

TECHNICAL REPORT

**CISPR
16-3**

Second edition
2003-11

INTERNATIONAL SPECIAL COMMITTEE ON RADIO INTERFERENCE

**Specification for radio disturbance and immunity
measuring apparatus and methods –**

**Part 3:
CISPR technical reports**

Withdrawn



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Part 3: CISPR technical reports

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WORLDWIDE

INTERNATIONAL ELECTROTECHNICAL COMMISSION

**SPECIFICATION FOR RADIO DISTURBANCE
AND IMMUNITY MEASURING APPARATUS AND METHODS –****Part 3: CISPR technical reports****FOREWORD**

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CISPR 16-3, which is a technical report, has been prepared by CISPR subcommittee A: Radio interference measurements and statistical methods.

This second edition of CISPR 16-3, together with CISPR 16-4-1, CISPR 16-4-3 and CISPR 16-4-4 cancels and replaces the first edition of CISPR 16-3, published in 2000, and its amendment 1 (2002). It contains the relevant clauses of CISPR 16-3 without technical changes.

This publication has been drafted in accordance with the ISO/IEC Directives, Part 2.

A bilingual version of this publication may be issued at a later date.

The committee has decided that the contents of this publication will remain unchanged until 2006. At this date, the publication will be

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- withdrawn;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

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

Report 33 – p/o CISPR 8, 1969; Report 38 – p/o CISPR 8, 1969; Report 49 – p/o CISPR 8C, 1980; Report: CIS/A(Sec)67 + CIS/A(Sweden)29; RM 2828/CISPR/A, 1985; CIS/A(CO)32, 1985; CIS/A(Sec)58, 1983; CIS/A(Sec)58A, 1983; CIS/A(Sec)67, 1985; CIS/A(CO)77A, 1993; CIS/A(CO)81, 1987; CIS/A(CO)82, 1994; CIS/A(CO)84, 1994; CIS/A(Sec)84, 1987; CIS/A(Sec)88, 1988; CIS/A(Sec)88A, 1988; CIS/A(Sec)94, 1989; CIS/A(Sec)115, 1991; CIS/A(Sec)115A, 1991; CIS/A(Sec)116, 1991; CIS/A(Sec)124, 1991; CIS/A(Sec)128, 1992; CIS/A(Sec)132, 1993; CIS/A/166/CD, 1995.

Withdrawn

INTRODUCTION

CISPR 16-1, CISPR 16-2, CISPR 16-3 and CISPR 16-4 have been reorganised into 14 parts, to accommodate growth and easier maintenance. The new parts have also been renumbered. See the list given below.

Old CISPR 16 publications		New CISPR 16 publications	
CISPR 16-1	Radio disturbance and immunity measuring apparatus	CISPR 16-1-1	Measuring apparatus
		CISPR 16-1-2	Ancillary equipment – Conducted disturbances
		CISPR 16-1-3	Ancillary equipment – Disturbance power
		CISPR 16-1-4	Ancillary equipment – Radiated disturbances
		CISPR 16-1-5	Antenna calibration test sites for 30 MHz to 1 000 MHz
CISPR 16-2	Methods of measurement of disturbances and immunity	CISPR 16-2-1	Conducted disturbance measurements
		CISPR 16-2-2	Measurement of disturbance power
		CISPR 16-2-3	Radiated disturbance measurements
		CISPR 16-2-4	Immunity measurements
CISPR 16-3	Reports and recommendations of CISPR	CISPR 16-3	CISPR technical reports
		CISPR 16-4-1	Uncertainties in standardised EMC tests
		CISPR 16-4-2	Measurement instrumentation uncertainty
		CISPR 16-4-3	Statistical considerations in the determination of EMC compliance of mass-produced products
CISPR 16-4	Uncertainty in EMC measurements	CISPR 16-4-4	Statistics of complaints and a model for the calculation of limits

More specific information on the relation between the 'old' CISPR 16-3 and the present 'new' CISPR 16-3 is given in the table after this introduction (TABLE RECAPITULATING CROSS REFERENCES).

Measurement instrumentation specifications are given in five new parts of CISPR 16-1, while the methods of measurement are covered now in four new parts of CISPR 16-2. Various reports with further information and background on CISPR and radio disturbances in general are given in CISPR 16-3. CISPR 16-4 contains information related to uncertainties, statistics and limit modelling.

TABLE RECAPITULATING CROSS-REFERENCES

First edition of CISPR 16-3
Clauses, subclauses

Second edition of CISPR 16-3
Clauses, subclauses

1.1
1.2
1.3

1
2
3

4
4.1
4.2
4.3
4.4
4.5
4.6

4
4.1
4.2
4.3
4.4
4.5
4.6

5
5.1
5.2

5
5.1
5.2

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SPECIFICATION FOR RADIO DISTURBANCE AND IMMUNITY MEASURING APPARATUS AND METHODS –

Part 3: CISPR technical reports

1 Scope

This part of CISPR 16 contains specific technical reports and information on the history of CISPR.

Over the years, the CISPR prepared a number of recommendations and reports that have significant technical merit but were not generally available. Reports and recommendations were for some time published in CISPR 7 and 8.

At its meeting in Campinas, Brazil, in 1988, subcommittee A agreed on the table of contents of part 3 and to publish the reports for posterity by giving the reports a permanent place in part 3.

With the reorganization of CISPR 16 in 2003, the significance of CISPR limits has been moved to CISPR 16-4-3, whereas recommendations on statistics of disturbance complaints and on the report on the determination of limits has been moved to CISPR 16-4-4. The contents of Amendment 1 (2002) has been moved to CISPR 16-4-1.

2 Normative references

The following referenced documents are indispensable for the application of this document. For dated references, only the edition cited applies. For undated references, the latest edition of the referenced document (including any amendments) applies.

CISPR 11, *Industrial, scientific and medical (ISM) radio-frequency equipment – Electro-magnetic disturbance characteristics – Limits and methods of measurement*

CISPR 16-1 (all parts), *Specification for radio disturbance and immunity measuring apparatus and methods – Radio disturbance and immunity measuring apparatus*

CISPR 16-2 (all parts), *Specification for radio disturbance and immunity measuring apparatus and methods – Methods of measurement of disturbances and immunity*

CISPR 16-4 (all parts), *Specification for radio disturbance and immunity measuring apparatus and methods – Uncertainties, statistics and limit modelling*

ITU-R Recommendation BS. 468-4, *Measurement of audio-frequency noise voltage level in sound broadcasting*

3 Definitions

For the purpose of this part of CISPR 16, the definitions of CISPR 16-1 and IEC 60050(161) as well as the following definitions apply.

3.1

bandwidth (B_n)

width of the overall selectivity curve of the receiver between two points at a stated attenuation, below the midband response. The bandwidth is represented by the symbol B_n , where n is the stated attenuation in decibels

3.2 impulse bandwidth (B_{imp})

$$B_{\text{imp}} = A(t)_{\text{max}} / (2G_o \times IS)$$

where

$A(t)_{\text{max}}$ is the peak of the envelope at the IF output of the receiver with an impulse area IS applied at the receiver input;

G_o is the gain of the circuit at the centre frequency.

Specifically, for two critically coupled tuned transformers,

$$B_{\text{imp}} = 1,05 \times B_6 = 1,31 \times B_3$$

where B_6 and B_3 are respectively the bandwidths at the –6 dB and –3 dB points (see CISPR 16-1-1 for further information)

3.3 impulse area (sometimes called impulse strength) (IS) the voltage-time area of a pulse defined by the integral:

$$IS = \int_{-\infty}^{+\infty} V(t)dt \text{ (expressed in } \mu\text{Vs or dB}(\mu\text{Vs))}$$

NOTE Spectral density (D) is related to impulse area and expressed in $\mu\text{V/MHz}$ or $\text{dB}(\mu\text{V})/\text{MHz}$. For rectangular impulses of pulse duration T at frequencies $f \ll 1/T$, the relationship $D (\mu\text{V/MHz}) = 2 \times 10^6 / IS (\mu\text{Vs})$ applies since D is calibrated in r.m.s. values of a corresponding sine wave.

3.4 electrical charge time constant (T_c)

time needed after the instantaneous application of a constant sine-wave voltage to the stage immediately preceding the input of the detector for the output voltage of the detector to reach 63 % of its final value

NOTE This time constant is determined as follows. A sine-wave signal of constant amplitude and having a frequency equal to the mid-band frequency of the i.f. amplifier is applied to the input of the stage immediately preceding the detector. The indication, D , of an instrument having no inertia (for example, a cathode-ray oscilloscope) connected to a terminal in the d.c. amplifier circuit so as not to affect the behaviour of the detector, is noted. The level of the signal is chosen such that the response of the stages concerned remains within the linear operating range. A sine-wave signal of this level, applied for a limited time only and having a wave train of rectangular envelope is gated such that the deflection registered is $0,63D$. The duration of this signal is equal to the charge time of the detector.

3.5 electrical discharge time constant (T_D)

time needed after the instantaneous removal of a constant sine-wave voltage applied to the stage immediately preceding the input of the detector for the output of the detector to fall to 37 % of its initial value

NOTE The method of measurement is analogous to that for the charge time constant, but instead of a signal being applied for a limited time, the signal is interrupted for a definite time. The time taken for the deflection to fall to $0,37D$ is the discharge time constant of the detector.

3.6 mechanical time constant (T_M) of a critically damped indicating instrument

$$T_M = T_L / 2\pi$$

where T_L is the period of free oscillation of the instrument with all damping removed.

NOTE 1 For a critically damped instrument, the equation of motion of the system may be written as

$$T_M^2(d^2\alpha / dt^2) + 2T_M(d\alpha / dt) + \alpha = ki$$

where

α is the deflection;

i is the current through the instrument;

k is a constant.

It can be deduced from this relation that this time constant is also equal to the duration of a rectangular pulse (of constant amplitude) that produces a deflection equal to 35 % of the steady deflection produced by a continuous current having the same amplitude as that of the rectangular pulse.

NOTE 2 The methods of measurement and adjustment are deduced from one of the following:

- The period of free oscillation having been adjusted to $2\pi T_M$, damping is added so that $\alpha_{TM} = 0,35 \alpha_{max}$.
- When the period of oscillation cannot be measured, the damping is adjusted to be just below critical such that the overshoot is not greater than 5 % and the moment of inertia of the movement is such that $\alpha_{TM} = 0,35 \alpha_{max}$.

3.7

overload factor

ratio of the level that corresponds to the range of practical linear function of a circuit (or a group of circuits) to the level that corresponds to full-scale deflection of the indicating instrument.

The maximum level at which the steady-state response of a circuit (or group of circuits) does not depart by more than 1 dB from ideal linearity defines the range of practical linear function of the circuit (or group of circuits).

3.8

symmetric voltage

in a two-wire circuit, such as a single-phase mains supply, the symmetric voltage is the radio-frequency disturbance voltage appearing between the two wires. This is sometimes called the differential mode voltage. If V_a is the vector voltage between one of the mains terminals and earth and V_b is the vector voltage between the other mains terminal and earth, the symmetric voltage is the vector difference ($V_a - V_b$).

3.9

asymmetric voltage

radio-frequency disturbance voltage appearing between the electrical mid-point of the mains terminals and earth. It is sometimes called the common-mode voltage and is half the vector sum of V_a and V_b , i.e. $(V_a + V_b)/2$.

3.10

unsymmetric voltage

amplitude of the vector voltage, V_a or V_b defined in 3.8 and 3.9. This is the voltage measured by the use of an artificial mains V-network

3.11

CISPR indicating range

range specified by the manufacturer which gives the maximum and the minimum meter indications within which the receiver meets the requirements of CISPR 16-1-1.

4 Technical reports

4.1 Correlation between measurements made with apparatus having characteristics differing from the CISPR characteristics and measurements made with CISPR apparatus

4.1.1 Introduction

CISPR standards for instrumentation and methods of measurement have been established to provide a common basis for controlling radio interference from electrical and electronic equipment in international trade.

The basis for establishing limits is that of providing a reasonably good correlation between measured values of the interference and the degradation it produces in a given communications system. The acceptable value of signal-to-noise ratio in any given communication system is a function of its parameters including bandwidth, type of modulation and other design factors. As a consequence, various types of measurements are used in the laboratory in research and development work in order to carry out the required investigations.

The purpose of this subsection is to analyse the dependence of the measured values on the parameters of the measuring equipment and on the waveform of the measured interference.

4.1.2 Critical interference measuring instrument parameters

The most critical factors in determining the response of an instrument for measuring interference are the following: the bandwidth, the detector, and the type of interference being measured. Considered to be of secondary importance, but, nevertheless, quite significant in correlating instruments under particular circumstances, are: overload factor, AGC design (if used), image and other spurious responses, and meter time constant and damping.

For purposes of discussion, reference is made to three fundamental types of radio noise: impulse, random and sine wave. The dependence of the response to each of these on the bandwidth and the type of detector is given in table 4.1-1. In this table, δ is the magnitude of the impulse strength, Δf_{imp} is the impulse bandwidth, Δf_{rn} is the random noise bandwidth, $P(\alpha)$ is the pulse response for the quasi-peak detector, f_{PR} is the pulse repetition frequency, and E' is the spectral amplitude of the random noise. The relative responses of various detectors to impulse interference for one instrument are shown in figure 4.1-1.

Table 4.1-1 shows that the dependence of the noise meter response on bandwidth is different for all three types of interference. If the waveform being measured can be defined as being any of the three types listed in table 4.1-1, and if a standard source provides that type of waveform, then by using the substitution method, a satisfactory calibration can be obtained for any instrument with adequate overload factor independent of its bandwidth. Thus, with a purely random interference or a purely impulsive interference of known repetition rate, calibration can be made using a corresponding source, or a correlation factor calculated on the basis of known circuit parameters.

If a particular interference waveform is of a type intermediate between these three types, then the correction or correlation factors will also be intermediate. In any given case, it will be necessary to classify the noise waveform in such a manner that a significant correlation factor can be established. Hence, in order to develop this subject to any significant extent, it will be necessary to examine typical interference sources and to determine the extent to which they are of impulsive, random, or sine-wave type.

If an interference measuring set with several types of detectors is available, for example, peak, quasi-peak and average, the type of interference can be assessed by measuring the ratios of the readings obtained with these detectors. These ratios will, of course, depend

upon the bandwidth and other characteristics of the instrument being used for the measurement.

4.1.3 Impulse interference – Correlation factors

The quasi-peak detector response of any interference measuring set to regularly repeated impulses of uniform amplitude can be determined by the use of the "pulse response curve" which is shown in figure 4.1-2. This figure shows the response of the detector in percentage of peak response for any given bandwidth and value of charge resistance and discharge resistance. Applying this curve, it should be noted that the peak itself is dependent upon the bandwidth, so that as the bandwidth increases, peak value increases, but the percentage of peak, which is read by the detector, decreases; over a narrow range of bandwidth, these effects tend to counteract each other. The bandwidth used in this curve is the 6 dB bandwidth, which, for the passband characteristics typical of most interference measuring equipment, is about 5 % less than the so-called impulse bandwidth. A theoretical comparison of instruments having various bandwidths and detector parameters with the CISPR instrument is shown in figure 4.1-3.

The response of the average detector to impulsive noise is an interesting case. The reading of an average detector for impulsive noise is independent of the bandwidth of the pre-detector stages. It is, of course, directly proportional to the repetition rate. In most cases, the reading obtained with an average detector for impulsive noise is so low as to be of no practical value unless the noise meter bandwidth is exceedingly narrow, such as of the order of a few hundred hertz. For a repetition rate of 100 Hz and a bandwidth of the order of 10 kHz, the average value would be approximately 1 % of the peak value. Such a value is too low to measure with any degree of precision. Furthermore, for many communication systems, the annoyance effect may be well above the reading obtained with the average meter. This, of course, is one of the justifications for the use of the quasi-peak instrument.

4.1.4 Random noise

The response of a noise meter to random noise is proportional to the square root of the bandwidth. This result is independent of the type of detector used. The ratio of the random noise bandwidth to the 3 dB bandwidth is a function of the type of filter circuit. On the other hand, it has been shown that for many circuits typical of those used in interference measuring equipment, a ratio of effective random noise bandwidth to the 3 dB bandwidth of about 1,04 is a reasonable figure.

4.1.5 The r.m.s. detector

One of the advantages of the r.m.s. detector in correlation work is that for broadband noise the output obtained from it will be proportional to the square root of the bandwidth, i.e. the noise power is directly proportional to the bandwidth. This feature makes the r.m.s. detector particularly desirable and is one of the main reasons for adopting the r.m.s. detector to measure atmospheric noise. Another advantage is that the r.m.s. detector makes a correct addition of the noise power produced by different sources, for example, impulsive noise and random noise, thus for instance allowing a high degree of background noise.

The r.m.s. values of noise often give a good assessment of the subjective effect of interference to a.m. sound and television reception. However, the very wide dynamic range needed when using very wide-band instruments for measuring impulsive noise, limits the use of r.m.s. detectors to narrow-band instruments.

4.1.6 Discussion

The preceding paragraphs have indicated the theoretical basis for comparing measurements obtained with different instruments. As mentioned previously, the possibility of establishing significant correlation factors depends upon the extent to which noise can be classified and identified so that the proper correlation factors may be used. In many frequency ranges, impulsive interference appears to be the most serious; however, for

power lines where corona interference is the primary concern, random interference would be expected to be more characteristic. Additional quantitative data are needed on typical interference characteristics. Another important parameter is the overload factor.

4.1.7 Application to typical noise sources

Commutator motors

The noise generated by commutator motors is usually a combination of impulse and random noise. The random noise originates in the varying brush contact resistance, while the impulse noise is generated from the switching action at the commutator bars. For optimum adjustment of commutation the impulse noise can be minimized. However, where variable loading is possible, measurements have confirmed that for the peak and quasi-peak detectors, the dominant noise is of impulse type and the random component may be neglected. While the repetition rate may be of the order of 4 kHz, the effective rate is lower because the amplitude of the impulses is usually modulated at twice the line frequency. Hence, experimental results have shown that quasi-peak readings are consistent with bandwidth variations if the repetition rate of the impulse is assumed to be twice the line frequency.

Peak measurements show fluctuating levels on such noise because of the irregular nature of the commutator switching action.

The quasi-peak to average ratio is lower than would be obtained for pure impulse noise for two reasons.

- 1) The modulation of the commutator switching transients by line frequency produces many pulses below the measured quasi-peak level. These pulses do not contribute to the quasi-peak value but do contribute to the average.
- 2) The relatively low level, but continuous, random noise can likewise contribute substantially only to the average value. Experimental values of quasi-peak to average ratio ranged from 13 dB to 23 dB with the highest ratios for the widest bandwidths (120 kHz).

Impulsive sources

Tests on an ignition model, commutator motor appliances, and appliances using vibrating regulators showed reasonable agreement on instruments with the same nominal bandwidth, but with time constant ratios of the order of 3 to 1 on restricted portions of the output indicator scale. Deviations at higher scale values are without explanation. Relatively poor correlation was obtained on sources producing very low repetition rate pulses.

Ignition interference

CISPR Recommendation 35 recognizes that correlation between quasi-peak and peak detectors can be established as a practical matter. The conversion factor of 20 dB is explained partly on the basis of theory for uniform repeated impulses, and partly on the basis of the actual irregularity of the amplitude and wave shape of such impulses.

Noise from high-voltage lines

Comparative tests were made with an instrument meeting CISPR specification and one meeting those of the U.S.A. Standards Institute. The latter read 0 dB to 1 dB higher in one test and 1 dB to 3 dB higher in another (see IEEE Special Publication 31C44).

Dependence on bandwidth

Comparisons of measurements made in the U.K. with two instruments having bandwidths of 90 kHz and 9 kHz respectively have been reported to show that for most interference sources, the values obtained are in the ratio 14 dB to 18 dB. This figure is consistent with the concept that the interference is dominated by impulse type noise but that some random components are present.

4.1.8 Conclusions

Analysis of data comparing the responses of various instruments shows that it is possible to explain in almost every case the differences in measured values on the basis of theoretical and practical considerations. In many instances, it is indicated that waveform characteristics are known to predict correlation factors adequately with an accuracy of 2 dB to 4 dB.

Further studies are needed

- to characterize in some detail the waveforms of various sources of interference, and
- to correlate these waveform characteristics with measured values and the instrument parameters.

Table 4.1-1 – Comparative response of slideback peak, quasi-peak and average detectors to sine wave, periodic pulse and Gaussian waveform

Input waveform	Detector type			
	Slideback peak (sb)	Quasi-peak:1/600 (qp)	Field intensity (average)	RMS
CW sine wave	$e^\dagger$	e	e	e
Periodic pulse (no overlap)	$1,41 \delta \Delta f_{\text{imp}}$	$1,41 \delta \Delta f_{\text{imp}} P(\alpha)^*$	$1,41 \delta f_{\text{PR}}^+$	$1,41 \delta \sqrt{f_{\text{PR}} \Delta f_{\text{imp}}}$
Random ++	—	$1,85 \sqrt{\Delta f_{\text{m}} E'}^{**}$	$0,88 \sqrt{\Delta f_{\text{m}} E'}$	$\sqrt{\Delta f_{\text{m}} E'}$
† e is the r.m.s. value of the applied sine wave. * $P(\alpha)$ is given in figure 4.1-2. ** E' is spectral strength in r.m.s. volts/hertz bandwidth. + δ is impulse strength. It is assumed the instrument is calibrated in terms of the r.m.s. value of a sine wave. ++ It is assumed that characteristics of the envelope are measured by the detector on random noise.				

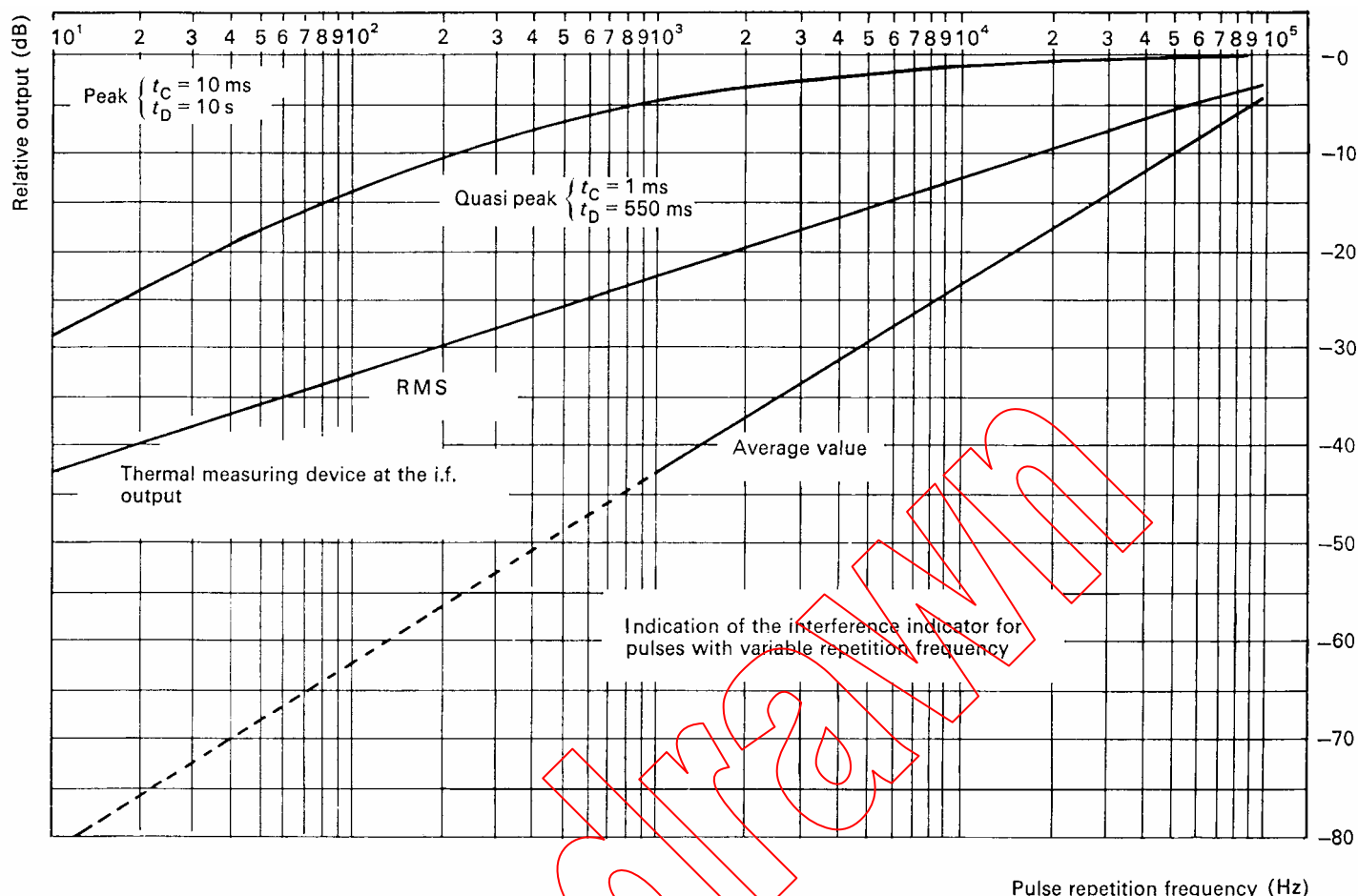
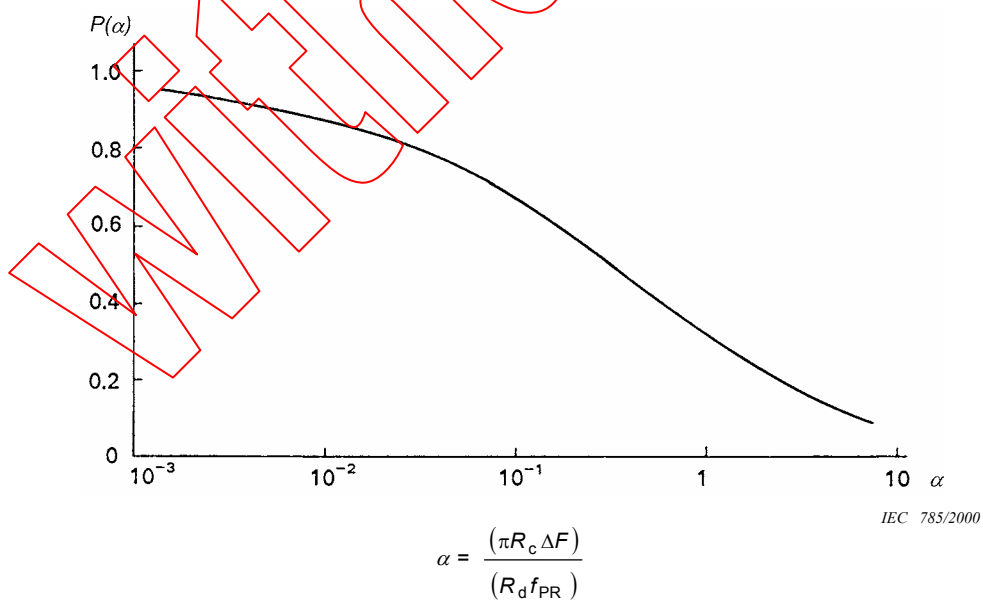


Figure 4.1-1 – Relative response of various detectors to impulse interference



R_c = charging resistance, in ohms

R_d = discharging resistance, in ohms

ΔF = 6 dB bandwidth, in hertz

f_{PR} = pulse repetition frequency, in hertz

Figure 4.1-2 – Pulse rectification coefficient $P(\alpha)$

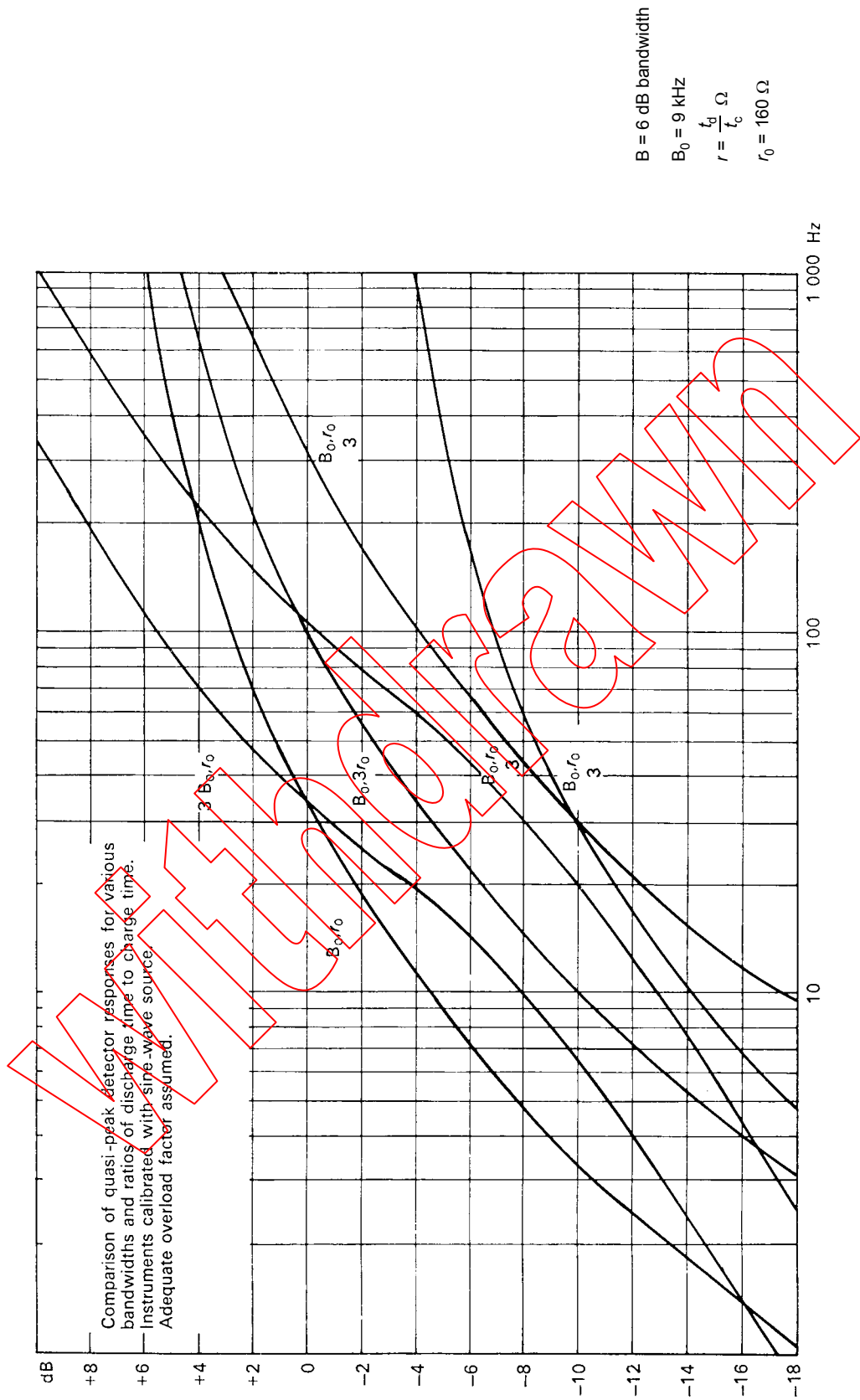


Figure 4.1-3 – Pulse repetition frequency

4.2 Interference simulators

4.2.1 Introduction

Interference simulators can be used for various applications, in particular to study signal processing in systems and equipment in the presence of interference (for example, overloading of receivers, synchronization of TV receivers, error rate of data signals, etc.) and for assessment of the annoyance caused by disturbances in broadcast and communication services.

A simulator should produce a stable and reproducible output signal, which requirement is normally not fulfilled by an actual interference source, and the simulator output waveform should show a good resemblance to the actual interference signal.

4.2.2 Types of interference signals

The following interference sources can be simulated.

- a) Narrowband interference sources generating sine-wave signals, for example receiver oscillators and ISM equipment. These sources can be simulated by an appropriate RF standard signal generator. ISM interference is often modulated by the mains voltage which can be simulated by modulating the RF signal with a full-wave rectified mains signal.
- b) Broadband interference sources producing continuous broadband noise, for example, gaseous discharges and corona. For simulating purposes a standard broadband noise source (saturated vacuum tube diode, zener diode or gas tube followed by a suitable broadband amplifier) can be used. In mains-fed sources of this type, mains modulation is present, but because of the non-linear behaviour of gaseous discharges the envelope of the actual noise signal can deviate appreciably from the normal full-wave rectified mains waveform. In this case, gating the noise of the simulator at a repetition frequency of twice the mains frequency can yield a good correspondence with the actual interference signal.
- c) Thyristor controlled regulators with phase control generate in a RF-channel narrow pulses of constant amplitude at a repetition frequency equal to twice the mains frequency. They can be simply simulated by standard pulse generators with narrow output pulses (10^{-7} s to 10^{-9} s width) of the same repetition frequency.
- d) Ignition systems, mechanical contacts and commutator motors generate short periods (bursts) of quasi-impulsive noise. This type of noise is caused by very short pulses of regular or irregular height at random time intervals; if the average interval between adjacent pulses is less than the reciprocal of the channel bandwidth under test ($\tau_{av} < 1/B$) the pulses overlap and because of the random phase conditions a random fluctuating output signal (noise) results. Therefore, bursts of quasi-impulsive interference of this type can be simulated by a gated broadband noise signal.

The duration and the repetition frequency of the bursts depend on the type of interference source (see table 4.2-1).

Ignition interference is characterized by burst durations between 20 μ s and 200 μ s and repetition frequencies between 30 bursts/s and 300 bursts/s depending on the number of cylinders and revolutions/minute of the engine.

Mechanical contacts produce bursts (clicks) which can vary between some milliseconds (snap-off switches) and more than 200 ms. In the case of a contact device in a mains-fed circuit, the noise during the burst is modulated with the full-wave rectified mains voltage.

Commutator motors produce much shorter bursts with durations between 20 μ s and 200 μ s at repetition frequencies between 10^3 bursts/s and 10^4 bursts/s, depending on the number of commutator bars and revolutions/minute of the rotor. Also in this case, mains supply causes a similar envelope modulation of the noise bursts.

4.2.3 Circuits for simulating broadband interference

Simulators of this type should generate gated noise bursts with or without mains modulation according to the characteristics laid down in table 4.2-1. Figure 4.2-1 shows a straightforward design with a noise source followed by an appropriate amplifier of 70 dB to 80 dB gain, a gating circuit to simulate the bursts, a mains envelope modulator and an output attenuator to adjust the required output level.

Table 4.2-1 – Characteristics of gate generator and modulator to simulate various types of broadband interference

Simulator signal	Burst duration	Burst repetition frequency	Mains modulation ^a
Gaseous discharge		Continuous	Yes/no ^b
Ignition	20-200 μ s	30-300 bursts/s	No
Switches	5-500 ms	0,2-30 bursts/min or single	Yes/no
Commutator motors	30-300 μ s	10^3 - 10^4 bursts/s	Yes/no

^a Depending on a.c. or d.c. supply.

^b In the case of mains modulation, gating at a repetition frequency $2f_{\text{mains}}$ and gate width of 1 ms to 2 ms may be more appropriate.

The disadvantage of this layout is that a wide usable frequency range requires a broad bandwidth for the entire circuit between noise source and output terminal. The most critical part in this respect is the high-gain amplifier. For applications in a wide frequency range (for example, 0 MHz to 1 000 MHz) such a range can be split up in several smaller ranges or a tunable amplifier may be used. Such a design complicates the construction of the simulator appreciably.

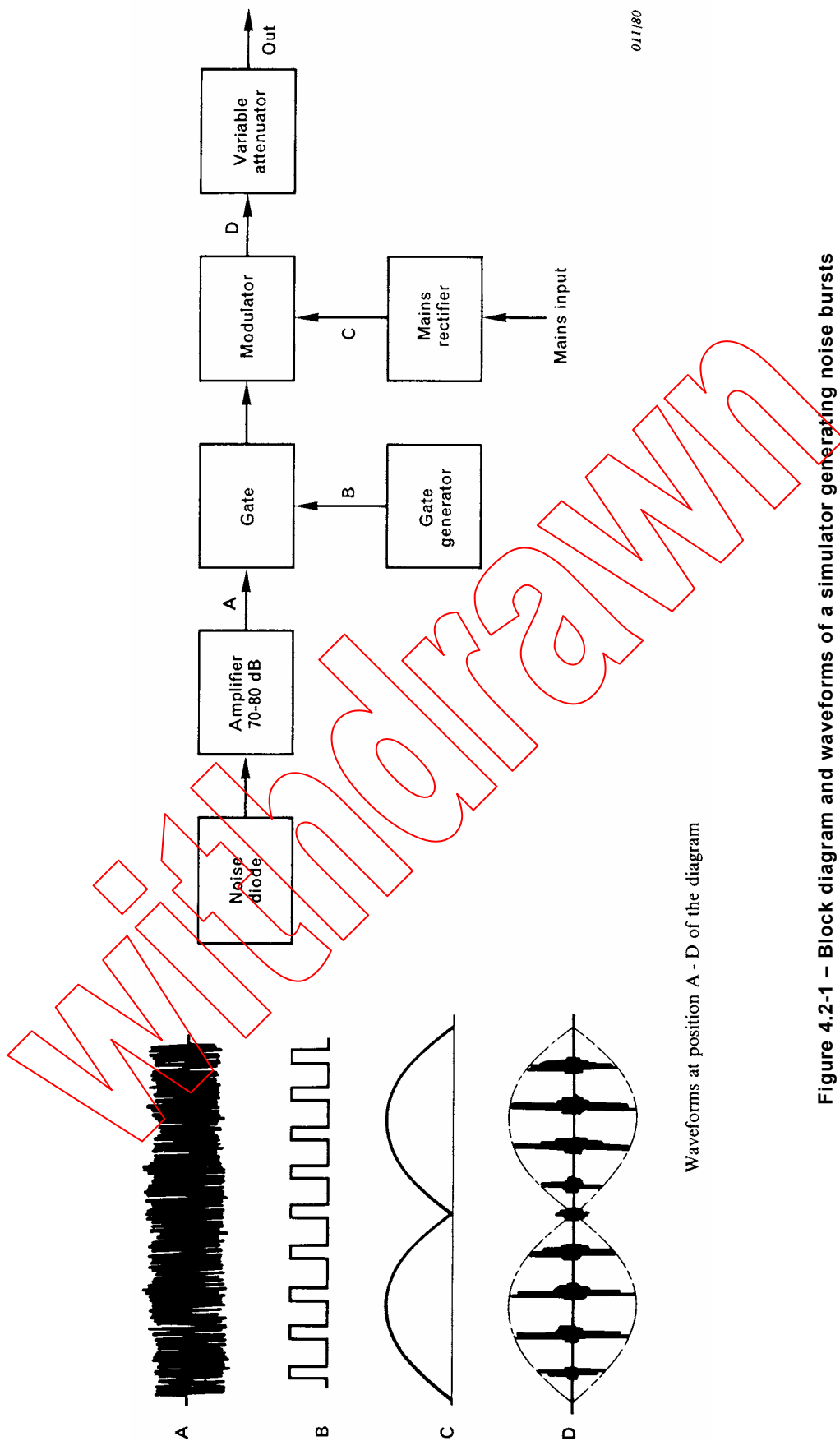
Another way to produce a gated wideband noise signal is given in the diagram of figure 4.2-2. In this design, nanosecond pulses are generated in the output stage, for example, a step recovery diode or similar device. These pulses of constant height are triggered at random time intervals and at a sufficiently high repetition rate to cause overlap in the RF channel under test in order to result in quasi-impulsive noise in the output of the channel. Average repetition rates of a few megahertz are required for measurements in a TV channel of at least 100 kHz for measurements in an FM channel and of at least 10 kHz in an AM channel. The random occurrence of the trigger pulses is obtained from the zero crossings of a broadband signal. For this purpose the output of a noise source is fed to an appropriate amplifier which is followed by a gating circuit for burst simulation. The gated noise signal is fed to a bistable multivibrator which converts the zero crossings into pulses of random varying width from which narrow trigger pulses at random distances are generated by the monostable multivibrator.

The advantage of this system over the circuit of figure 4.2-1 is that the usable frequency range is determined by the output pulses of the step recovery diode only. An example of such a circuit is given in figure 4.2-3, in which circuit output pulses are generated by the step recovery diode HP0102, the pulse width is determined by the length of a short-circuited coaxial cable L . Ringing effects are suppressed by the fast switch diode HP2301 and mains modulation can be effected simply by modulating the supply voltage of the step recovery diode with a full-wave rectified mains voltage. Pulses of 1 ns duration and 5 V amplitude are applicable and offer an output spectrum flat to about 500 MHz. Such a single pulse causes a 50 mV pulse in a TV channel and a 1 mV pulse in an FM channel; overlapping pulses add up, and the peak and quasi-peak value of the resulting signal is considerably higher.

The bandwidth of the preceding stages which generate the trigger signal (noise source, amplifier and gating circuit) should be sufficient for the required pulse repetition rate, so for measurements in a TV channel a bandwidth of 5 MHz to 10 MHz is quite satisfactory. Moreover, the linearity of these stages is not critical because only the position of the zero crossings is important. The multivibrators have to generate steep pulses of short duration (about 0,1 μ s) to drive the step recovery diode.

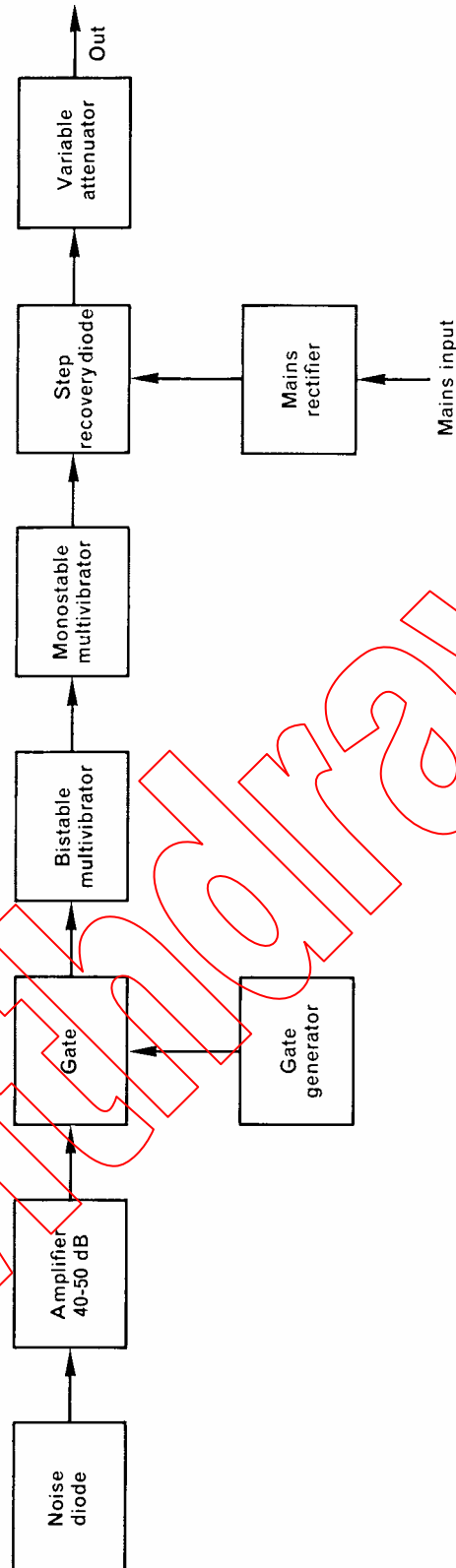
In summary, the circuit according to figure 4.2-1 is very useful for broadband interference simulators to be operated in a limited frequency range, whereas the circuit of figure 4.2-2 is more suitable for simulators intended for wideband use.

Withdrawn



Waveforms at position A - D of the diagram

Figure 4.2-1 – Block diagram and waveforms of a simulator generating noise bursts



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Figure 4.2-2 – Block diagram of a simulator generating noise bursts according to the pulse principle

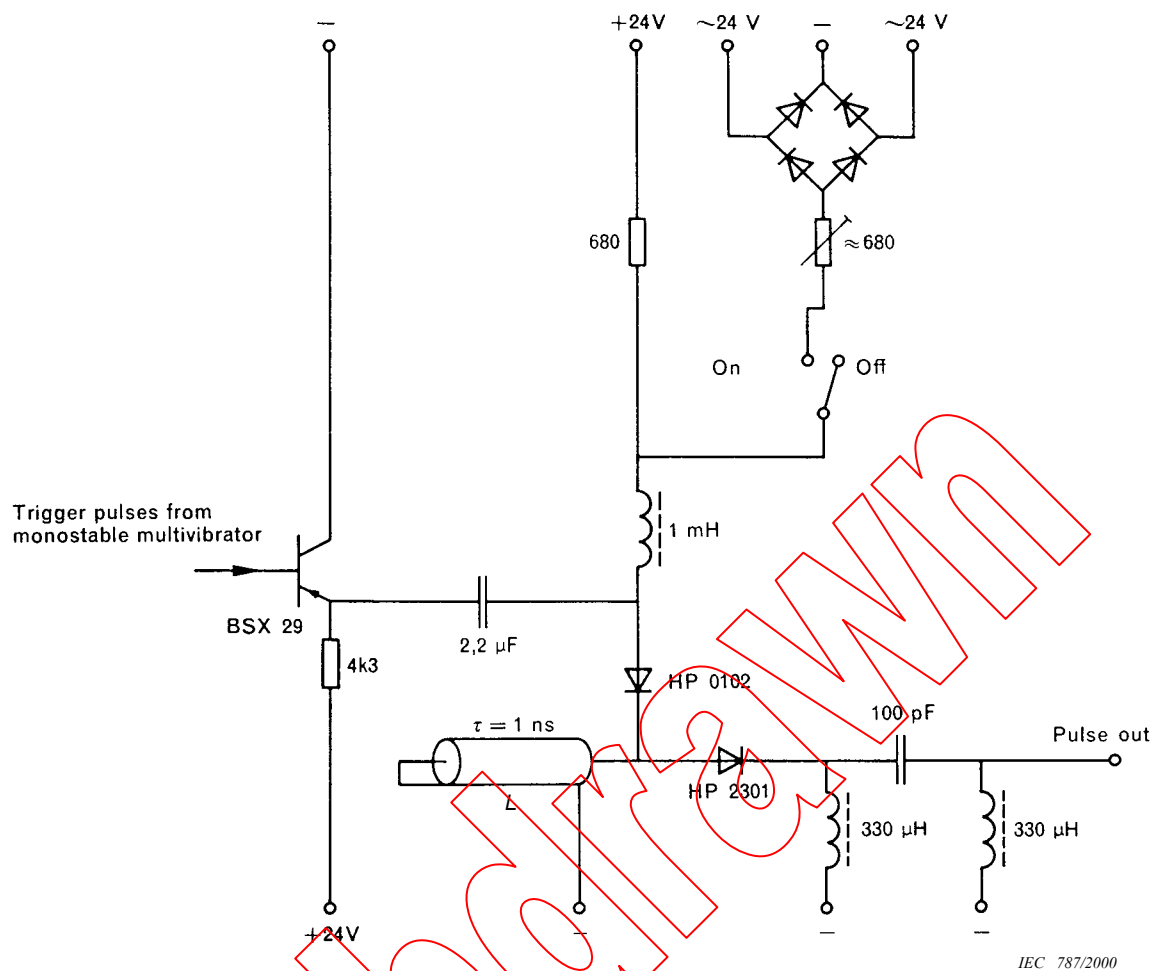


Figure 4.2-3 – Details of a typical output stage

4.3 Relationship between limits for open-area test site and the reverberating chamber

4.3.1 Introduction

At present there are limits for use with the open-area test site method of measurement specified in several CISPR publications. For equipment which can be measured using the reverberating chamber method, a procedure is required to relate the limit to be used with the open-area test site limit. The procedure is described in this report.

4.3.2 Correlation between measurements of the reverberating chamber and the open-area test site

The open-area test site measurement sets out to find the maximum level of radiation of the EUT (equipment under test). Whether the measurement is of the field strength or of power density at the measurement antenna, or of the power into an antenna in substitution of the EUT, the measured results can be expressed in the equivalent, radiated power from a half-wave dipole. Let this equivalent radiated power be P_q dB(pW).

The reverberating chamber measures the total radiated power of the EUT. Let the measured power be P_t dB(pW).

The two measurements are related to each other by the gain of the EUT as a radiator with respect to an isotropic radiator. Let this EUT gain be G dB. The relationship is given by equation (4.3-1). The equation is derived in the appendix.

$$P_t + G = P_q + 2 \quad (4.3-1)$$

4.3.3 Limits for use with the reverberating chamber method

Consider the case of an EUT which is exactly on the limit, L_o , when measured in the open area test site, i.e. $P_q = L_o$.

This EUT should also be exactly on the corresponding limit, L_r , when it is measured in the reverberating chamber, i.e. $P_t = L_r$.

From equation (4.3-1), we can relate the two limits as in equation (4.3-2).

$$L_r = L_o + 2 - G \quad (4.3-2)$$

The value of L_r is dependent not only on L_o , but also on G . Since G will not be the same for EUTs, it is not possible to set L_r to treat EUTs in a manner identical to the effect of L_o . If say $L_r = L_o$, then it is correct only for EUTs with $G = 2$. EUTs with a G greater than 2 will find it easier to pass the reverberating limit, and vice versa.

It is necessary to determine the value of G . This can be done from measurements of P_q and P_t . Figure 1 shows the lines of P_w versus P_t for various values of G . The shaded region is for negative values of G . (Experimental points appearing in this region are caused by failure to locate the maximum open-site radiation, probably due to the maximum radiation lying outside of the horizontal plane.)

An example is given in figure 4.3-2. A number of microwave ovens were measured for P_q and P_t . It can be seen that

- for points lying in the positive G region, the majority have values around 2;
- more points lying in the positive G region as the frequency goes up, indicating that the radiation pattern is becoming more directional in the vertical direction.

On this evidence, the reverberating chamber can be related to the open site. It appears, in fact, to be a more effective method in the ability to measure a quantity representative of the maximum radiation.

4.3.4 Procedure for the determination of the reverberating chamber limit

The procedure is as follows.

- i) Measure a sample of equipment for the maximum radiation on the open-area site. Convert the measured quantities to the equivalent power from a half-wave dipole. Call this quantity P_q , in dB(pW).
- ii) Measure the same sample in the reverberating chamber for total radiated power. Call this quantity P_t , in dB(pW).
- iii) This relationship between the reverberating chamber limit and the open area test site limit can be found by the graphical method of figure 1, or by calculating the gain of each equipment, obtaining a representative value of G for the equivalent type using statistical methods, and applying equation (4.3-2).

Withdrawal

Annex 4.3-A

Derivation of the formula

Consider the general case of a lossless antenna with gain g and input power p . At distance d in the direction of maximum radiation, the field strength e is given by the well-known formula:

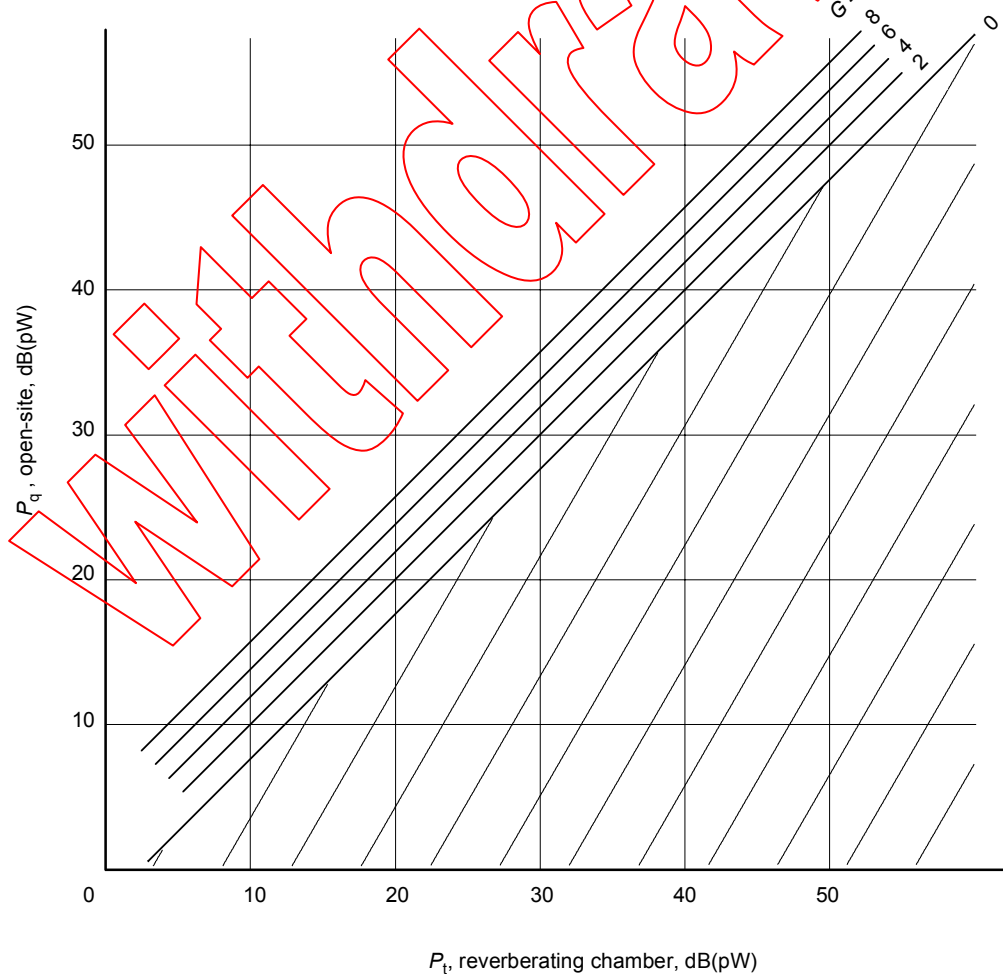
$$e = (30 \times p \times g)^{1/2} / d$$

Thus, for two antennas, with gains g_1 and g_2 and input powers p_1 and p_2 producing the same field strength at a point at the same distance in the direction of maximum radiation

$$p_1 \times g_1 = p_2 \times g_2, \text{ or, in log form, } P_1 + G_1 = P_2 + G_2$$

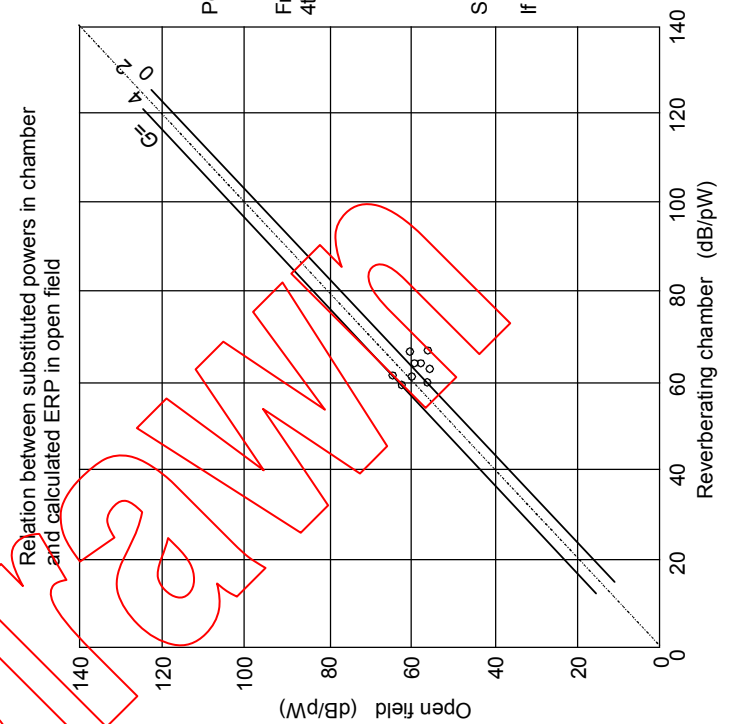
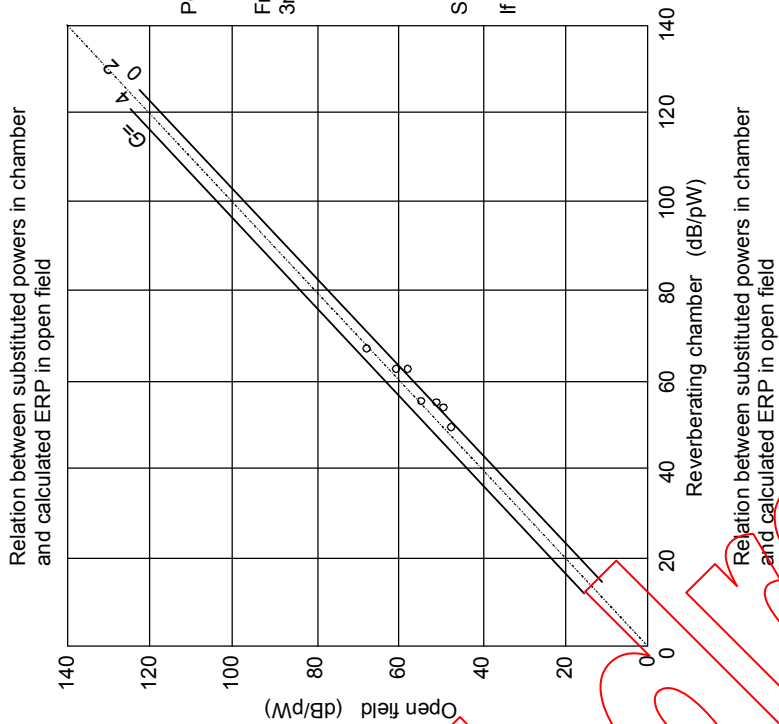
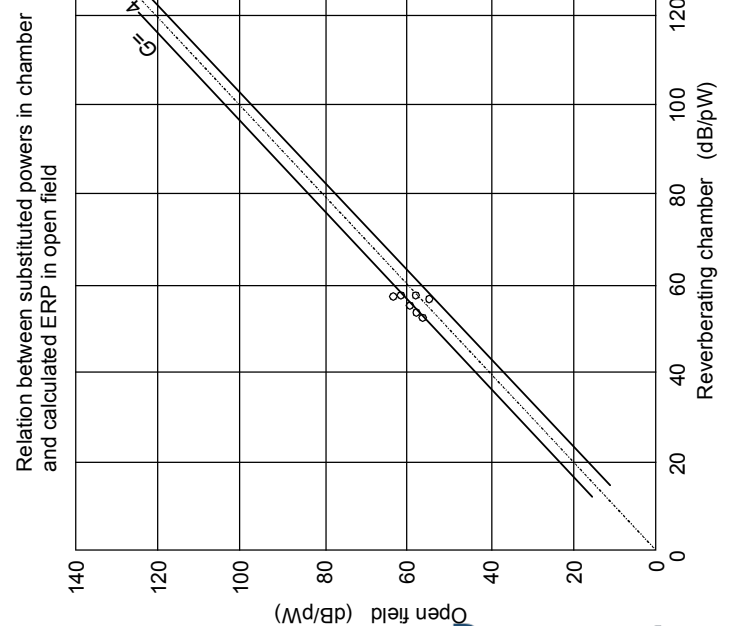
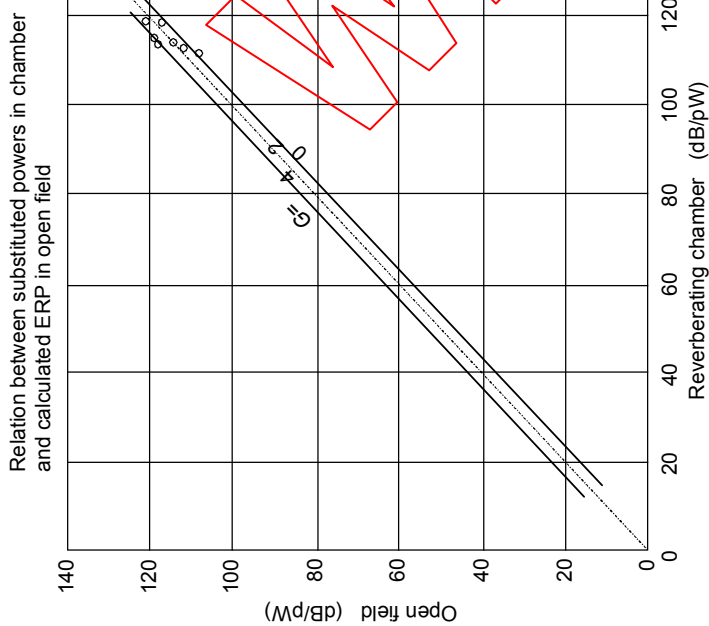
and, in the case given in 4.3.2, $P_1 = P_t$ and $G_1 = G$; $P_2 = P_q$ and $G_2 = 2$ therefore the relationship is

$$P_t + G = P_q + 2$$



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Figure 4.3-1



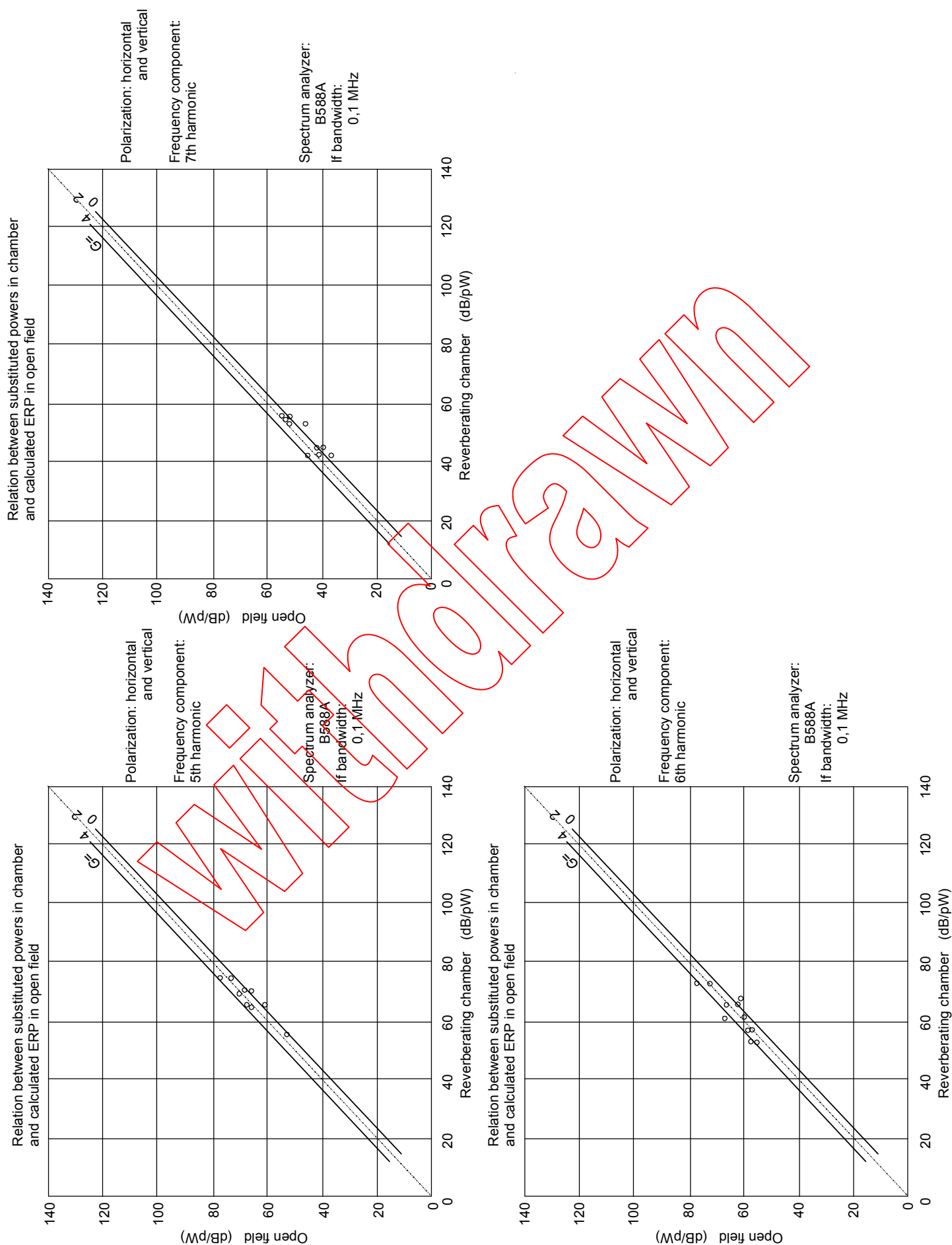


Figure 4.3-2 – Examples of a number of microwave measured for P_{sub} and P_t

4.4 Characterization and classification of the asymmetrical disturbance source induced in telephone subscriber lines by AM broadcasting transmitters in the LW, MW and SW bands

4.4.1 Introduction

The use of semiconductor devices in telephone terminal equipment (TTE) has created the need to verify the immunity to RF fields of this equipment, as non-linear semiconductor devices demodulate the induced RF signals [1-5]. The latter effect gives rise to a d.c. shift which may alter the operating point of such a device, thus, for example, reducing the noise margin of digital devices. In the case of amplitude-modulated RF fields, the non-linearity gives rise to a base-band signal which may become audible in the telephony system. An important class of RF-field sources is formed by AM broadcasting transmitters in the LW, MW and SW bands.

Because of the relatively small dimensions of TTE (compared to the wavelength of the disturbance signal) the asymmetrical (common-mode) source induced in telecommunication lines is expected to be the dominant disturbance source. Therefore, a conducted-immunity test (current-injection test) is relevant for this equipment. In this test, the disturbance signal is applied via the telecommunication lines. As a consequence, this report deals with the characterization of the unwanted antenna properties of telecommunication lines and with prediction models supplying information about the probability that certain parameter values will be met in practice. Moreover, it discusses the parameters which are of relevance when specifying the disturbance source used in the immunity test. The various considerations will be limited to parameters relevant at the subscriber end of the telephone lines.

In 4.4.2 the antenna properties will be expressed in terms of an antenna factor of the subscriber lines i.e. the induced asymmetrical open-circuit voltage normalized to the RF field strength, and an equivalent resistance of the induced asymmetrical source. The prediction models are needed in the classification of the RF fields and the induced asymmetrical disturbances, 4.4.3, and when setting immunity limits, 4.4.4. This report takes the view that the disturbance source in the immunity test is to be specified by an open-circuit voltage and a source impedance.

All mathematical relations associated with the derivation of the models and those needed by the user of this report when applying the models to the respective geographical area are given in annex 4.4-A through 4.4-D.

This report is based on results of experimental investigations carried out on buried telephone-subscriber lines in Germany [6] [7] and in the Netherlands [8]. In these investigations induced-voltage and current data and magnetic-field-strength data were recorded at a large number of locations, permitting a statistical evaluation of the parameter values. A statistical approach was needed, as the telephone lines have random routing in the buildings and, consequently, random orientation with respect to the RF field have random coupling with nearby metal objects, while the buildings cause a random scattering of the RF fields.

It is to be expected that the contents of this report will also be applicable to other types of lines running through buildings in a similar manner to telephone-subscriber lines, for example, bus-system lines and signal and control lines.

4.4.2 Experimental characterization

A full description of the experimental characterization has been presented in [7] [8]. Therefore, this section contains only a summary of this method with regard to the parameters needed in this report.

The induced asymmetrical voltage was measured at the outlet of a telephone-subscriber line using a modified T-network [9] [10]. As a result of this modification, a voltage U_h could be measured across a $10\text{ k}\Omega$ resistor and a voltage U_l across a $150\ \Omega$ resistor. The investigations showed that U_h can be considered as the induced open-circuit voltage. In practice, the reference for this voltage is generally unknown. During the measurements that reference was a special metal measuring cart connected via a copper strap to the central heating system (CH). The equivalent resistance R_a of the induced source is estimated from data pairs $\{U_h, U_l\}$.

At each location two magnetic-field-strength data of the broadcasting transmitter were measured using a loop antenna positioned in a vertical plane and rotated about its vertical axis to find the maximum reading. One datum, H_i , was measured near the outlet under investigation inside the building, and one datum, H_o , was measured outside the building at a distance of about 10 m from that building. In order to obtain a sufficiently high induced-signal-to-ambient-noise ratio, the measurements were carried out in areas with a relatively high value of the RF field strength. This is not expected to influence the applicability of the results, as the presence of a broadcasting transmitter is not taken into account in the layout of telephone-subscriber lines. Moreover, as mentioned in 4.4.1, the induced voltage will be normalized to the field strength and the resulting "antenna factor" will represent a property of the subscriber lines measured.

4.4.2.1 Field strength and building effect

Although the RF field is not a characteristic of the subscriber lines, it forms the origin of the induced disturbances. Two aspects of the RF field will be considered in this section:

- a) The measured field strength H_o outside the buildings compared to the field strength H_c calculated from the simple far-field relation for a half-wave dipole (in its main direction)

$$H_c = \frac{7\sqrt{P}}{Z_0 r} \quad (4.4.2-1)$$

where P is the transmitter power, Z_0 is free-space wave impedance ($377\ \Omega$) and r the distance between the transmitter and the point of observation. In the calculations, the values of P as given in [13] will be used.

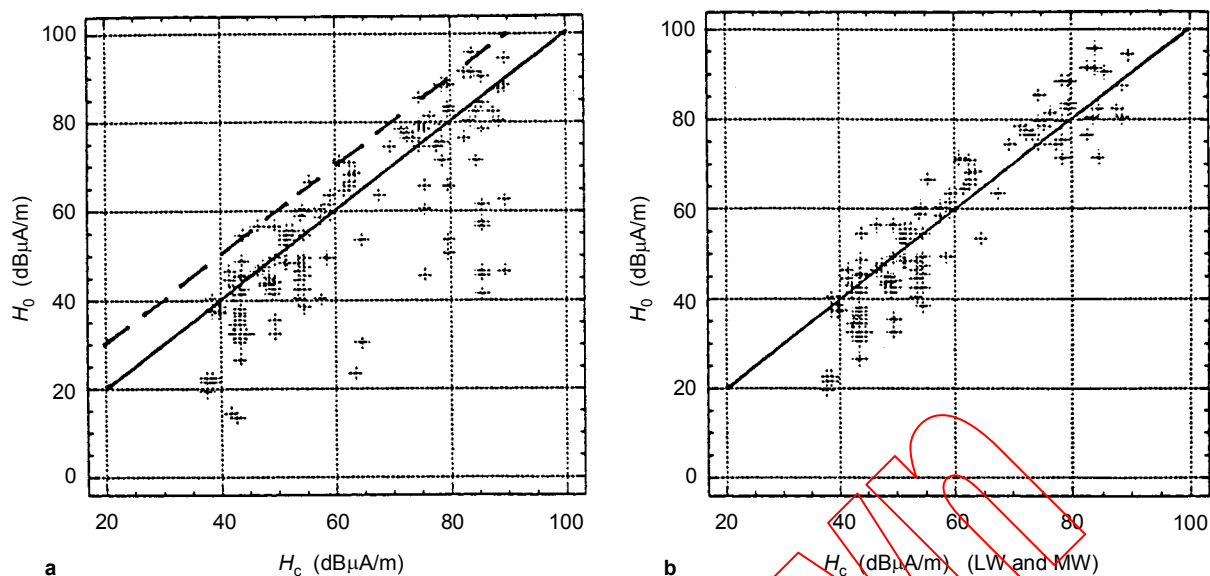
NOTE Although broadcast transmitter antennas usually are monopoles (in the frequencies of interest), the half-wave dipole formula (4.4.2.1) has been used for convenience.

- b) The effect of the building on the field strength, which can be expressed in a building-effect parameter A_b defined by

$$A_b = H_o\ (\text{dB}\mu\text{A/m}) - H_i\ (\text{dB}\mu\text{A/m}) \quad (4.4.2-2)$$

This factor is often called the building attenuation. However, this factor not only depends on the attenuation properties of the building material itself, but also on the reradiation properties of metallic structures in and near the building, and, on the height above ground, H_o and H_i were measured. Therefore the term building effect is used in this report.

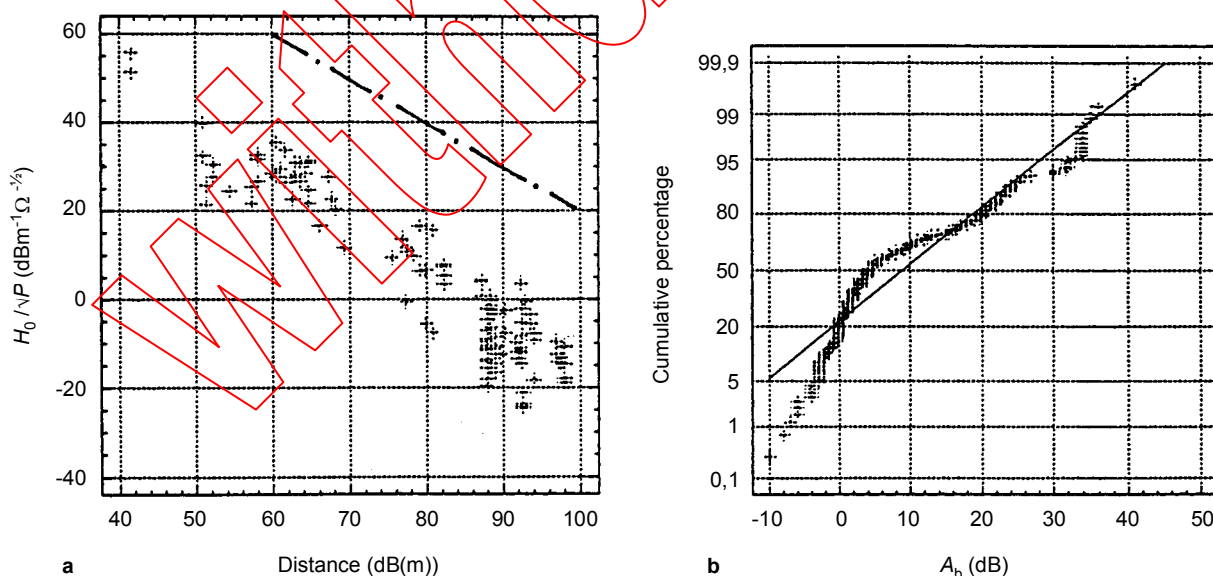
A consideration of these two aspects is needed in view of the antenna factors to be discussed in 4.4.2.2 and in view of the prediction models to be discussed in 4.4.2.3.



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Figure 4.4.2-1 – Scatter plot of the measured outdoor magnetic field strength H_0 (dBμA/m) versus the calculated outdoor magnetic field strength H_c (dBμA/m)
a) all data; b) data of SW transmitters rejected

The results of $H_0(H_c)$ given in figure 4.4.2-1a show that large deviations from $H_0 = H_c$ are possible (drawn line), but that $H_0 \leq (H_c + 10)$ dBμA/m (broken line), hence the measured value is at most a factor of 3 higher than the value calculated from equation 4.4.2-1. The largest deviations concern data of SW transmitters. This is understandable, as SW transmitters normally have a beamed antenna pattern, whereas the antenna patterns of the LW and MW transmitters are generally close to circular. Figure 4.4.2-1b gives the scatter plot $H_0(H_c)$ after rejection of the SW transmitter data.



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Figure 4.4.2-2 –
a) Scatter plot of the measured outdoor magnetic field strength H_0 normalized to the square root of the reported power [11] versus the distance d (m) to the transmitter,
b) Normal probability plot of the building-effect parameter A_b (dB), all data

Figure 4.4.2-2a shows the ratio $H_0/\sqrt{P}$ versus the distance between the point of observation and the transmitter, and the dash-dot line indicates a slope -1 . It can be concluded that, on average, the data follow this slope fairly well. The associated intercept is higher than that expected from equation 4.4.2-1, which is in agreement with the $(H_c + 10)$ dB μ A/m limit observed in figure 4.4.2-1.

Figure 4.4.2-2b shows the normal probability plot of all building-effect data. If these data were normally distributed, a straight line would have resulted. This is not the case, and the data suggest that, in first-order approximation, two distributions are superimposed. The two distributions are found when distinguishing between data associated with buildings constructed from brick and/or wood (B/W) and data associated with buildings constructed from reinforced concrete (C). The normal probability plots of these distributions are given in figure 4.4.2-3a and b. The negative values of A_b predominantly stem from measurements where H_i was measured on an upper floor of the building, whereas H_0 was already measured at about 1,5 m above ground level outside the building. Effects of reradiation also influence the actual field-strength data.

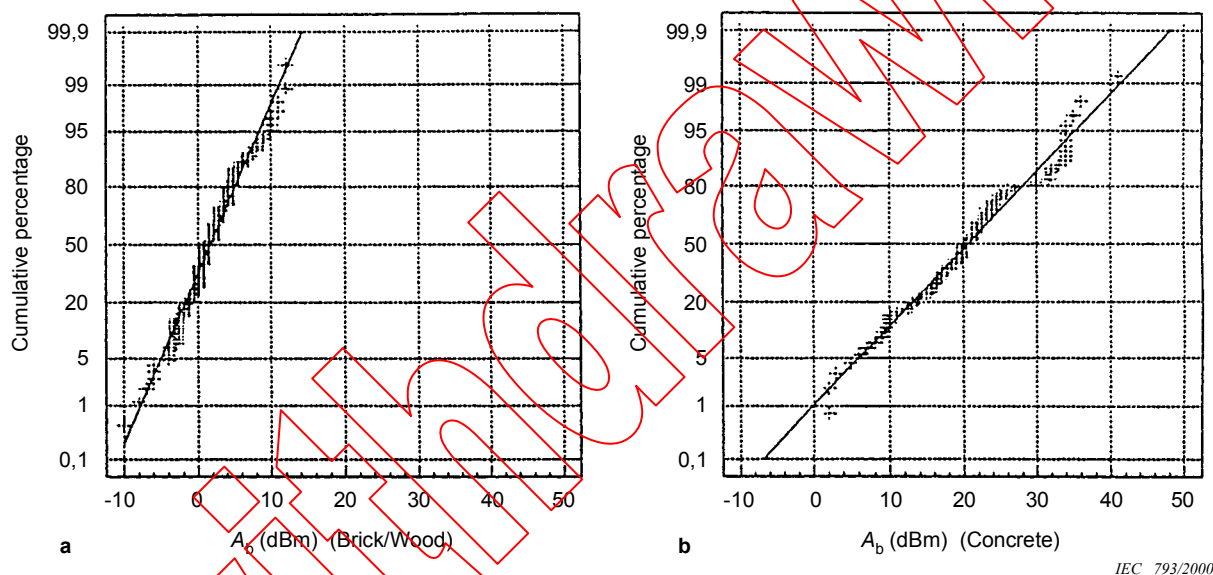


Figure 4.4.2-3 – Normal probability plot of the building-effect parameter A_b (dB)
a) Buildings constructed from brick and/or wood;
b) Buildings constructed from reinforced concrete

The numerical results have been summarized in table 4.4.2-1. No clear frequency dependence of the A_b data could be observed (see 4.4.2.4).

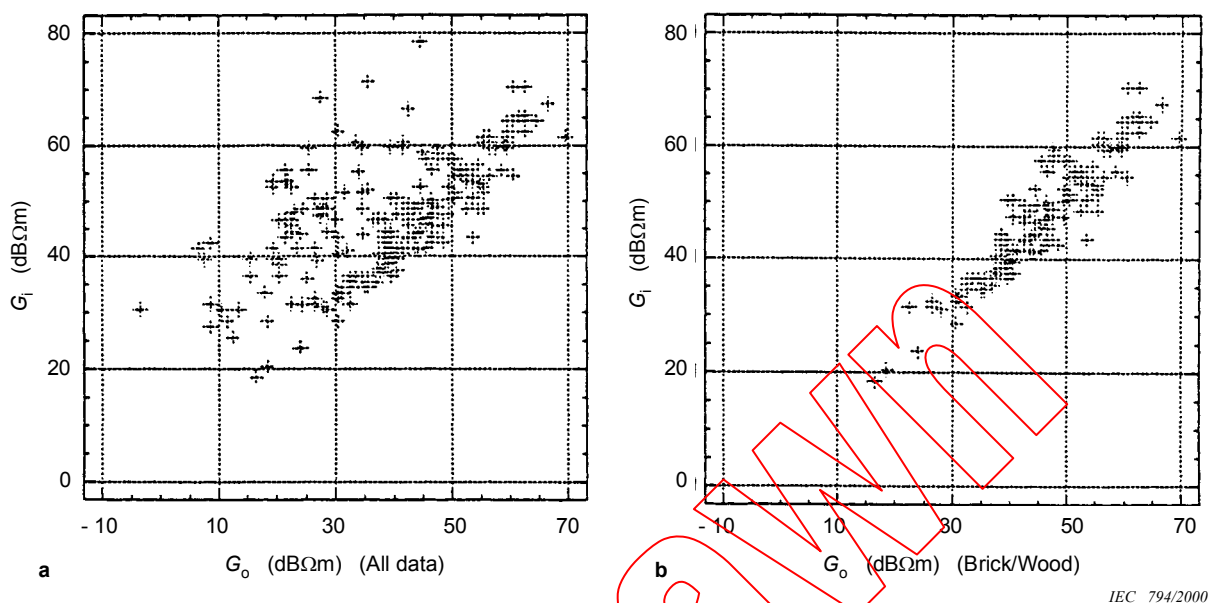
Table 4.4.2-1 – Summary results of building-effect, A_b , analysis

Building material	Average dB	Standard deviation dB	Median dB	Number of data
Brick and/or wood	1,6	4,0	1,0	138
Reinforced concrete	20,6	8,7	20,1	84

4.4.2.2 The asymmetrical open-circuit voltage normalized to the field strength

The interface for the voltage measurements was the outlet to which the telephone set was connected during the measurements. The investigations showed that the influence of the telephone set and its standard lead (4 m long) on the measured voltages could be neglected.

The measured voltage will be normalized to the measured magnetic field strength in 4.4.2.2.1 and, assuming far-field conditions, to the electric field strength in 4.4.2.2.2. After that, 4.4.2.2.3 deals with truncation of the distributions found in 4.4.2.2.1 and 4.4.2.2.2.



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Figure 4.4.2-4 – Scatter plot of the outdoor antenna factor G_o (dBΩm) versus the indoor antenna factor G_i :

a) all data; b) data associated with buildings constructed from brick and/or wood

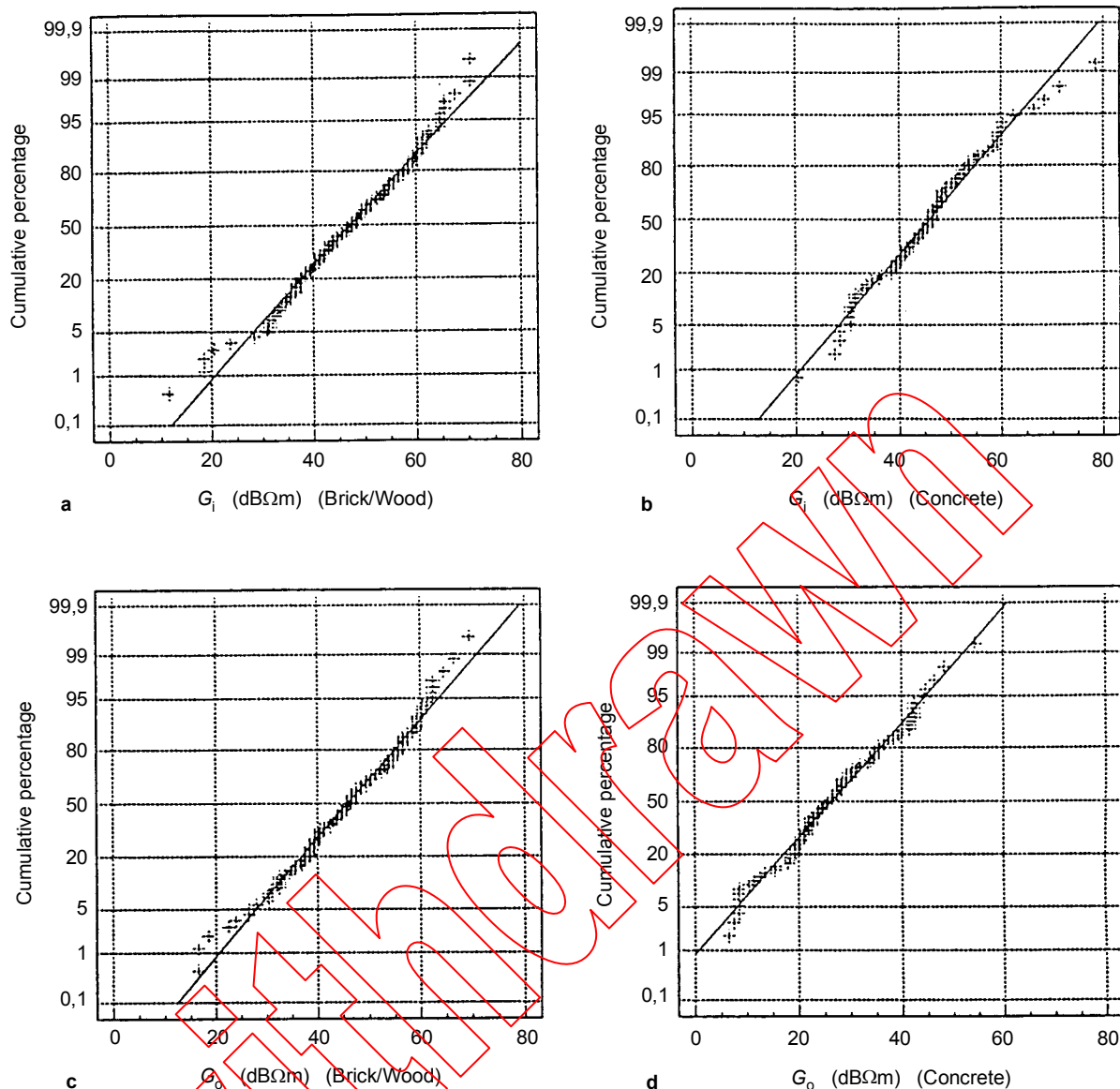
4.4.2.2.1 G factors

To obtain an antenna property of the subscriber lines, the open-circuit voltage U_h is normalized to the field strength (H_i or H_o), yielding the antenna factors G_i and G_o defined by

$$G_{i,o}(\Omega m) = \frac{U_h (\mu V)}{H_{i,o} (\mu A/m)} \quad (4.4.2-3)$$

Figure 4.4.2-4a shows the scatter plot $G_i(G_o)$ using all data. The plot suggests that there is one dominant "cloud" of data with a limited scattering and a second "cloud" with much more scattering. Further investigation revealed that the first cloud stems from data measured in buildings constructed from brick and/or wood, see figure 4.4.2-4b, while the other cloud is associated with buildings constructed (predominantly) from reinforced concrete. Consequently, the building effect discussed in 4.4.2.1 is of importance.

The normal probability plots of G_i (dBΩm) and G_o (dBΩm) associated with the two types of building material considered, are given in figure 4.4.2-5. It can be concluded that the data follow a normal distribution, which means lognormal distributions of the G factors in Ωm. The numerical results have been summarized in table 4.4.2-2, where G_U and G_L are the upper and lower limit of the range of experimental G data (see 4.4.2.2.3). The differences between G_i and G_o of the two classes of building material considered are in line with the building-effect data for these buildings (see table 4.4.2-1). No clear frequency dependence could be observed (see 4.4.2.4).



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Figure 4.4.2-5 – Normal probability plots of the antenna factors:
a) G_i (dBΩm) and c) G_o (dBΩm) associated with buildings constructed from brick and/or wood;
b) G_i (dBΩm) and d) G_o (dBΩm) associated with buildings constructed from reinforced concrete

Table 4.4.2-2 – Summary of results of G-factor analysis

G factor	Building material	Average dBΩm	Standard deviation dB	Median dBΩm	Number of data
G_i	B/W	47,3	11,2	47,5	135
G_i	C	45,9	10,5	46,5	88
G_o	B/W	45,8	10,6	46,4	134
G_o	C	26,5	10,9	26,0	90

4.4.2.2.2 L factors

In 4.4.2.2.1, U_h was normalized to the measured magnetic field strength, thus yielding the G factors. Assuming far-field conditions, the electric field strength follows from $E = H \cdot Z_0$ with $Z_0 = 377 \, \Omega$. If the outdoor field strength is considered, this assumption seems to be reasonable and the G_o factor can be converted into an L_o factor defined by

$$L_o = \frac{G_o}{Z_0} = \frac{U_h}{H_o Z_0} = \frac{U_h \, (\mu V)}{E_o \, (\mu V/m)} \quad (4.4.2-4)$$

The L factor can be considered as the effective length or effective height of the subscriber line acting as an antenna. The results for L_o have been summarized in table 4.4.2-3, where L_U and L_L are the upper and lower limit of the range of experimental L factors (see 4.4.2.2.3).

Table 4.4.2-3 – Summary of L_o factors (far-field)

L factor	Building material	Average* dBm	Standard deviation dB	L_U dBm	L_L dBm
L_o	B/W	–5,7	10,6	18,0	–35,0
L_o	C	–25,0	10,9	3,0	–55,0

* Note that dBm refers to dB with respect to 1 m.

In the literature an L factor of –3,0 dBm average (standard deviation 10 dB, number of data 10) has been reported [12] for a cable running 1 300 m underground and 1 000 m to 3 000 m overhead (aerial cable) towards the subscriber. Broadcasting frequencies were 594 kHz and 1 242 kHz. No details were given about the field-strength measurements, the reference for the asymmetrical voltage and the properties of the building material. The results reported in [12] are in line with the results for L_o (B/W), as given in table 4.4.2-3. However, more recent investigations by the same team [15] indicate an average L -factor of 0 dBm.

L_i factors might be derived from the G_i factors in a similar way as the L_o factors. However, it is to be expected that inside the buildings the far-field conditions are not satisfied and it has to be decided which wave impedance has to be taken. Therefore, no L_i data have been presented in table 4.4.2-3. See also note 2 at the end of 4.1.

4.4.2.2.3 Truncation

In 4.4.2.2.1 it was concluded that the distribution $f(G)$ of the G factors (antenna factors) is lognormal or in mathematical form

$$f(g)dG = \frac{1}{G\sigma\sqrt{2\pi}} e^{-\frac{(\ln G - \mu)^2}{2\sigma^2}} dG \quad (4.4.2-5)$$

However, by adopting this lognormal distribution it is automatically assumed that a G factor may range from zero to infinitely large. In practice, infinitely large will never occur as wavelength effects and effects of coupling with nearby structures create an upper limit (G_U or L_U) of the antenna factors [13]. Consequently, for correct use in the prediction models (4.4.3 and 4.4.4) $f(G)$ has to be truncated. Similarly, truncation has to be applied to the distribution of the building-effect parameters.

Unfortunately, no theoretical study is known which predicts the upper limit of G (or L) of an actual telephone-subscriber line taking into account the length and the routing of a line inside the building and, buried, outside the building. However, it has to be expected that such a limit exists and the best approach is to use the experimental upper limit (G_U or L_U).

In addition to the upper limit, one may also consider a lower limit (G_L or L_L) and truncate $f(G)$ at the lower end. It is found that in the range of parameter values to be considered in 4.4.3 and 4.4.4, the influence of G_L (or L_L) is negligible.

The truncated probability density function reads

$$f_t(G)dG = \frac{f(G)dG}{\int_{G_L}^{G_U} f(G)dG} = \frac{f(G)dG}{F(G_U) - F(G_L)} = \alpha_t f(G)dG \quad (4.4.2-6)$$

The mathematical form of the cumulative distributions $F(G_U)$ and $F(G_L)$ is given in annex 4.4-D. Table 4.4.2-4 summarizes the truncation data of the G factors and the building-effect parameter A_b . Note that α_t differs very little from 1, that is from the value of α_t if $-\infty \leq G$ (dBΩm) $\leq \infty$ or $-\infty \leq a_b$ (dB) $\leq \infty$ since $F(\infty) = 0,5$ and $F(-\infty) = -0,5$. The upper and lower limit in dBm of the L factor range are found by subtracting 51,5 dBΩ from the corresponding G factors in dBΩm.

Table 4.4.2-4 – Summary of truncation parameters of $f(G)$

G factor or A_b	Building material	G_U (dBΩm) A_{bu} (dB)	G_L (dBΩm) A_{bl} (dB)	$F(G_U)$ $F(A_{bu})$	$F(G_L)$ $F(A_{bl})$	α_{tG} α_{tA}
G_i	B/W	70,5	11,5	0,4805	-0,4993	1,021
G_i	C	78,5	20,5	0,4985	-0,4922	1,009
G_o	B/W	69,5	16,5	0,4873	-0,4971	1,016
G_o	C	54,5	-3,5	0,4950	-0,4971	1,008
A_b	B/W	12,0	-10,0	0,4953	-0,4981	1,007
A_b	C	41,0	2,0	0,4905	-0,4837	1,026

4.4.2.3 The equivalent asymmetrical-source resistance

The equivalent resistance of the induced asymmetrical source can be determined from data pairs $\{U_h, U_l\}$, where U_h is the open-circuit voltage and U_l the voltage measured across 150 Ω, using the simple relation

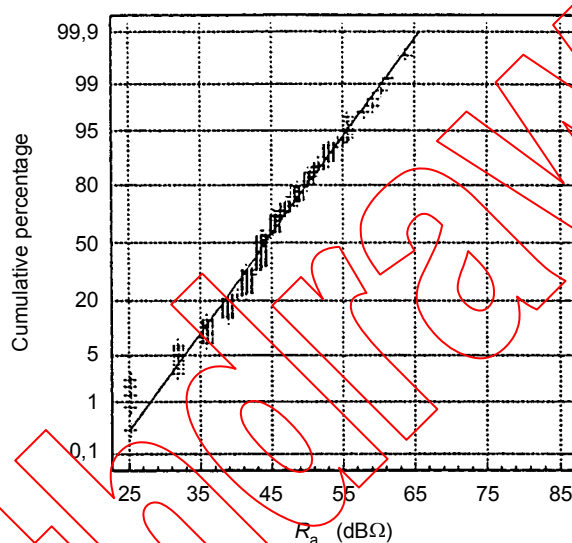
$$R_a = \frac{U_h - U_l}{U_l} 150 \text{ (}\Omega\text{)} \quad (4.4.2-7)$$

The normal probability plot of R_a (dBΩ) is given in figure 4.4.2-6. It can be concluded that R_a (dBΩ) follows a normal distribution and, hence, R_a in Ω follows a lognormal distribution. The numerical results have been summarized in table 4.4.2-5. The average value found is close to the value 150 Ω used in existing immunity tests [9] [14]. In table 4.4.2-5, R_{au} and R_{al} are the upper and lower limit of the range of experimental R_a -data. The relatively large and small values of R_{au} and R_{al} compared to the average value of R_a stem from

resonances and anti-resonances in the common mode circuit of the subscriber line. No clear-frequency dependence of R_a could be observed (see 4.4.2.4) and no influence of the building material was observed.

Table 4.4.2-5 – Summary results of equivalent-resistance analysis

R_a (average) dBΩ	Standard deviation dB	Median dBΩ	R_a (average) Ω	Number of data	R_{au} dBΩ	R_{al} dBΩ	R_{au} Ω	R_{al} Ω
44,2	6,8	43,5	162	204	63,7	25,2	1 531	18



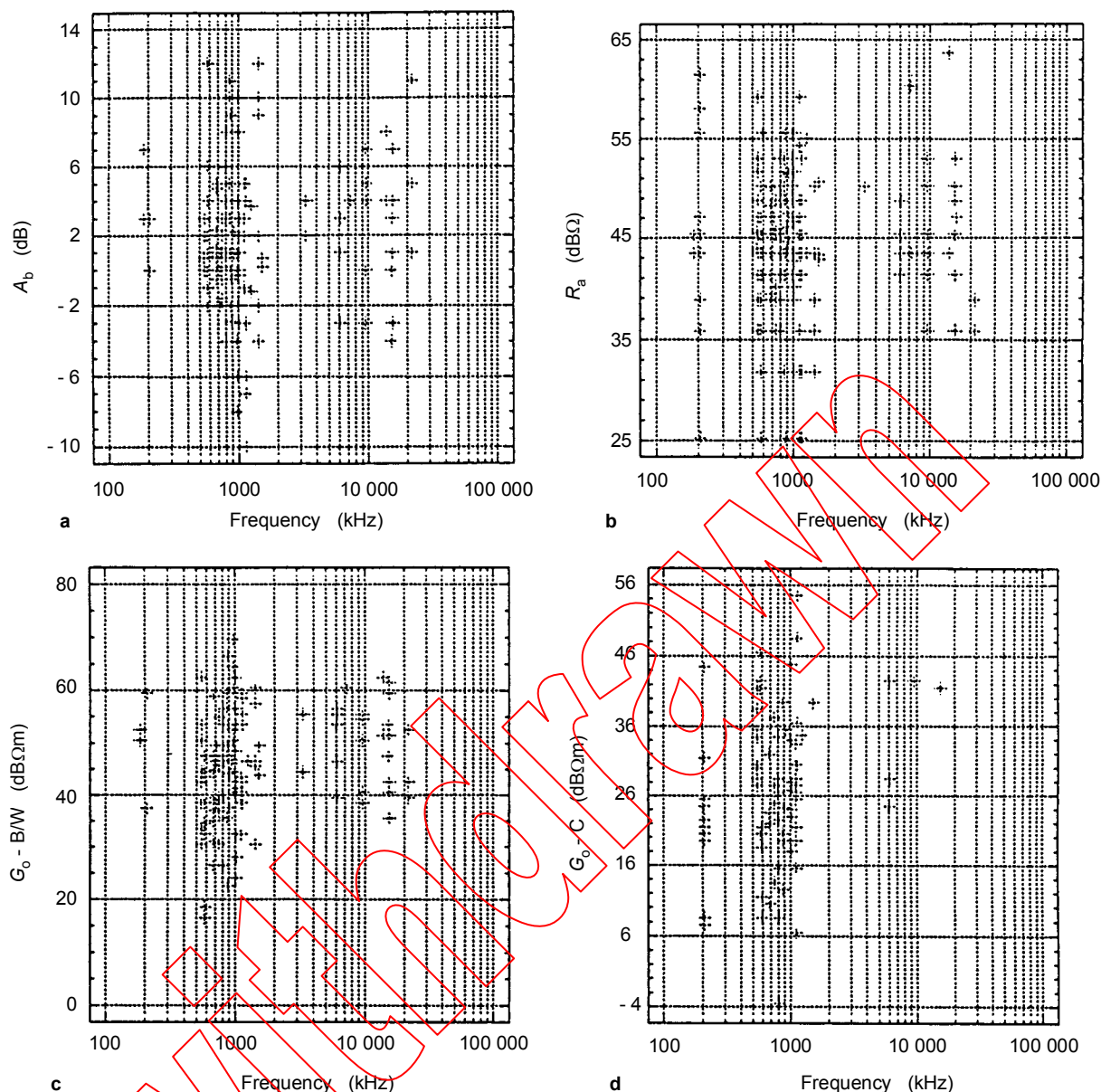
IEC 796/2000

Figure 4.4.2-6 – Normal probability plot of the equivalent asymmetrical resistance R_a (dBΩ)

4.4.2.4 Frequency dependence of the parameters

In the frequency range determined by the measurements in the LW, MW and SW bands, no clear frequency dependence of the building effect A_b , the G factors G_o and G_i , and the equivalent resistance R_a could be observed. This is illustrated in figures 4.4.2-7a and b the A_b data for brick/wood buildings, the R_a data, the G_o data for brick/wood buildings, and for reinforced concrete buildings.

Because no clear frequency dependence of the various quantities could be observed, it will be assumed in 4.4.3 and 4.4.4 that the building effect, the G and the L factors and the equivalent resistance are independent of the frequency in the frequency range of the LW, MW and SW bands. A possible frequency dependence is then incorporated in the standard deviation of the respective distributions.



IEC 797/2000

Figure 4.4.2-7 – Examples of the frequency dependence of some parameters

- a) the building effect A_b in the case of brick/wood buildings
- b) the equivalent resistance R_a
- c) the G_o factor for brick/wood buildings, and
- d) the G_o factor for reinforced concrete buildings

4.4.3 Prediction models and classification

This section presents some simple prediction models for fields and voltages needed in the process of classification of the electromagnetic environment and when setting immunity limits for the telephone sets to be connected to the subscriber lines.

Because the measuring locations were not chosen randomly in order to obtain a sufficiently high induced signal-to-ambient noise ratio, the basic data from which the parameters reported in 4.4.2 were derived cannot be used directly as they form a non-random sample from their actual distributions. The models to be discussed permit an estimate to be made of the complete distributions of the field strength and induced voltage. In addition, the

complete distributions allow for a classification of these quantities. This report gives only the procedure for this classification. The actual class limits are outside its scope.

4.4.3.1 Field-strength classification

As mentioned in 4.4.2.1, the field strength is not a property of the subscriber line. Nevertheless, information about the field strength is needed in order to make a prediction of the induced voltages.

From the results given in 4.4.2.1, it follows that in first-order approximation the outdoor field strength, to be indicated by E_o and H_o for the electric and magnetic field component respectively, is inversely proportional to the distance r between the point of observation and the transmitter, and proportional to the square root of the transmitter power. From the results summarized in 4.4.2.1 it follows that, in the worst case, the constant of proportionality is a factor of 3 (≈ 10 dB) larger than the constant of proportionality in the case of a half-wave dipole.

A classification of the outdoor electric or magnetic field strength may be based on the probability $pr\{E_o \geq E_L\}$ or $pr\{H_o \geq H_L\}$ that the outdoor field strength is equal to or larger than a given limit value, indicated by the subscript L . As explained in annex 4.4-A, this probability can be written as

$$pr\{H_o \geq H_L\} = \int_{H_L}^{H_{\max}} f_H(H) dH \quad \text{or} \quad pr\{E_o \geq E_L\} = \int_{E_L}^{E_{\max}} f_E(E) dE \quad (4.4.3-1)$$

where $f(H)$ and $f(E)$ represent the normalized field-strength distribution and $H_{\max}$ and $E_{\max}$ the maximum field strength in the geographical region in which the probability has to be estimated. Under far-field conditions both relations in equation (4.4.3-1) are equivalent.

Considering a ring-shaped area (RSA) around a transmitter having a circular antenna pattern, it follows (see annex 4.4-A) that

$$pr\{E_o \geq E_L\} = \frac{(E_{\max}^2 - E_L^2)E_{\min}^2}{(E_{\max}^2 - E_{\min}^2)E_L^2} \approx \frac{E_{\min}^2}{E_L^2} \quad (4.4.3-2)$$

where $E_{\max}$ is the field strength at the inner boundary of the RSA and $E_{\min}$ the field strength at the outer boundary of the RSA. A similar expression is valid for the magnetic field strength, see annex 4.4-A.

The inner boundary of the RSA has to be specified, as the relation between E_o and r derived from the measuring data will not be valid arbitrarily close to the transmitter, i.e. in the near-field region of the transmitter. A non-zero outer boundary field strength is needed, as $E_{\min} = 0$ would mean that $pr\{E_o \geq E_L\} = 0$ for all values of E_L . Note that by definition $pr\{E_o \geq E_L\} = 0$ and that $pr\{E_o \geq E_{\min}\} = 1$ ($= 100\%$). The approximation given in equation (4.4.3-2) is valid if $E_{\max} \gg \{E_{\min}, E_L\}$, which is normally the case. Hence, it can be concluded that in the present model the value of $E_{\min}$ to be specified is very important. The choice of $E_{\min}$ and $E_{\max}$ will be discussed further in 4.4.4.

As an example, table 4.4.3-1 gives the values of E_L for a number of probability values, assuming $E_{\max} = 60$ V/m, which is an example of a radiation-hazard limit in the MW band of frequencies, and $E_{\min} = 0,01$ V/m ($= 80$ dB μ V/m). The latter value is of the order of

magnitude of the minimum field strength in the service area of a broadcasting transmitter. Note that the probability values are almost completely determined by $E_{\min}$.

Table 4.4.3-1 – Example of field-strength classification

$pr\{E_o \geq E_L\}$	E_L V/m	$\frac{E_{\min}}{\sqrt{pr\{E_o \geq E_L\}}}$	$P = 500 \text{ kW}$	
			$k = 7$ R_L m	$k = 22$ R_L m
0	60	–	(80)	(260)
100	0,01	–	495000	1550000
10^{-1}	0,32	0,33	15652	49193
10^{-2}	1,00	1,00	4950	11556
10^{-3}	3,16	3,16	1565	4919
10^{-4}	9,86	10,00	495	1556

By expressing the RSA boundaries in terms of a field strength, and not, for example, in terms of the distance between the transmitter and the point of observation, equation (4.4.3-1) is applicable to any transmitter producing a field strength which is inversely proportional to the distance. However, after the classes have been established, a certain transmitter will have a certain value of the constant of proportionality k . Then class boundaries can be associated with distances $R_L = k/P/E_L$ between transmitter and point of observation. In table 4.4.3-1 examples of R_L are given, assuming $k = 7$ (as in the case of a half-wave dipole) and $k = 22$ (the worst-case value found in 4.4.2.1), while the transmitter power $P = 500 \text{ kW}$. The R_L values for $E_L = 60 \text{ V/m}$ have been put in between brackets, as, in the considered frequency range, the far-field condition is not valid at these distances.

The advantage of choosing the field-strength boundaries first is that the classes are the same for all kinds of transmitter, while the choice of a class is then determined by the probability that victim equipment will be at a certain distance from the chosen class of transmitters. In general, an estimate of the latter probability is easier than an estimate of the field-strength probability.

4.4.3.2 Asymmetrical-voltage classification

A classification of the induced open-circuit common-mode voltage U_h may be based on the probability $pr\{U_h \geq U_L\}$ that U_h is equal to or larger than a given limit value U_L . If $f_t(G)$ describes the truncated distribution of G factors (see 4.4.2.2.3), $f_n(H_o)$ the normalized field-strength distribution and use is made of the relation $U_h = G_o \cdot H_o$, it is shown in annex 4.4-B that this probability can formally be written as

$$pr\{U_h \geq U_L\} = \int_{G_1}^{G_2} dG_o \int_{U_1}^{U_2} dU_h \frac{1}{G_o} f\left(\frac{U_h}{G_o}\right) \cdot f_t(G_o) \quad (4.4.3-3)$$

where G_1 , G_2 , U_1 and U_2 are suitably chosen boundaries (see annex 4.4-B). In this equation the product of the two distributions, i.e. the joint distribution, is needed as $pr\{U_h \geq U_L\}$ depends on simultaneously meeting a certain field-strength value $H_o = (=U_h/G_o)$ and a certain value of G_o . In equation (4.4.3-3) the factor $1/G_o$ stems from the transformation of $f(H_o)$ into $f(U_h/G_o)$.

Note that equation (4.4.3-3) is not an explicit function of the distance between transmitter and point of observation as a consequence of the fact that the boundaries of the RSA have been defined by field-strength values. A similar remark has been made in connection with equation (4.4.3-1) and similar conclusions are possible here.

Considering again the ring-shaped area as introduced in 4.4.3.1, examples of the classification of U_h , i.e. U_L values corresponding with chosen values $pr\{U_h \geq U_L\}$, are given in table 4.4.3-2 classification. The relations used can be found in annex 4.4-B. As in 4.4.3.1, it was assumed that the outdoor field strength $E_{\max} = 60$ V/m ($H_{\max} = 0,16$ A/m) and $E_{\min} = 0,01$ V/m ($H_{\min} = 27$ μ A/m) have been specified. When using G_i and specifying the outdoor field strength, the building effect has to be taken into account, as is explained in annex 4.4-B.

Table 4.4.3-2 – Example of voltage classification assuming for the outdoor field strength:
 $E_{\max} = 60$ V/m and $E_{\min} = 0,01$ V/m

Building material	G_i		G_o	
	B/W	C	B/W	C
A_b dB	1,6	20,6	–	–
S_b dB	4,0	8,7	–	–
A_{bu} dB	12,0	41,0	–	–
A_{bi} dB	–10,0	2,0	–	–
$G_{i,o}$ dB Ω m	47,3	45,9	45,8	26,5
S dB	11,2	10,5	10,6	10,9
G_U dB Ω m	70,5	78,5	69,5	54,5
G_L dB Ω m	11,5	20,5	16,5	–3,5
$pr\{U_h \geq U_L\}$	U_L dB μ V	U_L dB μ V	U_L dB μ V	U_L dB μ V
10^{-1}	115	101	114	96
10^{-2}	125	111	124	106
10^{-3}	135	121	134	116
10^{-4}	145	131	143	125

4.4.4 Characterization of the immunity-test disturbance source

The results presented in 4.4.2 and 4.4.3 of this report may be used to specify the open-circuit voltage and the internal impedance of the disturbance source in a conducted-immunity test needed to achieve a sufficiently high probability that TTE to be connected to the subscriber lines will be electromagnetically compatible.

The specification of the open-circuit voltage U_h should be based on the distribution of the voltages over all telephone outlets to be considered. Therefore, this distribution is calculated first, using models and parameter values derived in the preceding sections. Once the distribution is known it is possible to calculate $N_o(U_h \geq U_L)$, i.e. the number of outlets in the respective geographical region showing a voltage $U_h \geq U_L$, where U_L may be considered as the open-circuit voltage in the immunity test. The internal impedance may be chosen directly using the results given in table 4.4.2-3. After that, the relevant parameters for the specification of the disturbance source in the immunity test will be summarized in 4.4.4.2.

This section only gives the procedures to arrive at a specification of the parameters needed. The assignment of actual values is the prerogative of the Product Committees.

4.4.4.1 Outlet-voltage distribution

The derivation of the outlet-voltage distribution is described in detail in annexes 4.4-B and 4.4-C. In this derivation, the following steps have been taken.

- 1) Determine the total probability density $n(H_o)$ of the telephone outlets experiencing an outdoor magnetic field strength H_o in ring-shaped areas around the N transmitters to be considered, where the inner boundary of the areas is specified by a maximum field strength $H_{\max}$ and the outer boundary by a minimum field strength $H_{\min}$.

(Again, the magnetic field strength is considered first, as this field strength was measured in the experimental characterization procedure. If far-field conditions are valid, so that the magnetic and electric field strength have a constant ratio, the results can be converted directly in terms of the electric field strength.)

- 2) Determine the joint probability density $f(H_o, G_o) = f(H_o) \cdot f(G_o)$ describing the density of outlets where the field strength has a magnitude H_o and, simultaneously, the antenna factor of the subscriber lines has the magnitude G_o , and convert the result into the joint probability density $f(U_h, G_o)$, by using the relation $U_h = H_o G_o$.
- 3) Calculate the probability $pr\{U_h \geq U_L\}$. If N_T is the total number of outlets in the respective geographical region and the boundary conditions are taken into account such that $N_o(U_h \geq U_L) = N_T$ (or $= 0$) if $pr\{U_h \geq U_L\} = 1$ (or $= 0$), then the number of outlets $N_o(U_h \geq U_L)$ equals $N_T \cdot pr\{U_h \geq U_L\}$.

It is the prerogative of a Product Committee to choose a value of $N_o(U_h \geq U_L)$ from which U_L follows and hence the open-circuit voltage of the disturbance source in the immunity test.

Assuming the field strength to be inversely proportional to the distance between the outlet and the transmitter and assuming constant densities of outlets around the transmitters, it is shown in annex 4.4-C that the field-strength distribution $n(H_o)$ can be written as

$$n(H_o) = \frac{2\pi \sum_{j=1}^N \mu_j k_j^2 P_j}{E_o^3} = \frac{-C_{Eo}}{E_o^3} \quad (4.4.4-1)$$

where μ_j is the outlet density, k_j the constant of proportionality of the j -th transmitter and N the total number of transmitters considered. If the density μ is the same around all transmitters and all transmitters have the same constant of proportionality k , CH_o is simply a constant times the sum over all transistor powers.

When considering the electric field strength $E_o = (k/P)/r$, the distribution $n(E_o)$ reads

$$n(E_o) = \frac{2\pi \sum_{j=1}^N \pi_j k_j^2 P_j}{E_o^3} = \frac{-C_{Eo}}{E_o^3} \quad (4.4.4-2)$$

so that $C_{Eo} = C_{Ho} \cdot Z_o^2$. In annexes 4.4-A and 4.4-B it is explained how the various relations change when the indoor field strength H_i or E_i is to be used.

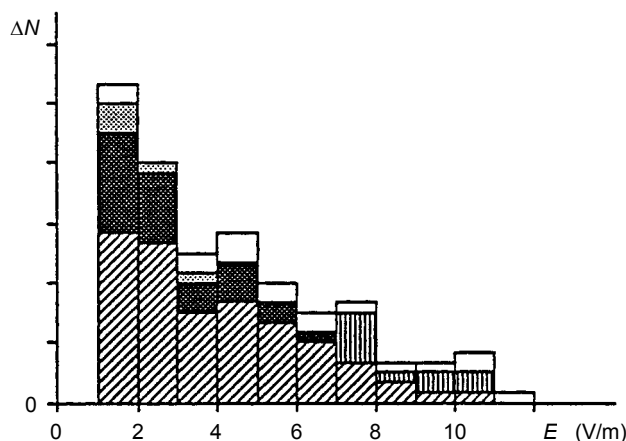


Figure 4.4.4-1 – Example of the frequency histogram $\Delta N(E_o, \Delta E_o)$. The various shadings indicate schematically the contributions from various transmitters and non-homogeneous outlet densities. In this example the field-strength step $\Delta E_o = 1$ V/m.

The outlet density will, in general, not be homogeneous around a transmitter. To derive $n(H_o)$ in that case, a possible procedure is to determine a frequency histogram $\Delta N(H_o, \Delta H_o)$ so that $n(H_o)$ is approximated by $n(H_o) = \Delta N(H_o, \Delta H_o) / \Delta H_o$.

In practice, the magnitude of the electric field strength is mostly considered, so $n(E_o)$ may be determined first, after which $n(H_o)$ follows after assuming far-field conditions. An example of $\Delta N(E_o, \Delta E_o)$ is given in figure 4.4.4-1, where the different shadings indicate the various contributions resulting from various transmitters and non-homogeneous outlet densities around these transmitters.

A drawback of the method leading to figure 4.4.4-1 and a drawback of the model leading to equations (4.4.4-1) and (4.4.4-2) is that, in particular in the lower field strength regions the fields of the various transmitters overlap. As a result, the same outlets in these regions are counted more than once if no discrimination is made with respect to the broadcast frequency. In 4.4.2.4 it has been explained that no real frequency dependence could be observed, so this discrimination is not possible, which leads to an over-estimate of $\Delta N(E_o, \Delta E_o)$ at the lower field-strength values.

A procedure which might be followed is then to determine $n_m(E_o)$, the distribution of the maximum field strength in the respective geographical region resulting from the N relevant transmitters in (and in the direct vicinity of) that region. An example of such a distribution is given in figure 4.4.4-2. This distribution resulted for a part of Germany (having an area of $2,5 \cdot 10^5$ km² and $42 \cdot 10^6$ outlets) by calculating the maximum field strength in each node at a 1 km by 1 km grid over this region as caused by one of the 79 actual broadcasting transmitters in the respective frequency range, having a total ERP of 12,2 MW. The resolution of the field strength, i.e. ΔE_o was taken to be 0,1 dBμV/m. It was assumed that the density μ was a constant ($42 \cdot 10^6 / 2,5 \cdot 10^5 = 168$ km⁻²) throughout the region. When performing these calculations it was found that $n_m(E_o)$ does not vary much after a certain number (50 in this case, with a total power of 7,5 MW) of transmitters had been taken into account.

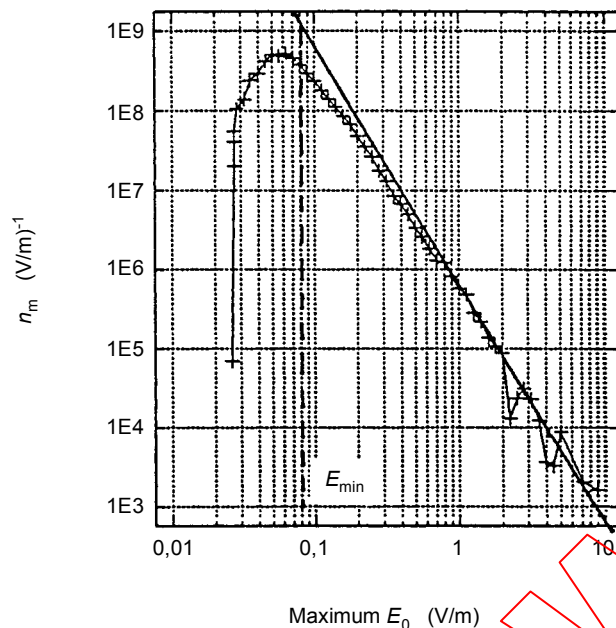


Figure 4.4.4-2 – Example of $n_m(E_0)$ the distribution of the outlets experiencing a maximum field strength E_0 resulting from a given number of transmitters in (or near) the respective geographical region. The drawn line represents $n(E_0) = -C_{E_0}/E_0^3$ with $n(E_0 = 1 \text{ V/m}) = n_m(E_0 = 1 \text{ V/m}) \cdot E_{\min}$ has been chosen such that the integral over the field strength of $n_m(E_0)$ and $n(E_0)$ both yield the total number of outlets in that region

The drawn line in figure 4.4.4-2 represents $n(E_0) = -C_{E_0}/E_0^3$ with C_{E_0} taken such that $n_m(E_0) = n(E_0)$ when $E_0 = 1 \text{ V/m}$. This value is somewhat smaller than C_{E_0} calculated from equation 4.4.4-2 due to the aforementioned fact that $n_m(E_0)$ seems to "saturate" when the number of transmitters is increased.

When integrating the distributions over the whole of the respective region, the total number of outlets N_T must follow, i.e.

$$N_T = \int_{E_{\min}}^{E_{\max}} n(E_0) dE_0 = \frac{C_{E_0}}{2} \left\{ \frac{1}{E_{\min}^2} - \frac{1}{E_{\max}^2} \right\} \quad (4.4.4-3)$$

where the right-hand side of equation (4.4.4-3) follows when equation (4.4.4-2) is used. Equation (4.4.4-3) indicates that for a given or agreed maximum field strength $E_{\max}$, the value of $E_{\min}$ follows when C_{E_0} is given or that C_{E_0} follows when $E_{\min}$ is given. The former approach has been used to calculate $E_{\min} = 0,08 \text{ V/m}$ as indicated in figure 4.4.4-2, assuming $n_m(E_0 = 1 \text{ V/m}) = C_{E_0}/E_0^3 = C_{E_0} \cdot 5,4 \cdot 10^5 \text{ (V}^2/\text{m}^2)$.

Table 4.4.4-1 – Summary of the parameters used in the numerical examples presented in figures 4.4.4-3 and 4.4.4-4. Some curves have the same parameter values: Figure 4.4.4-4a) curve 2 and figure 4.4.4-3, and figure 4.4.4-4b curve 1 and figure 4.4.4-4a curve 3

Figure	4.4.4-3	4.4.4-4a			4.4.4-4b		
Curve	–	1	3	4	2	3	4
L-factor	L_o	L_o	L_o	L_o	L_i^*	L_o	L_i^*
Building material	B/W	B/W	B/W	B/W	B/W	C	C
M_L dBm	–5,7	–5,7	–5,7	–5,7	–4,2	–25,0	–5,6
S_L dB	10,6	10,6	10,6	10,6	11,2	10,9	10,5
L_u dBm	18,0	$+\infty$	18,0	18,0	19,0	3,0	27,0
L_u m	7,9	$+\infty$	7,9	7,9	8,9	1,4	22,4
L_i dBm	–35,0	$-\infty$	–35,0	–35,0	–40,0	–55,0	–31,0
L_i m	0,02	$-\infty$	0,02	0,02	0,01	0,002	0,03
M_A dB	–	–	–	–	1,6	–	20,6
S_A dB	–	–	–	–	4,0	–	8,7
A_{bu} dB	–	–	–	–	12,0	–	41,0
A_{bl} dB	–	–	–	–	–10,0	–	,0
N_T millions	42	42	42	42	42	42	42
C_{Eo} V ² /m ²	$5,4 \cdot 10^5$	$5,4 \cdot 10^5$	$5,4 \cdot 10^5$	$5,4 \cdot 10^5$	–	$5,4 \cdot 10^5$	–
C_{Ei} V ² /m ²	–	–	–	–	$3,3 \cdot 10^4$	–	–
E_{max} V/m	10,0	10,0	3,0	10,0	3,0	3,0	3,0
$E_{i,max}$ V/m	–	–	–	–	9,5	–	2,4
E_{min} V/m	0,08	0,08	0,08	0,008	0,08	0,08	0,08
$E_{i,min}$ V/m	–	–	–	–	0,02	–	0,0007
U_{max} V	79	$+\infty$	24	79	85	42	53

* See note 3 at the end of this subclause.

This subclause is concluded by giving several numerical examples of $N_o(U_h \geq U_L)$, i.e. of the number of outlets in the respective geographical region showing an induced open-circuit voltage $U_h \geq U_L$, where U_L may be considered as the open-circuit voltage in the immunity test. The region in these examples is the aforementioned part of Germany. The relations used in the calculation of $N_o(U_h \geq U_L)$ can be found in annexes 4.4-B and 4.4-C. The values of the various parameters used in these calculations have been summarized in table 4.4.4-1.

Figure 4.4.4-2 shows an example of $N_o(U_h \geq U_L)$ in the voltage range $U_L \leq U_{max} = 79$ V, based on L_o data for buildings constructed from brick and/or wood (B/W) and assuming that all 42 million outlets are in these types of building. U_{max} (see also table 4.4.4-1) is given by $U_{max} = E_{max}L_u$, where E_{max} is the maximum outdoor field strength in the respective geographical region and L_u the upper limit of the range of L_o (B/W) data (see 4.4.2.2.2). U_{max} is quite large in this case. The maximum value measured in the experimental investigations was 22 V.

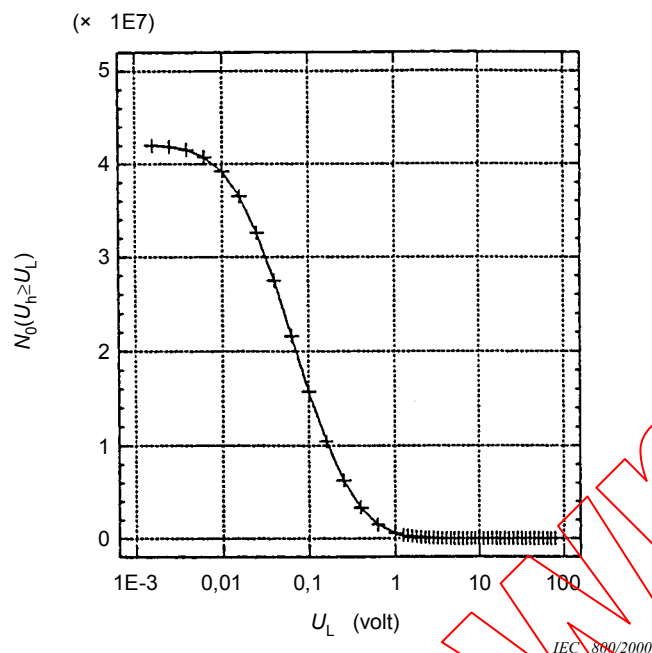
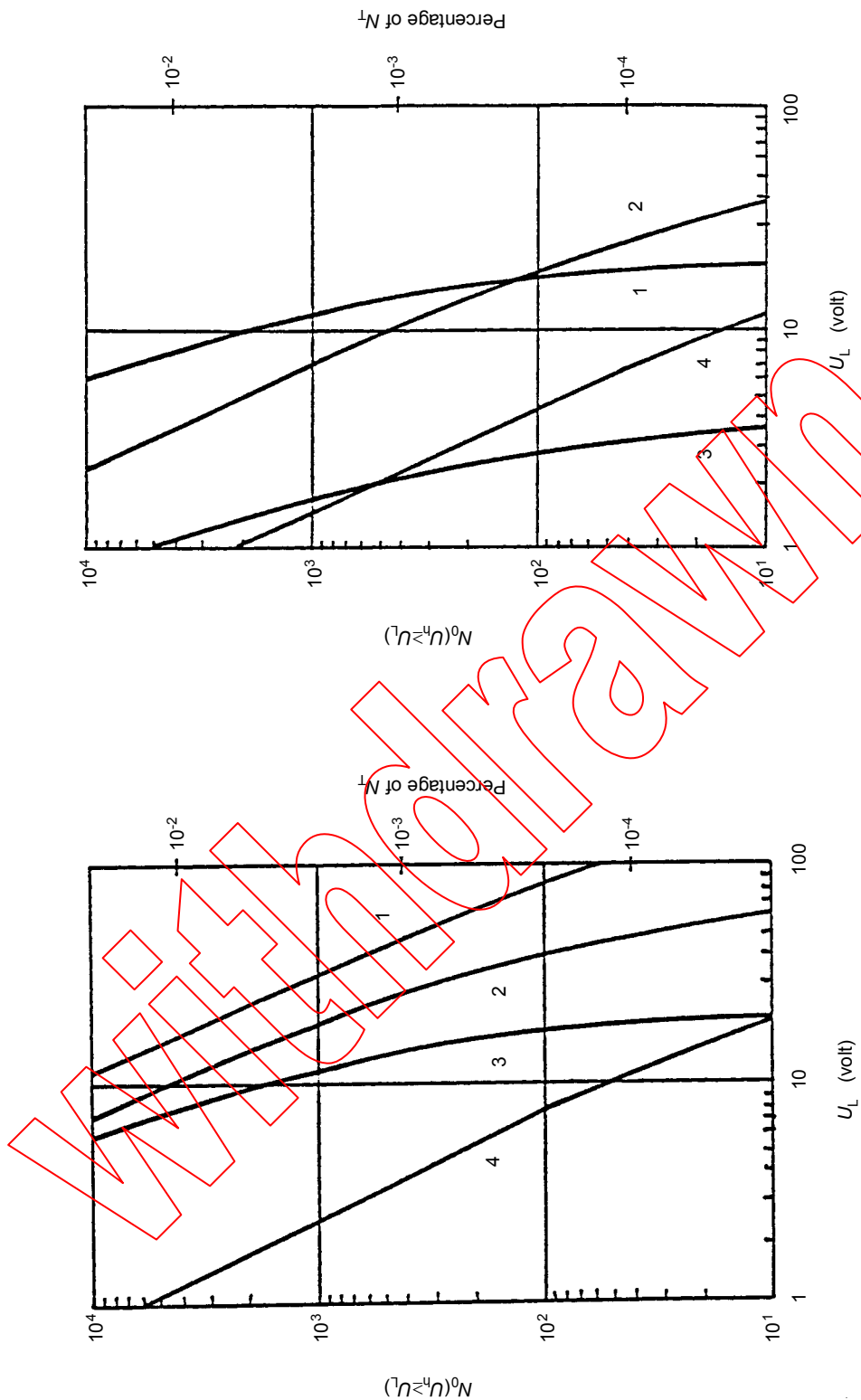


Figure 4.4.4-3 – Example of the number of outlets with an induced asymmetrical open-circuit voltage $U_L \leq U_h \leq U_{\max} = 79 \text{ V}$, see table 4.4.4-1.

Total number of outlets $N_T = 42 \cdot 10^6$

To set the specifications for an immunity test the higher range of U_L values is of interest. Therefore, the results given in figure 4.4.4-3 have been replotted as Curve 2 in figure 4.4.4-4a. Curve 1 in figure 4.4.4-4a gives $N_o(U_h \geq U_L)$ for the same situation as in figure 4.4.4-3 but neglecting the truncation of the L factor distribution. In that case $U_{\max}$ is infinitely large, which is not very realistic. Curve 3 in figure 4.4.4-4a demonstrates how the results represented by Curve 2 are modified when $E_{\max}$ is reduced from 10 V/m to 3 V/m. Then $U_{\max} = 24 \text{ V}$. Finally, Curve 4 in figure 4.4.4-4a demonstrates how the results represented by Curve 2 are modified when $E_{\min}$ is reduced from 0,08 V/m to 0,008 V/m. The latter curve clearly demonstrates the importance of the minimum field strength in the geographical region.

In figure 4.4.4-4b the influence of the building material (B/W or reinforced concrete, C) and the choice of L factor (L_o or L_i) on the results can be observed. Curve 1 in that figure is identical to Curve 3 in figure 4.4.4-4a, so it concerns L_o data for B/W-buildings. When using L_i data the building effect has to be taken into account as there is no direct model to predict the indoor field-strength distribution. By doing so, the results represented by Curve 2 are found. As for Curve 1 the maximum outdoor field-strength $E_{\max} = 3 \text{ V/m}$, but due to the building effect the maximum indoor field strength $E_{i,\max} = E_{\max} A_{bi}$, where A_{bi} is the lowest building effect as determined from the experimental data. In this case $A_{bi} = -10 \text{ dB}$, so that $E_{i,\max} = 9,5 \text{ V/m}$, hence an amplification of the outdoor field strength.



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Figure 4.4.4-4 – Examples of number (left-hand scale) and relative number (right-hand scale) of outlets with $U_L \leq U_h \leq U_{\max}$. See table 4.4.4-1 for the various parameter values used. Total number of outlets $42 \cdot 10^6$, homogeneous density of the outlets: $\mu = 168 \text{ km}^{-2}$

a) B/W buildings

Curve 1: nontruncated L_o factor,	$E_{\max} = 10 \text{ V/m},$	$E_{\min} = 0,08 \text{ V/m}$
Curve 2, truncated L_o factor,	$E_{\max} = 10 \text{ V/m};$	$E_{\min} = 0,08 \text{ V/m}$
Curve 3, truncated L_o factor,	$E_{\max} = 3 \text{ V/m};$	$E_{\min} = 0,008 \text{ V/m}$
Curve 4, truncated L_o factor,	$E_{\max} = 10 \text{ V/m};$	$E_{\min} = 0,08 \text{ V/m}$

b) Truncated L factors, maximum outdoor field strength 3 V/m

- Curve 1: B/W buildings, L_o factor
- Curve 2: B/W buildings, L_i factor
- Curve 1: C buildings, L_o factor
- Curve 2: C buildings, L_i factor

One might argue that a negative value of A_{bi} is not realistic. As mentioned in 4.4.2.1, the negative values predominantly originated from measurements where the indoor field strength was measured at an upper floor of the building and the outdoor field strength at 1,5 m above ground level. Furthermore, reradiation effects also influenced the actual field-strength data. One may decide to truncate the building-effect distribution at the lower end at 0 dB. In that case $E_{i,\max}$ and $U_{\max}$ reduces from 85 V to $L_u E_{\max} = 8,9 \times 3 = 27 \text{ V}$.

Assuming that all 42 million outlets are located in reinforced concrete buildings the results represented by Curve 1 in figure 4.4.4-4b are modified into those represented by Curve 3, while Curve 4 follows if the L_i data are used. In practice, the outlets will be distributed over the B/W and C buildings. Then N_T in the calculations is the total number of outlets in a given type of building, and the results for both types of building have to be added.

NOTE 1 Although the number of outlets $N_o(U_h \geq U_L)$ as given in figure 4.4.4-4 may have noticeable values, the relative number is very low. This stresses the importance of the "tails" of the various distributions used and of the applied truncation. In addition, if the analytical expressions given in the annexes are not used, but a full-numerical calculation is made (for example, one starting with $n_m(E_o)$), care has to be taken that the accuracy of the numerical calculations is sufficiently high. The very low values of the relative number $N_o(U_h \geq U_L)/N_T$ form an indication of the accuracy needed. The relative number is identical to $pr\{U_h \geq U_L\}$ in per cent.

NOTE 2 In all calculations the density of the outlets μ was considered to be a constant over the whole respective region. The results can be improved when, for example, the calculations leading to $n_m(E_o)$, see figure 4.4.4-2, take a location-dependent μ into account.

NOTE 3 Curves 2 and 4 in figure 4.4.4-4b are based on L_i and A_b data, and a warning has to be given here. As mentioned in 4.4.2, it is not very likely that inside the buildings the far-field condition will be satisfied and, formally speaking L_i cannot be derived from G_i . In addition A_b has been determined from magnetic field-strength data and A_b for the electric field need not be the same. In the calculations leading to curves 2 and 4, it has tacitly been assumed that $L_i = G_i/Z_0$ ($Z_0 = 377 \Omega$). However, the same curves for $N_o(U_h \geq U_L)$ would have been found when correctly using the G_i and A_b , but quoting on Row 15 and 16 of table 4.4.4-1 the magnetic field quantities $H_{\max} = E_{\max}/Z_0$ etc. and replacing C_{Ei} by $C_{Hi} = C_{Ei}/Z_0^2$ (see clause 4.4.A-3 in annex 4.4-A).

4.4.4.2 Summary of disturbance source parameters

Assuming that the disturbance source in the conducted-immunity test is sufficiently described by an open-circuit voltage and an internal impedance, the following parameters are of importance.

The internal impedance

The internal impedance may be specified as a purely resistive quantity, of which the magnitude is chosen on the basis of the results for the equivalent asymmetrical resistance R_a , as given in 4.4.2.3. The choice depends on the reference for the asymmetrical disturbance source considered to be relevant in the situations where interference problems

have to be prevented. An often used value is $R_a = 150 \, \Omega$ [9] [10] [14] which is not in conflict with R_a results presented in 4.4.2.3.

The open-circuit voltage

Because the induced asymmetrical voltage has been measured at non-random locations in order to have a sufficiently large induced-signal-to-ambient-noise ratio, the bare U_h data cannot be used and the procedure described in 4.4.4.1 has to be followed. From that procedure, described in more detail in annexes 4.4-B and 4.4-C, it can be concluded that the following parameters have to be considered (see also table 4.4.4-1).

- a) N_T The total number of outlets in the geographical region (country) to be considered. N_T is either the grand total of outlets or the total number of outlets in a certain type of building (brick/wood or reinforced concrete).
- b) M_G or M_L The average value of the G or L factor, in $\text{dB}\Omega\text{m}$ or dBm , see tables 4.4.2-2 and 4.4.2-3. If the G_i or L_i factors are used, the following building effect parameters must be known (see tables 4.4.2-1 and 4.4.2-4):
 M_A : the average building effect A_b in dB
 S_A : the standard deviation of A_b in dB
 A_{bu} : the maximum building effect
 A_{bl} : the minimum building effect
- c) S_G or S_L The standard deviation (in dB) associated with M_G or M_L
- d) G_U, G_L The upper and lower limit of the G -factor range or the L -factor range
 L_U, L_L
- e) $H_{\max}$ or $E_{\max}$ The upper boundary of the outdoor field strength in A/m or V/m which determines the inner boundary of the areas around the transmitters to be considered.
This value may be chosen after considering radiation hazards. However, in particular in the case of mass-produced appliances the following consideration seems to be relevant. Choose (or agree on) a maximum field strength of X (A/m or V/m) such that there is a high probability that all equipment will be electromagnetically compatible when the field strength is equal to or smaller than X (A/m or V/m), and agree that special EMC hardening measures are to be taken in that part of the region where the field strength is larger than X (A/m or V/m) and (simultaneously) a complaint occurs.
- f) $H_{\min}$ or $E_{\min}$ The minimum field strength determining the outer boundary of the areas around the transmitters (see 4.4.4.1).
This value has only to be chosen when the field-strength distribution $n(H)$ or $n(E)$ is unknown. If this distribution is known, $H_{\min}$ or $E_{\min}$ is calculated from an equation like equation (4.4.4-3).
- g) $n(H)$ or $n(E)$ The field-strength distribution as discussed in 4.4.4.1 and annex 4.4-A.

4.4.5 Bibliography

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