the measurement of spread-spectrum modulated emissions, the measured value will be proportional to the square-root of the measurement bandwidth. For the frequency range of CISPR bands C and D, the radiocommunication signal bandwidths have always had a wide range of values. Narrowband FM with as few as 7,5 kHz on the one hand and the amplitude-modulated TV signal spectrum including the residual sideband with as much as 6 MHz on the other were in use until recently and the 120 kHz was used as the measurement bandwidth. This situation has not changed very much with the introduction of TETRA (bandwidth approx. 25 kHz) and DVB-T (bandwidth 6,6 MHz (VHF) and 7,6 MHz (UHF)).

4.8.8 Analysis of the various weighting characteristics and proposal of a weighting detector

Looking at various results of weighting characteristics in the sections above, we can see that above a certain corner frequency, the weighting function decreases with approximately 10 dB per decade of pulse repetition frequency. A decrease of 10 dB per decade corresponds to the weighting function of an r.m.s. detector (see Figure 4.8.2). Below this corner frequency, the weighting function decreases with a higher rate. A higher rate of decrease (20 dB/decade) can be achieved using the linear average detector function. This behaviour can be approximated by a combination of two detectors, the r.m.s. and the linear average detector. The average detector applies the meter time constant as described in CISPR 16-1-1 for intermittent, unsteady and drifting narrowband disturbances. Figure 4.8.29 serves to understand the meaning of the corner frequency. It is not possible to satisfy the protection requirements of all services with the same perfection, therefore the selection of the various corner frequencies between the proposed average and r.m.s. weighting functions in each band can be regarded as a compromise. Where corner frequencies for different pulse widths are different, the corner frequency for the shorter pulse widths apply, as the detector weighting always applies to the shortest possible pulse width, which is determined by the measurement bandwidth. It is proposed to keep the measurement bandwidths specified in CISPR 16-1-1 for the CISPR bands A through E .

Table 4.8.6 – Survey of the corner frequencies found in the various measurement results

System	Reference	f_c kHz	Comment
DRM	2, 10	0,1/0,005	10 Hz chosen for feasibility
DVB-T	2, 11	0,1 - 10	f_c depending on w_p, modulation and coding
DAB	2, 9	5	f_c partially depending on w_p
TETRA	9	0,5	narrowband system, mainly used below 1 GHz
GSM 900	7, 8, 9	1,5	above f_c:very close to r.m.s.
FM	(2) unpubl.	< 0,5	weighting characteristics follows r.m.s. down to 0,5 kHz
GSM 1800	7, 8, 9	1,5	above f_c:very close to r.m.s.
DECT	9	50	above f_c: flatter than r.m.s.
IS-95	9	2	very similar to J-STD 008; above f_c close to r.m.s.
J-STD 008	9	5	very similar to IS-95; above f_c close to r.m.s.
CDMA2000	9	50	data rate 9,6 kb/s; above f_c, curves are very close to r.m.s.
CDMA2000	9	10	data rate 76,8 kb/s; above f_c, curves are very close to r.m.s.

Where f_c is the corner frequency and w_p is pulse width

As a result of the values found in Table 4.8.6, the following corner frequencies were selected:

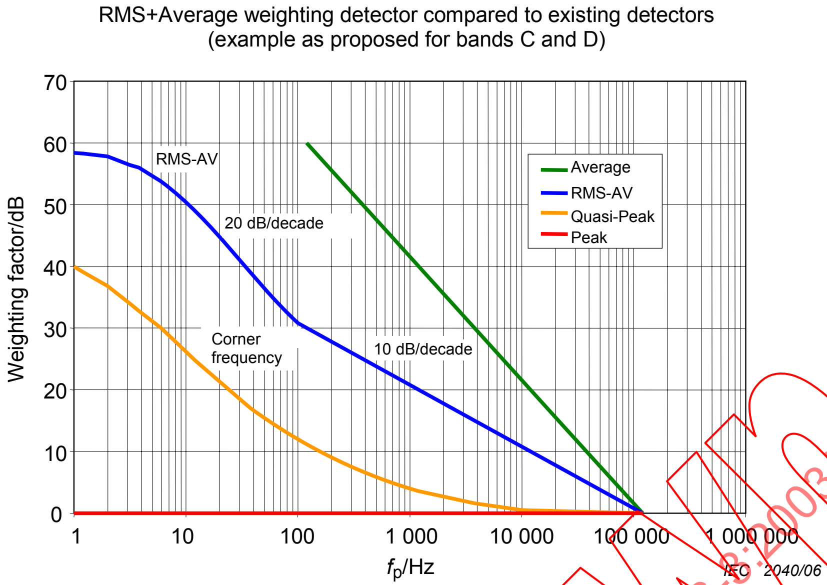
Band A: 10 Hz (same as band B)

Band B: 10 Hz

Bands C/D: 100 Hz

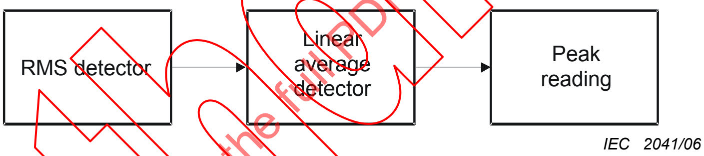
Band E: 1000 Hz

NOTE The corner frequency of 10 Hz was selected for band A, in order to give the r.m.s. detector a function similar to the one of band B and in addition to allow the use of the meter time constant in order to provide an asymptote for the purpose of measuring intermittent, unsteady and drifting narrowband disturbances with the r.m.s. detector.



The asymptote of 58,7 dB near 1 Hz is due to the average detector meter time constant.

Figure 4.8.29 – The proposed r.m.s.-average detector for CISPR bands C and D with a corner frequency of 100 Hz



The linear average detector has an inherent meter time constant. The maximum output of the average detector is taken using a peak reading function.

Figure 4.8.30 – RMS-average detector function by using an r.m.s. detector followed by a linear average detector and peak reading

A digital r.m.s. detector with r.m.s. computing times of 10 ms, followed by a digital linear average detector, results in the r.m.s.-average weighting curve of Figure 4.8.29 for the shortest pulse width allowed by the measurement bandwidth of 120 kHz. RMS computing times of 10 ms will give r.m.s. values of the disturbance signal within 10 ms. The 10-ms packets are then weighted using a linear average function. The peak reading function after a meter time constant of 100 ms is effective then for low repetition pulses (f_p below 10 Hz) which causes the weighting curve to approximate the asymptote of 58,7 dB.

Conclusion: it has been shown experimentally and partly numerically that some detector functions that are currently in use in CISPR product standards

– either indicate a higher interference potential of impulsive disturbance than the interferer actually represents (i.e. they overweigh the disturbance) if “peak” and “quasi-peak” detectors are used, or

– indicate a lower interference potential of impulsive disturbance than the interferer really represents (i.e. underweigh or de-emphasize the disturbance) for the “average” detector with respect to the possible interference effect on digital radiocommunication systems, whereas using the r.m.s.-average detector represents the interference effect rather well.

4.8.9 Properties of the r.m.s.-average weighting detector

For CISPR weighting functions, the pulse width is always assumed to be defined as the inverse of the measurement bandwidth, corresponding to the response to the Dirac pulse. Therefore the weighting functions of the r.m.s.-average detector are shown in Figure 4.8.31 for the shortest pulse widths allowed by the bandwidths specified.

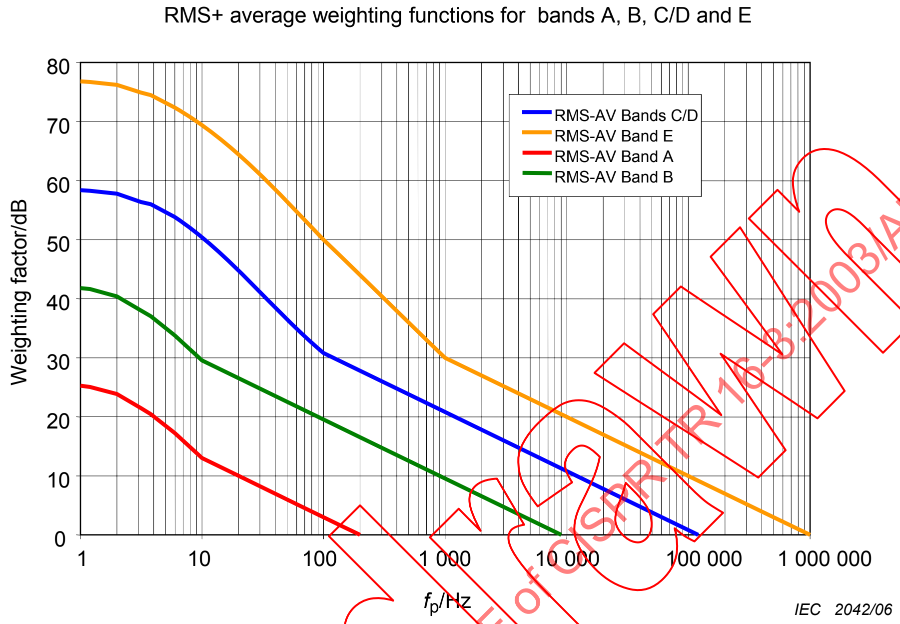


Figure 4.8.31 – RMS-average weighting functions for CISPR bands A, B, C/D and E for the shortest pulse widths allowed by the measurement bandwidths

Figure 4.8.31 shows theoretical weighting curves. In practice, the weighting factors apply up to values of approximately 40 dB for broadband emissions. If wider pulse widths, e.g. pulse-modulated carriers are measured, then the weighting function will change depending on the pulse width. If e.g. the pulse width is 10 times the shortest pulse width, this will shift the weighting curve by an amount of 10 dB, if the PRF is above the corner frequency, and by 20 dB, if PRF and reciprocal of pulse width are below the corner frequency.

If the r.m.s.-average weighting detector is used with a wider measurement bandwidth than the one specified, then the weighting curve will be shifted due to the shorter pulse width as shown in Figure 4.8.32.

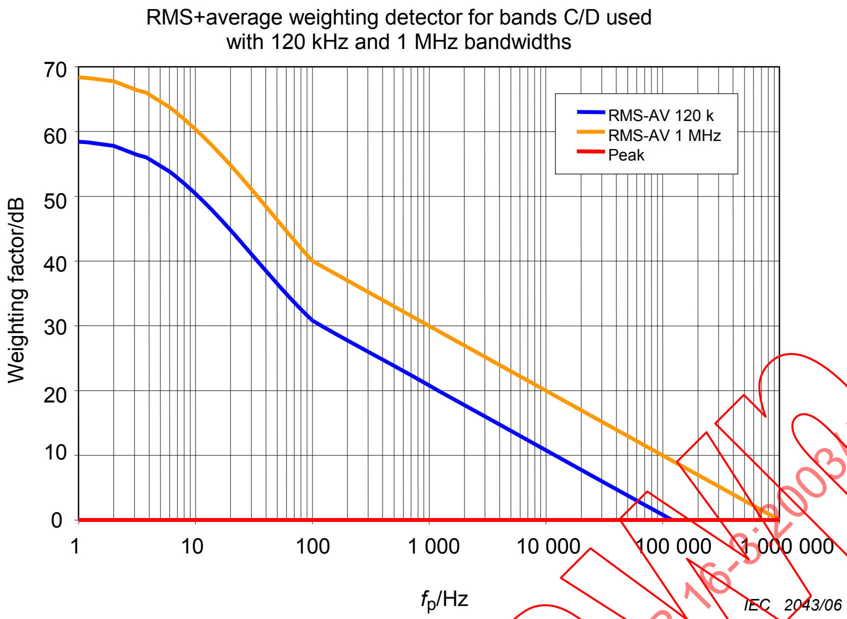


Figure 4.8.32 – Shift of the r.m.s.-average weighting function for CISPR band C/D by using a bandwidth of 1 MHz instead of 120 kHz, if the shortest possible pulse widths are applied

Measurement speeds: Measurement times and scan rates for the r.m.s.-average measuring receiver can be made similar to those of measuring receivers using the average detector (see CISPR 16-2-1, 16-2-2 and 16-2-3), i.e. measurements can be made substantially faster than with the quasi-peak measuring receiver. The definition of measurement times will have to take the characteristics of the disturbance into account.

The process of defining limits that are based on the r.m.s.-average detector has to take into account the effects of different types of disturbances on digital communication systems. For unmodulated sinewave signals, all detectors will yield the same result. For Gaussian noise, the r.m.s.-average measuring receiver will indicate a level approximately 1 dB higher than the average detector level, 6 dB lower than the quasi-peak detector (for bands C and D) level and 10 dB lower than the peak detector indication. Measurements of impulsive noise will result in levels between the average detector level and the quasi-peak detector or peak detector indications. In general, the r.m.s.-average detector is applicable for all types of continuous disturbance.

Table 4.8.7 shows examples of measurement results for some broadband disturbance sources, measured with the average, r.m.s.-average and quasi-peak detectors at frequencies in bands B and C. The measurements were taken in a small round-robin test, conducted in Germany in 2004.

Table 4.8.7: Measurement results for broadband disturbance sources (measurements with r.m.s.-average and quasi-peak detectors are normalized to average detector values)

EUT	*f* MHz	Average value dB(μV)	RMS-average minus average dB	Quasi-peak minus average dB
Hairdryer	1,0	32,75	+3,39	+11,81
Hairdryer	35	33,80	+8,49	+26,84
Washing machine 1	0,768	20,67	+4,74	+21,79
Washing machine 1	124	13,68	+3,80	+19,91
Washing machine 2	0,71	26,98	+1,71	+9,22
Washing machine 2	116	18,90	+3,92	+22,04

Taking into account that the r.m.s.-average measuring receiver addresses disturbance effects of all types of continuous emissions, it is possible to define one limit only, i.e., a single limit could be used and the limits for average and quasi-peak (or peak) detectors could be merged into one single limit, except for cases of discontinuous disturbances and disturbances like microwave oven emissions. However, this decision is ultimately to be made by product committees i.e., the committees responsible for the definition of emission limits. The application of the r.m.s.-average detector for measurements of discontinuous disturbances is to be discussed.

4.8.10 References

[1] CISPR publication 1, 2nd edition 1972, *Specification for C.I.S.P.R. radio interference measuring apparatus for the frequency range 0,15 MHz to 30 MHz.*

[2] CISPR/A/WG1(Hasenpusch/Dunker)04-01 Investigations of broadcast radio services in Germany

[3] CISPR/A/WG12(Ad hoc Weight Convenor)05-01: Possible Effects of Spread-Spectrum-Clock Interference on Wideband Radiocommunication Services

[4] STECHER, M. *Possible Effects of Spread-Spectrum-Clock Interference on Wideband Radiocommunication Services.* In Proc. IEEE EMC Symposium Chicago, USA, 2005.

[5] CISPR/A/WG1(Stecher-Stenumgaard)03-01 Weighting of interference according to its effect on digital communication services – results for convolutional codes.

[6] STENUMGAARD, P. *A Promising Solution of How to Evolve Present Radiated Emission Standards in Order to Protect Digital Communication Services.* Proc. IEEE EMC International Symposium on EMC, Istanbul, Turkey May 2003.

[7] CISPR/A/WG1(Stecher/Neibig)01-01: Weighting of Disturbance taking the Effect on Digital Radiocommunication Systems into account. Example: The Effect on GSM-900.

[8] Neibig, U. Störempfindlichkeit digitaler Mobilfunkempfänger am Beispiel von GSM-900" (Susceptibility of digital mobile receivers for the example of GSM 900). In *Frequenz*, 54 (2000), pp. 257 – 262.

[9] Measurement results published in CISPR/A/WG1(Stecher)00-1

[10] Measurement results published in CISPR/A/WG1(Stecher)03-1

[11] Measurement results published in CISPR/A/WG1(Stecher)04-1

[12] CISPR 16-1-1, *Specification for radio disturbance and immunity measuring apparatus and methods – Part 1-1: Radio disturbance and immunity measuring apparatus – Measuring apparatus*
Amendment 1[2)]

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2) To be published.

[13] CISPR 16-2-1, *Specification for radio disturbance and immunity measuring apparatus and methods – Part 2-1: Methods of measurement of disturbances and immunity – Conducted disturbance measurements*

[14] CISPR 16-2-2, *Specification for radio disturbance and immunity measuring apparatus and methods – Part 2-2: Methods of measurement of disturbances and immunity – Measurement of disturbance power*

[15] CISPR 16-2-3, *Specification for radio disturbance and immunity measuring apparatus and methods – Part 2-3: Methods of measurement of disturbances and immunity – Radiated disturbance measurements*

[16] ITU Recommendation BT.1368, *Planning criteria for digital terrestrial television services in the VHF/UHF bands*

4.9 Common mode absorption devices (CMAD)

4.9.1 Introduction

4.9.1.1 Purpose and application of CMAD

Common mode absorption devices (CMAD) are applied on cables leaving the test volume during radiated disturbance tests at a test site. The purpose of this part of CISPR 16 is to define the common mode (CM) impedance at the point where the cable leaves the test volume.

Figure 4.9.1 shows an example of an EUT in a radiated emission measurement test set-up for table-top equipment. The cables (e.g. the power supply cable, telecom cables, or other external connections used to exercise the EUT during the test) leave the test volume at the centre of the turntable. Radiation at frequencies between 30 MHz and about 200 MHz is from the cable acting as an antenna. The EUT together with the cable and the ground plane of the test site behave like a "top loaded monopole antenna." The radiation of such an antenna depends on the HF source (at the EUT connection of the cable), on the current distribution (CM current on the cable), and length of this unintended antenna. The current distribution depends upon the CM impedance at both ends of the antenna.

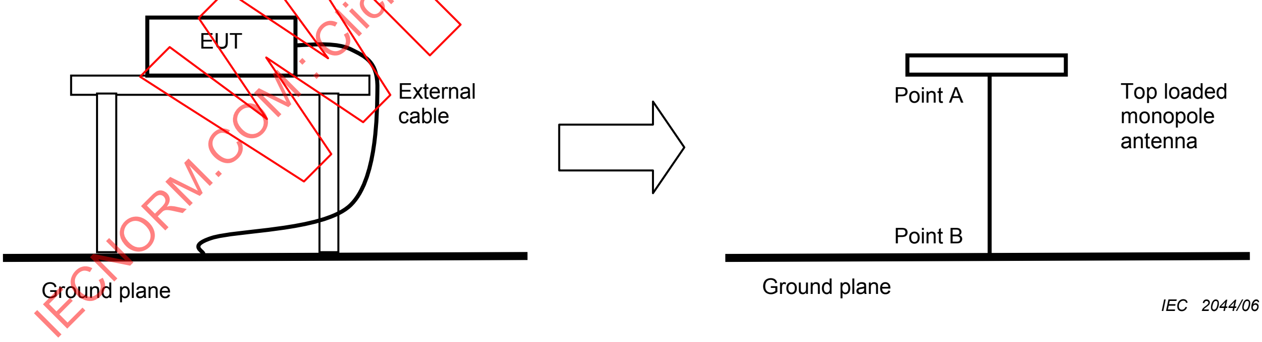

Figure 4.9.1 – Example of a simple EUT model

The CM impedance is unknown at both ends of the cable (points A and B in Figure 4.9.1). The CM impedance at point A is determined by the EUT and can have any value. For a given EUT, the value is fixed whenever the test is performed with an identical EUT. However, the CM impedance at point B ($Z_{apparent}$) may vary at each test laboratory, and can have any value because actual test procedures give no requirements for the CM impedance at this point. The actual value depends on the construction and layout of the test laboratory outside the test volume. Examples are given in [1].

It has been shown [1, 2] that the variation of the undefined CM impedance at point B can lead to variations as shown in Table 1 for radiated emissions measured from small EUTs. The sizes of these small EUTs were between 20 cm × 10 cm × 10 cm and 50 cm × 30 cm × 30 cm.

Table 1 – Expected deviations between different laboratories for small EUTs due to variations of the impedance $Z_{apparent}$ at point B

Frequency range	Possible maximum deviations of the radiated emission results between different laboratories
30 MHz – 50 MHz	10 dB – 25 dB
50 MHz – 120 MHz	5 dB – 15 dB
120 MHz – 200 MHz	2 dB – 7 dB

NOTE The variations of the cable layout in the test volume are not considered in this context. Table 1 does not include the variations of the radiation emission results due to variations in the cable layout.

The purpose of a CMAD is to reduce the influence of the CM impedance at point B upon the compliance uncertainty to a negligible amount.

4.9.1.2 Important properties of CMADs

The main purpose of a CMAD is to ensure that the CM impedance $Z_{apparent}$ at the point B of Figure 4.9.1 is always the same, independent of the undefined impedance at the cable entrance to the test volume in the different laboratories. Therefore the following two properties are important:

- The cable including the CMAD should have a CM impedance $Z_{apparent}$ (or $S_{11apparent}$) within a specified tolerance.
- The CMAD impedance $Z_{apparent}$ (or $S_{11apparent}$) should be independent of the CM impedance at the other end of the CMAD.

An additional purpose of a CMAD can be to attenuate disturbance signals not produced by the EUT, in order to distinguish between the EUT as a disturbance source and other disturbance sources. For this purpose, the insertion loss A_{IL} of the CMAD can be used as a figure-of-merit.

NOTE 1 The insertion loss A_{IL} is comprised of two components:

a) loss due to dissipation inside the device, and

b) loss due to mismatch between CMAD and line.

If two CMADs are used in cascade, the resulting insertion loss in general is not the sum of the individual insertion losses.

NOTE 2 The primary function of the absorbing clamp described in Clause 4 and Annex B of CISPR 16-1-3, is for the measurement of interference power. Depending on the ferrite material used, some types of absorbing clamps are suitable as CMAD.

NOTE 3 The EM clamp defined in IEC 61000-4-6 for RF-injection immunity testing is not suitable for use as a CMAD as described herein.

4.9.2 CMAD as a two-port device

4.9.2.1 Simple model of a CMAD

Usually CMADs are constructed using multiple ferrite clamps. Ferrite clamps have the advantage of being applicable to any types of cables, within a range of diameters. For measurements of the CMAD characteristics, the test cable is replaced by a well-defined test conductor. In CISPR 16-1-4, Clause 9, a test conductor of 4 mm diameter is defined, located above a ground plane at the height defined by the dimensions of the CMAD (typically 30 mm). The CMAD (ferrite clamp) along with the test conductor above the ground plane is regarded as a two-port device – see Figure 4.9.2.

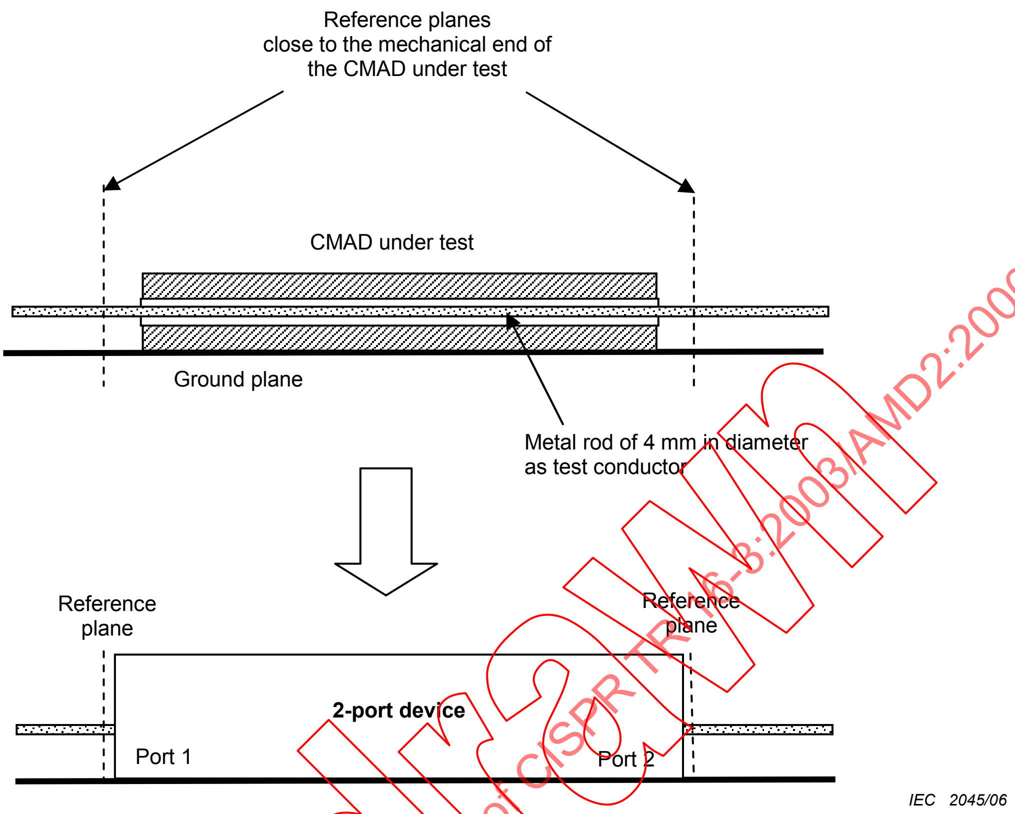

Figure 4.9.2 – Representation of a CMAD as a two-port device

A two-port device is completely described by the *S*-parameters measured at ports 1 and 2. The *S*-parameters are referenced to the characteristic impedance, Z_{ref}, of the test conductor above the ground plane without the CMAD in place, given by

$$Z_{ref} = \frac{Z_0}{2\pi} \cosh^{-1}\left(\frac{2h}{d}\right) \quad \text{in } \Omega$$

where

Z_0 is the free-space impedance (120π) in Ω;

d is the test conductor diameter (defined to be 4 mm);

h is the height of the centre of the test conductor above the ground plane.

EXAMPLE Typical values of Z_{ref} for various heights *h* are:

h = 30 mm	>>	Z_{ref} = 204 Ω
h = 65 mm	>>	Z_{ref} = 248 Ω
h = 90 mm	>>	Z_{ref} = 270 Ω

Any two-port network may be represented using various sets of parameters; each of these gives a complete characterisation of the two-port device. Examples of two-port parameter sets are:

- S_{11}, S_{21}, S_{12} and S_{22} – *S*-parameters: four complex numbers, related to a reference impedance Z_{ref} ;
- *A*, *B*, *C*, *D* (*ABCD* matrix: 4 complex numbers);
- other types of two-port parameter representations are described in the literature, but do not offer any advantages in the present context.

4.9.2.2 Parameters of a CMAD represented as a two-port device

The performance of a CMAD can basically be defined by the four complex *S*-parameters when measured as a two-port device in a test jig. The test conductor in the test jig has a diameter of 4 mm. The height above the ground plane, *h*, is defined by the dimensions of the CMAD. These two parameters define the reference impedance, Z_{ref}, for the *S*-parameter measurements. If the CMAD is symmetrical, S_{11} and S_{22} have the same value. If the device is not symmetrical, the test report must describe which port was used for the S_{11} test (the end closed to the EUT to be used for radiated emissions measurements), or the results must be reported for both ports of the CMAD.

4.9.2.3 Conversion between S-parameters and ABCD-parameters for a two-port network element

The conversion from *S*-parameters to *ABCD*-matrix representation is given by the following equations (Z_{ref} is the reference impedance to which the *S*-parameters are referred):

$$A = \frac{(1+S_{11})(1-S_{22})+S_{12}S_{21}}{2S_{21}} \qquad (4.9.1)$$

$$B = \frac{(1+S_{11})(1+S_{22})-S_{12}S_{21}}{2S_{21}} Z_{ref} \qquad (4.9.2)$$

$$C = \frac{(1-S_{11})(1-S_{22})-S_{12}S_{21}}{2S_{21}} \Big/ Z_{ref} \qquad (4.9.3)$$

$$D = \frac{(1-S_{11})(1+S_{22})+S_{12}S_{21}}{2S_{21}} \qquad (4.9.4)$$

The inverse equations are:

$$S_{21} = \frac{2}{A+B'+C'+D} \qquad (4.9.5)$$

$$S_{11} = \frac{A+B'-C'-D}{A+B'+C'+D} \qquad (4.9.6)$$

$$S_{12} = \frac{2(A\cdot D-B\cdot C)}{A+B'+C'+D} \qquad (4.9.7)$$

$$S_{22} = \frac{-A+B'-C'+D}{A+B'+C'+D} \qquad (4.9.8)$$

where

$$B' = B/Z_{ref} \qquad (4.9.9)$$

$$C' = C\cdot Z_{ref} \qquad (4.9.10)$$

NOTE All operations in preceding equations are for complex numbers. All parameters are functions of frequency. The equations are valid at each frequency point.

4.9.2.4 Range of variations for S11 due to undefined impedance at the far end of a CMAD

The apparent impedance of a two-port network element characterized by its *ABCD*-parameters is given by:

$$Z_{\text{apparent}} = \frac{A \cdot Z_{\text{end}} + B}{C \cdot Z_{\text{end}} + D}$$

From this equation the S_{11} parameter can be calculated using:

$$S_{11\text{apparent}} = \frac{Z_{\text{apparent}} - Z_0}{Z_{\text{apparent}} + Z_0} = \frac{(A - C \cdot Z_0)Z_{\text{end}} + (B - D \cdot Z_0)}{(A + C \cdot Z_0)Z_{\text{end}} + (B + D \cdot Z_0)}$$

Z_{apparent} and $S_{11\text{apparent}}$ are the values seen at port 1 if port 2 is connected to an impedance of Z_{end}.

Both quantities Z_{apparent} and $S_{11\text{apparent}}$ are a conformal mapping of Z_{end}, expressed as:

$$f(Z_{\text{end}}) = \frac{a \cdot Z_{\text{end}} + b}{c \cdot Z_{\text{end}} + d}$$

The general form of the equation for this type of conformal mapping is:

$$f(z) = \frac{a \cdot z + b}{c \cdot z + d}$$

This type of function has the property that it transforms straight lines and circles in the *z*-plane into either straight lines or circles in the *f*-plane. In particular, if the values of *z* are restricted to positive real values, the transformation of this half plane results in a circle in the *f*-plane, as shown in Figure 4.9.3.

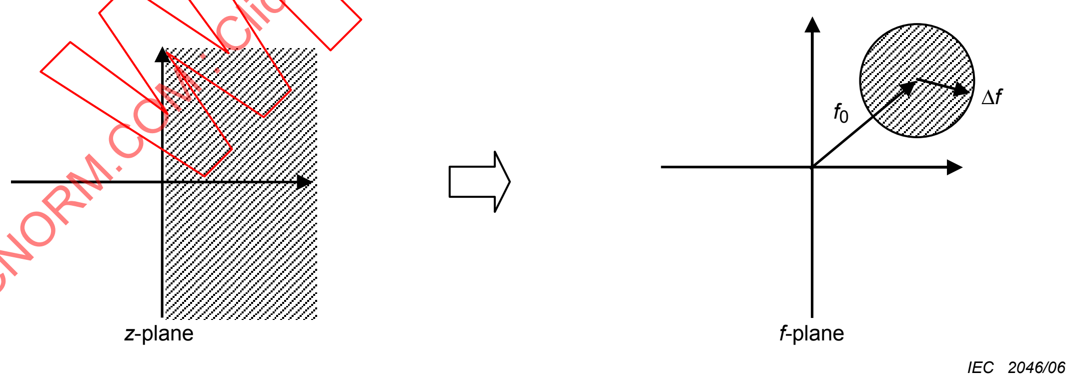

Figure 4.9.3 – Conformal mapping between *z*-plane and *f*-plane

The centre of this circle is at:

$$f_0 = \frac{b \cdot c - a \cdot d}{2c^2 \operatorname{Re}(d/c)} + \frac{a}{c} \qquad \text{(complex value)}$$

The radius of this circle is:

$$|\Delta f| = \left| \frac{b \cdot c - a \cdot d}{2c^2 \operatorname{Re}(d / c)} \right| \qquad \text{(scalar value)}$$

The maximum value of $|f|$ is then:

$$|f|_{\text{max}} = |f_0| + |\Delta f| \qquad \text{(scalar value)}$$

The minimum value of $|f|$ is then:

$$|f|_{\text{min}} = |f_0| - |\Delta f| \qquad \text{if } |f_0| > |\Delta f| \qquad \text{else} \qquad |f|_{\text{min}} = 0$$

Using these relations for Z_{apparent} gives the following parameters:

Position of the centre of the circle:

$$Z_{\text{apparent/center}} = \frac{B \cdot C - A \cdot D}{2C^2 \operatorname{Re}(D / C)} + \frac{A}{C} \qquad \text{(complex value)}$$

Radius of the circle:

$$|\Delta Z_{\text{apparent}}| = \frac{B \cdot C - A \cdot D}{2C^2 \operatorname{Re}(D / C)} \qquad \text{(scalar value)}$$

Maximum value of Z_{apparent}:

$$|Z_{\text{apparent}}|_{\text{max}} = |Z_{\text{apparent/center}}| + |\Delta Z_{\text{apparent}}|$$

Minimum value of Z_{apparent}:

$$|Z_{\text{apparent}}|_{\text{min}} = |Z_{\text{apparent/center}}| - |\Delta Z_{\text{apparent}}| \quad \text{if} \quad |Z_{\text{apparent/center}}| > |\Delta Z_{\text{apparent}}| \quad \text{else}$$

$$|Z_{\text{apparent}}|_{\text{min}} = 0$$

For S_{11} the relevant parameters are given by:

Position of the centre of the circle:

$$S_{11/\text{center}} = \frac{(B \cdot C - A \cdot D) \cdot Z_0}{(A + C \cdot Z_0)^2 \operatorname{Re}\left(\frac{B + D \cdot Z_0}{A + C \cdot Z_0} \right)} + \frac{A - C \cdot Z_0}{A + C \cdot Z_0}$$

Radius of the circle:

$$|\Delta S_{11}| = \left| \frac{(B \cdot C - A \cdot D) \cdot Z_0}{(A + C \cdot Z_0)^2 \,\mathrm{Re}\left(\dfrac{B + D \cdot Z_0}{A + C \cdot Z_0}\right)} \right|$$

Maximum value of $|S_{11\mathrm{apparent}}|$:

$$|S_{11\mathrm{apparent}}|_{\max} = |S_{11/\mathrm{center}}| + |\Delta S_{11}|$$

Minimum value of $|S_{11\mathrm{apparent}}|$:

$$|S_{11\mathrm{apparent}}|_{\min} = |S_{11/\mathrm{center}}| - |\Delta S_{11}| \quad \text{if} \quad |S_{11/\mathrm{center}}| > |\Delta S_{11}| \quad \text{else} \quad |S_{11\mathrm{apparent}}|_{min} = 0$$

4.9.3 Measurement of CMAD

4.9.3.1 Introduction

The CMAD parameters are defined as parameters of the two-port network measured at the reference planes with the reference impedance Z_{ref} given by the dimensions of the test jig cross section at the reference plane. Vector network analysers (VNA) used to measure the *S*-parameters operate with coaxial connectors having 50 Ω characteristic impedance. Between this 50 Ω coaxial connection and the non-coaxial configuration of the two-port device to be measured, an adaptor is needed to convert the 50 Ω coaxial connector to the geometry of the two-port device to be measured. Figure 4.9.4 illustrates the relevant set-up.

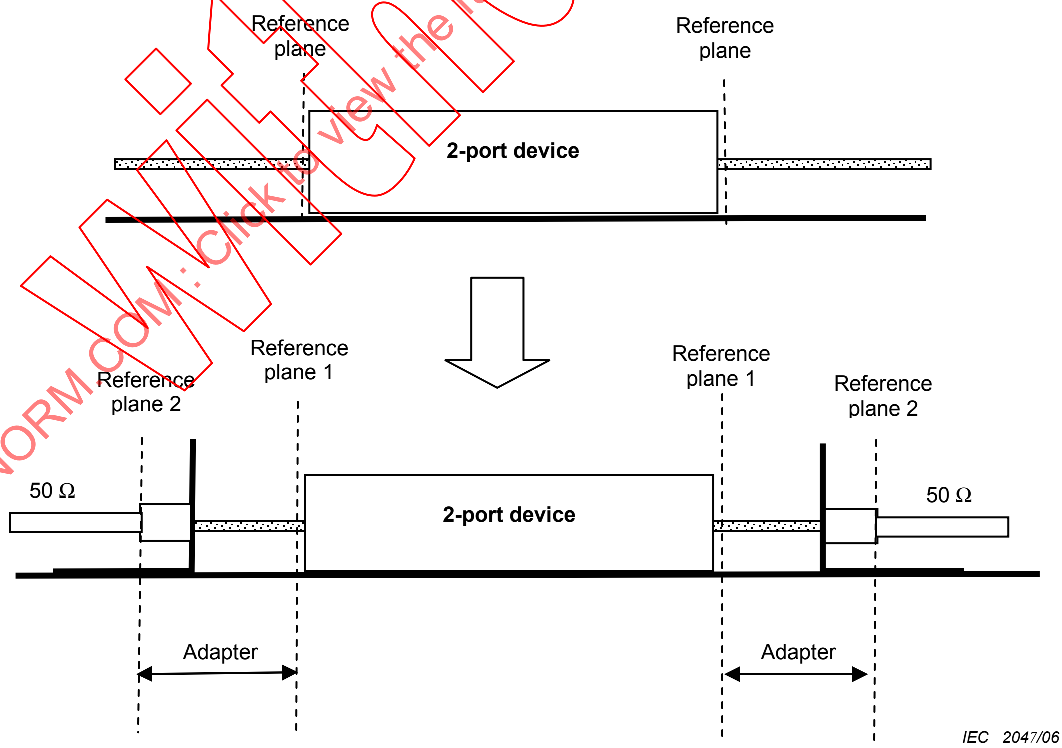

Figure 4.9.4 – Conversion from 50 Ω coaxial system to the geometry of the two-port device-under-test

If the TRL (thru-reflect-line) calibration method is used, it is possible to define calibration at reference plane 1 of Figure 4.9.4. The measurement result is directly referred to the connections of the two-port device at reference plane 1, and the measurement result does not include the adaptors. Measurements based on TRL calibrations are therefore recommended for accurate measurements of CMAD characteristics. The details of the TRL calibration method are described in 3.2.

The classical SOLT (short-open-load-thru) calibration method for a VNA is made at the reference plane 2 of Figure 4.9.4, for which the necessary calibration kits are commercially available. If this calibration at reference plane 2 is used, the properties of the adaptors are included in the measurement result.

The effect of the adaptors can be compensated partially using other "simplified" methods – two alternative methods are described in 3.3 and 3.4:

a) Measurement with SOLT calibration and position shifting (matching adaptors)
b) Measurement with SOLT calibration and transformation to Z_{ref} (lossless 50 Ω adaptors)

4.9.3.2 Measurement with TRL calibration method

4.9.3.2.1 Introduction

The TRL calibration method is based on the model shown in Figure 4.9.5 [3, 4].

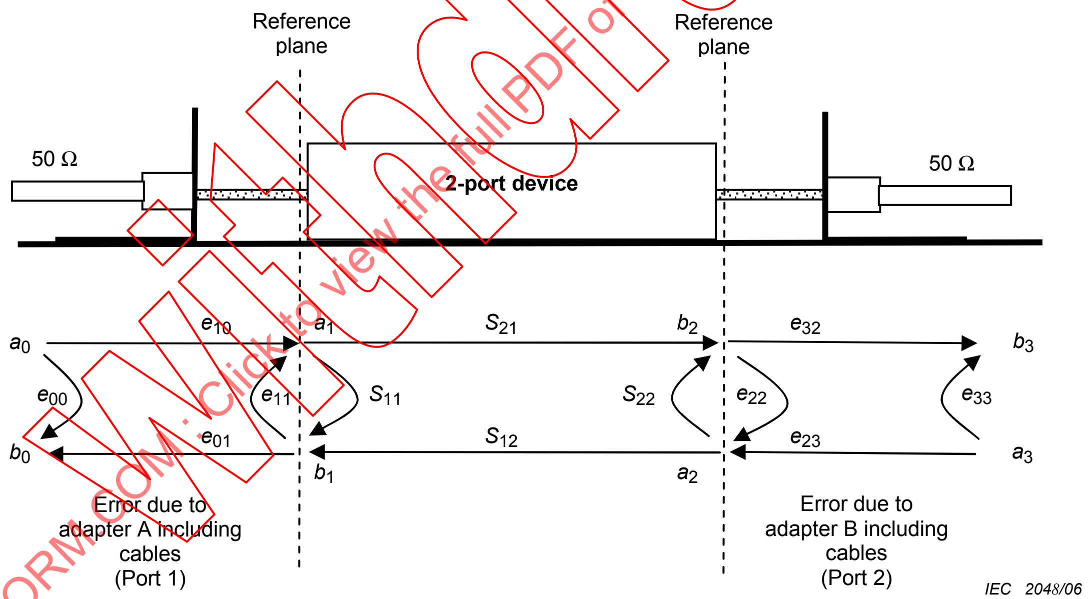

Figure 4.9.5 – Basic model for the TRL calibration

The four *S*-parameters are the true values of the measured two-port device. The parameters of the two adapters A and B are unknown and need to be derived from the calibration measurements with different external connections.

Four calibration configurations are necessary for the TRL calibration:

a) "Reflect" (Port 1): Measuring the complex value S_{11} of the adaptor section and adapter at port 1 without any other connection (simulating an open) – see Figure 4.9.6 a)
b) "Reflect" (Port 2): Measuring the complex value S_{22} of the adapter section and adapter at port 2 without any other connection (simulating an open) – see Figure 4.9.6 b)

c) "Through": Measuring the complex values S_{11}, S_{12}, S_{21}, S_{22} with the two adapter sections directly connected together (without the transmission line section in between) – see Figure 4.9.6 c)

d) "Line": Measuring the complex values S_{11}, S_{12}, S_{21}, S_{22} with the transmission line section introduced – see Figure 4.9.6 d).

The results of these calibration measurements are 10 complex numbers for each frequency. Many VNAs have firmware for the TRL calibration included. If the VNA includes firmware for TRL calibration, it will use these reference measurements to calculate the proper corrections for the TRL measurement. If the VNA does not support the TRL calibration, the necessary corrections can be made externally according to the procedure described below (4.9.3.2.2 to 4.9.3.2.5).

The characteristic impedance of the "line" section has to be known exactly and is introduced into the calibration data used by the firmware of the VNA. Some firmware also asks for the electrical length of the "line" section, but theoretically only the impedance is needed. The properties of the adaptor section and adaptors outside the calibration plane do not need to be known for the TRL calibration. These properties are measured in the calibration procedure and are compensated directly by the TRL calibration. Any type of adaptors may be used.