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A NATIONAL MEASUREMENT GOOD PRACTICE GUIDE

No. 73

Calibration and use of antennas, focusing on EMC applications

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Measurement Good Practice Guide No. 73

Calibration and use of antennas, focusing on EMC applications.

M J Alexander, M J Salter, D G Gentle, D A Knight, B G Loader, K P Holland

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National Physical Laboratory, Teddington, UK

Abstract: This document addresses the calibration of antennas in the frequency range 30 Hz to 40 GHz. Guidance is given on the assessment of uncertainties in their use for radiated emission measurements according to EMC standards. The National Physical Laboratory operates calibration services for the measurement of antennas factors of monopole, loop, dipole, biconical, log-periodic and horn antennas which are used for emission testing on open area test sites and in fully anechoic chambers. Most of the calibration methods and the advice given applies to the use of antennas in other applications.

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National Physical Laboratory Teddington, Middlesex, United Kingdom, TW 11 OLW

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Approved on behalf of Managing Director, NPL by S Pollitt, Director of Enabling Metrology Division Contents:

| 1. | Scope | 1 |
|---------|---|-----------|
| 2. | Introduction | 1 |
| 3. | The role of a Standards Laboratory in EMC antenna calibration | 3 |
| 3 | 3.1. Measurement facilities | 4 |
| 3 | 3.2. Calculable antennas | 5 |
| 3 | 3.2.1. The calculable rectified standard dipole | 5 |
| 3 | 3.2.2. The calculable standard dipole. | 6 |
| 3 | 3.3. Electro-optic antennas | 9 |
| 4. | Uncertainties in the use of antennas | 9 |
| Z | 4.1. Antenna related uncertainties in emission testing | .11 |
| Z | 4.2. Uncertainties in the validation of test sites | .14 |
| | 4.2.1. Normalised site attenuation (NSA) | 15 |
| | 4.2.2. Site attenuation comparison method (SACM) | .16 |
| 5. | Problems encountered in calibrations for customers. | 17 |
| 4 | 5.1 Biconical antennas | 17 |
| 4 | 5.2 LPDA antennas | 20 |
| 4 | 5.3 Monopole antennas | 21 |
| 4 | 5.4 Connector problems | 21 |
| 6 | Self calibration using transfer standards | 24 |
| 0. 7 | Calibration interval | 25 |
| ,. 8 | Calibration test sites | 25 |
| υ. | R 1 Onen area test site (OATS) | 27 |
| Ş | Case study for design of an OATS | 27 |
| Ş | R_3 Anechoic chambers | 20 |
| Ċ | 8.3.1 Ferrite lined chamber 30 MHz to 1 GHz | 30 |
| | 8.3.2 Pyramidal absorber lined chamber 1 GHz to 40 GHz | 31 |
| Ş | | 33 |
| 9 | Outline of methods for antenna calibration | 35 |
| ٦. ر | Checks of condition of antenna | 37 |
| , C | 2. Calibration of dinole and biconical antennas | 38 |
| 5 | 9.2. Calibration of typeable dipole antennas | 30 |
| | 9.2.1. Calibration of higonical antennas | 10 |
| C | 9.2.2. Calibration of LPDA antennas | 40 |
| 5 | 2.1 Dhase centre of LDDA entennes | 42 |
| 5 | 2.2. Cross polar porformance of LDDA antennas | 43 |
| 5 | 9.3.2. Closs-polar periormance of LFDA antennas | .44 |
| 5 | 7.4. Calibration of nybrid biconical / LFDA antennas | .44 |
| 5 | 9.5. Calibration of entennes for ensitied measurement geometries | .43 |
| 5 | 9.0. Canonation of antennas for specified measurement geometries | .40 |
| | 9.0.1. Measurement distance. | .4/ |
| | 9.6.2. Uncertainty of measurement in the near-field | .48 |
| | 9.6.5. Customer expectations regarding calibration distance | .49 |
| | 9.6.4. Geometry-specific calibration for 5 in OATS measurement and the ADD | .30 |
| | 9.6.5. Calibration of antennas using 1 m separation - comment on the ARPS | /58 51 |
| r | calibration method | .51 |
| 2 | 9.7. Calibration of antennas for site evaluation | .33 |
| 2 | 9.8. Considerations on the measurement of radiation patterns of low gain antennas | .33 |
| ç | 9.9. Calibration of norm antennas | .33 |
| ~ | 9.9.1. Comment on the AINSI and AKP958 calibration methods | .02 |
| 9 | 9.10. Calibration of rod antennas | .03 |

| 9.10.1 | The equivalent capacitor substitution method | 65 | | | | |
|--|---|-----|--|--|--|--|
| 9.11. | Calibration of loop antennas | 67 | | | | |
| 9.11.1. | Passive loop antennas | 67 | | | | |
| 9.11.2 | Active loop antennas | 69 | | | | |
| 9.11.3 | Induction field method | 70 | | | | |
| 9.11.4 | TEM cell method | 71 | | | | |
| 10. Unc | ertainties | 72 | | | | |
| 10.1. | Uncertainties budget for a biconical antenna calibrated on an OATS | 73 | | | | |
| 10.2. | Uncertainty due to receiver noise | 74 | | | | |
| 10.3. | Uncertainties in chambers | 76 | | | | |
| 10.4. | Uncertainties budget for calibration of rod antennas in a GTEM cell | 78 | | | | |
| 10.5. | Interpretation of uncertainties in NPL EMC antenna certificates | 79 | | | | |
| 11. Ack | nowledgements | 80 | | | | |
| 12. Glos | ssary | 81 | | | | |
| 13. Refe | erences | 86 | | | | |
| Appendices | | 90 | | | | |
| Appendix | 1 Formulae | 90 | | | | |
| A1.1 | Antenna factor of half wave dipole | 90 | | | | |
| A1.2 | Antenna factor | 90 | | | | |
| A1.3 | Conversion of realised gain to antenna factor | 91 | | | | |
| A1.4 | Calculation of gain by the three-antenna method | 91 | | | | |
| A1.5 | Calculation of antenna factor by the standard site method | 91 | | | | |
| A1.6 | Calculation of antenna factor by the standard antenna method | 92 | | | | |
| A1.7 | Site insertion loss | 92 | | | | |
| A1.8 | E-field from an elemental dipole | 92 | | | | |
| A1.9 | Theoretical value of normalised site attenuation in free-space | 93 | | | | |
| A1.10 | Derivation of H-field from a measurement of E-field | 93 | | | | |
| Appendix | 2 Measurement of field-strength in the near-field of an EUT | 94 | | | | |
| A2.1 | Measurement of absolute field strength | 94 | | | | |
| A2.2 | Calibration with 1 m separation according to SAE ARP958 | | | | | |
| Appendix | 3 Information on uncertainties in using biconical antennas. | | | | | |
| A3.1 Ant | enna Factor | | | | | |
| A3 2 Bal | ance Test | 98 | | | | |
| A3 3 Ret | urn Loss | 98 | | | | |
| A34 AR | P958 Antenna Factor | 98 | | | | |
| A35AN | SI Height Scan Method | 99 | | | | |
| Appendix | x_4 information on uncertainties in using LPDA antennas | 100 | | | | |
| A41 A | Interna Factor | 100 | | | | |
| A4 2 P | hase Centre | 100 | | | | |
| A4 2 1 | Assessment of uncertainty arising from phase centre position on LPDAs | 102 | | | | |
| A43 R | eturn Loss | 103 | | | | |
| A4.5 Return Loss | | | | | | |
| A45 Use of ARP958 Antenna Factor | | | | | | |
| A4 6 ANSI Height Scan Method | | | | | | |
| $\Delta 4.7$ Uncharacteristic resonances | | | | | | |
| A4 8 Mutual counting between large horn antennas 1 m apart | | | | | | |
| Annendi | 5 Emission measurements in screened rooms | 105 | | | | |
| Annondi | Antenna calibration services offered by NDI | 107 | | | | |
| Appendix | Antenna canoration services oncred by Nr L | 10/ | | | | |



Plate 1: 60 m x 30 m sheet steel ground plane on the NPL site



Plate 2: Ensemble of EMC antennas: Biconical, LPDA, bilog, rod, standard dipole.

1. Scope

This Good Practice Guide addresses the calibration of antennas in the frequency range 30 Hz to 40 GHz. Guidance is given on the assessment of uncertainties in their use for *EMC* radiated emission measurements and evaluation of test sites. It is not intended to give detailed methods of calibration, which can be found in textbooks or in published standards. Knowledge of these is useful in understanding the variations of these methods and their limitations as described in this Guide. Most of the information in this Guide is applicable to antennas used outside the field of EMC. NPL operates a calibration service for the measurement of *antenna factors* of dipole, biconical, log-periodic (*LPDA*), *broadband hybrid* and horn antennas which are used for emission testing on *open area test sites* (OATS) and in *anechoic chambers*. Words in italics (first instance) are explained in the Glossary.

Conical log-spiral, monopole (or rod) and loop antennas, which are mainly called for by military standards, are also calibrated at NPL. The use of antennas at a distance of 1 m from the emitter in screened rooms is a special case, for which calibrations and measurement uncertainties are addressed.

This Guide is aimed at technical users of antennas who wish to minimise the uncertainties of their radiated emission measurements. Typical uncertainties in the use of EMC antennas are included in the text. Allied services at NPL handle calibrations of field probes over the frequency range 10 Hz to 46.5 GHz, which are used in part of this frequency range to test for field uniformity in EMC immunity tests. A separate Good Practice Guide has been written about field probes¹.

This document is an updated version of the NPL Measurement Good Practice Guide No 4 "Calibration and use of EMC antennas" published in April 1997, in which sections on horn and rod antennas have been expanded, and more detail has been given of calibration methods and uncertainties. Sections on TEM cells, the calibration of loop antennas and on designing a ground plane have been added.

2. Introduction

Measurement of radiated emissions above 30 MHz is a requirement of Electromagnetic Compatibility (EMC) standards and of the European Community EMC Directive². Other standards require electric and magnetic field measurements to be made as low in frequency as 30 Hz. EMC is the ability of electrical equipment to co-exist without mutual electromagnetic interference, and it protects the users of broadcast services from undue radio interference. Improving the methods of calibrating antennas will make possible the reduction of *uncertainty* of EMC measurements, which will enable industry to make savings in the design and manufacture of products, and yet achieve the required EMC protection.

One component of the uncertainty budget for an emission test is the performance of the test site as gauged by the measurement of *Normalised Site Attenuation (NSA)* using a pair of antennas. There is a demand for highly accurate site attenuation (SA) measurements because a small shortfall in meeting the acceptability criterion can result in costly modifications to the site.

Calculations of uncertainty of measurement are required in test reports produced by EMC laboratories that are accredited under an internationally accepted accreditation scheme. This requirement is spelled out in ISO 17025³. Uncertainty calculations should be performed in accordance with the requirements in the ISO/IEC Guide to the Uncertainty of Measurement⁴. The UK Accreditation Service, *UKAS*, has published document LAB34⁵ which gives guidance on the calculation of uncertainty in the specific case of EMC, in which the magnitudes of uncertainty are typically larger than in most applications in RF & Microwave metrology. An example of an uncertainty budget for the NPL calculable standard dipole antenna is given in Ref. 6.

CISPR 16-4, parts 1 to 4, gives a good overview of EMC uncertainties⁷. It is hoped that, as the contributions to the uncertainty of emission measurements become better understood and quantified, the staff of more laboratories will be encouraged to include all known contributions in their uncertainty budgets. This will help to ensure that trading partners are offering their services on equal terms.

When assembling an uncertainty budget for an emission test, the antenna contribution is not merely the uncertainty of measurement of antenna factor stated in the certificate of calibration. This could be as low as ± 0.2 dB. Because of the interaction of the antenna with the emission test environment there may be other uncertainty contributions which amount to several decibels. Whilst some systematic errors are amenable to correction, others would require costly characterisation of the antenna and are usually left as uncertainty components. In practice the total uncertainty contribution of an antenna in an EMC emission test can economically be kept to less than ± 2 dB. A detailed calibration certificate, such at that from NPL, quantifies the main contributions.

Some standards specify the calibration of antennas with a separation of 3 m or 1 m between the source and receiving antennas; a problem with calibrating antennas-in-close-proximity by the three-antenna method is that mutual-coupling between two antennas is built into the antenna factor, yet the reason for the calibration is for compliance testing of an EUT in which the antenna is actually coupling to the EUT. The effect of the change of coupling has to be allowed for in the uncertainty of measurement.

There are many proprietary models of antenna that have undesirable, but avoidable, features in addition to the less than ideal intrinsic properties referred to above. These are *balunimbalance* in *tuneable* dipole and biconical antennas, and resonances in LPDAs. These properties can be unstable, depending on cable layout and usage of the antenna. NPL has developed tests to determine approximately the additional uncertainties that such antennas will introduce when used for emission testing. Fortunately there are commercial models available which are largely free of such defects. It is to be hoped that, as the EMC community becomes aware of the disadvantages of using unsatisfactory antennas, they will be phased out of use.

Test houses can obtain traceability of antenna factor to national standards by self-calibrating an antenna through substitution against a transfer standard which has been calibrated at an accredited laboratory. However the test house needs to be aware of the requirements for a high quality test site and of the pitfalls they may encounter in the calibration of antennas: for traceability to be claimed it is necessary to have the calibration method approved by a national accreditation body, such as *UKAS* in the UK. This is explained more fully in Section 6.

This Guide is an example of technology transfer and good practice support provided by NPL. It is one of a series⁸ devoted to practical metrology techniques, commissioned by the National Measurement System Policy Unit (NMSPU) of the Department of Trade and Industry (DTI). It provides a fitting introduction to the NPL calibration and testing facility that includes one of the world's best VHF antenna ranges, comprising a 60 m by 30 m contiguously welded sheet steel ground plane. The lead author is a member of the CISPR/A group working on a document for methods of antenna calibration and site evaluation in the frequency range 30 MHz to 18 GHz. This guide is consistent with the CISPR document, which is expected to be published in 2005.

3. The role of a Standards Laboratory in EMC antenna calibration

The role of a *National Measurement Institute (NMI)* in the reduction of uncertainties in radiated emission testing is twofold: firstly, to develop a method for measuring an optimum value of antenna factor at each frequency that will result in the minimum uncertainty of measurement in an emission test, and secondly, to evaluate the properties of the antennas to establish the uncertainties attributable to each characteristic that causes the antenna factor to vary from its ideal value during an emission test. The ideal value is that which gives the true E-field strength from the measured output voltage of the antenna. The definition of antenna factor and its relationship to antenna gain is given in Appendix 1. The evaluation of antenna characteristic. Unrealistically high antenna related uncertainties would unacceptably increase the overall uncertainty of an emission test. This is discussed more fully in Section 4.

An NMI would normally be accredited to ISO17025, but in any case be expert at identifying contributors to the uncertainty of measurement and calculating the overall uncertainty. Furthermore an NMI will participate in international inter-comparisons⁹ with peer NMIs in order to confirm the validity of their uncertainty assessments. The participation by countries in international Mutual Recognition Agreements, aided by inter-comparisons, in a particular subject makes it easier for one country to recognise a calibration performed by another country.

Two distinct methods of EMC testing are: 1) on a 10 m OATS employing height scanning of the receive antenna and 2) in a fully lined anechoic chamber. Some anechoic chambers have a ground plane on the floor instead of absorber - for the purposes of this article such facilities will also be treated in the same way as an OATS. NPL has installed specially designed facilities to enable measurements of the highest accuracy to be made. NPL is examining improved methods of emission testing and this includes investigation of testing in absorber lined rooms. The appropriate calibration of antennas for use in fully anechoic rooms is a measurement of their *free-space* AF, which is the default CISPR parameter for antennas.

3.1. Measurement facilities

OATS which are designed for radiated emission conformance testing may not meet the stringent requirements required for antenna calibration. An antenna range specifically for the calibration of EMC antennas was built at NPL, the UK's national standards laboratory, see Plate 1. This site is known to approximate very closely to the ideal site and can be regarded as a National Standard Site, against which measurements on other sites can be compared. Its performance is described in Section 8.1. It is preferred to call such a site a *CALTS* (CALibration Test Site) to distinguish it from the CISPR OATS whose quality criterion is a relaxed ± 4 dB (clause 5.6 of Ref.10) for the difference between measured and theoretical *site insertion loss (SIL)*. CISPR introduced a specification for a CALTS¹¹ in 1999.

A fully anechoic chamber was built at NPL in the year 2000 to conform with EN50147-2 as best as the suppliers could achieve; special pyramidal absorber was commissioned with the tips neither truncated nor painted, in order not to degrade the absorption up to 18 GHz. See Plate 2. The 9 m x 6 m x 6 m chamber is fully lined with ferrite tiles, on top of which is placed "hybrid" pyramidal urethane foam absorber up to 36 inch in depth, designed to achieve low reflections in the frequency range 30 MHz to 2 GHz and acceptable performance up to 18 GHz. The chamber is primarily used for the measurement of *free-space antenna factor (AFfs)* of wire antennas up to 5 GHz.

3.2. Calculable antennas

Fortunately, the theory of dipole antennas has been analysed in great depth. In the course of this work we found remarkable agreement between the predicted antenna factor and the measured antenna factor, which was derived from the coupling between a pair of antennas. The antenna factor was calculated both by analytical formulae and by numerical computation involving the method of moments. This is described in Ref. 12. After this paper was written it was pointed out that the analytical formulae are strictly valid for very thin wires only; including the radius involves an approximation. The agreement of ± 0.3 dB cited in the paper between analytical and numerical methods is, in fact, improved to better than ± 0.05 dB for very thin resonant dipole antennas. For wires with a realistic radius the numerical methods agreed with measurement to better than ± 0.2 dB for SIL, showing the method of moments copes well with thick wires. Also numerical methods accurately predict the antenna factor of a single dipole used over a broad bandwidth. For a pair of identical antennas, an agreement of \pm 0.2 dB for SIL implies an agreement of better than \pm 0.1 dB for AF of each dipole, because the site error is also included in the value for SIL. This achievement was a significant improvement on the state-of-the-art at the time, enabling an improvement in the calibration of low gain wire antennas in the VHF and UHF ranges from typically $\pm 1 \text{ dB}$ to typically ± 0.3 dB, and ± 0.15 dB best case. Up until this point the version of standard dipole in use was a thin wire with a Schottky diode connected across the gap in the centre of the dipole, described further in Section 3.2.1.

The ability to accurately measure antenna factor, verified by theory, underpins a wide variety of developments in antenna metrology. Advances made possible by precision measurements include the validation of calibration methods, the validation of theoretical models, the ability to set up a precise field strength, the measurement of the properties of antennas, the measurement of the effects of antenna support structures and feed cables and the evaluation of EMC open area test sites and anechoic chambers. The complex reflection coefficient of the ground can be measured, and also of RF absorbing material laid on the ground. The quality of the national standard ground plane has been verified: besides enabling accurate calibrations of EMC antennas, accurate measurements of free-space gain of circularly polarised VHF satellite antennas can be made by the three-antenna method above the ground plane (because the ground reflection undergoes a change in the sense of polarisation it is not seen by the receiving antenna).

3.2.1. The calculable rectified standard dipole

A thin wire with a diode at its centre is conceptually very attractive. This is because of the beauty of the formula for calculating antenna factor and because the dipole is insensitive to mutual coupling with its surroundings because of its very high self-impedance. The diode is connected to a voltmeter via high impedance leads, voltage being a parameter that can be measured very precisely. The main drawbacks of this antenna are its relative insensitivity and it is not selective: in regions where the regulatory authorities impose limits on the power that can be transmitted it is not permissible to generate a field needed to give about 1 volt across the diode. Also this dipole will measure the resultant field of all frequencies present, so, for example, a large ambient field from a TV broadcast cannot be excluded. In essence the antenna factor of a thin resonant high impedance dipole is given by:

$$AF = \frac{\pi}{\lambda}$$

where λ is the wavelength.

A working formula for a practical dipole with finite thickness is given in Ref. 50.

3.2.2. The calculable standard dipole

The calculable standard dipole consists of a set of dipole elements and a balun, as shown in Fig. 1.



Figure 1. Schematic diagram of a calculable standard dipole antenna.

The NPL dipole is constructed in the following manner. The dipole elements have two dipole arms fitted to a central dielectric support, with push fit connectors for connection to the balun. The balun consists of a 180° hybrid coupler that has a phase balance better than 1.5° and an amplitude balance of better than 0.2 dB over the frequency range 30 MHz to 1 GHz. The sum port (Σ) of this hybrid is terminated with a 50 Ω load. A matching pad is connected to each of the balanced ports of the hybrid. Two 50 Ω semi-rigid coaxial cables, phase matched to within 1°, connect the hybrid to the dipole element. At the dipole end, the shields of these cables are joined electrically together. The design of the balun allows for easy measurement of its complex S-parameters using an automated network analyser (ANA), and for the antenna elements to be changed quickly during site measurements without the need to reposition the antennas.

The derivation of the theoretical value for the site insertion loss of the calculable dipoles is presented in Refs. 12 and 13. SIL is obtained by combining a numerical model for the dipole elements with measured data on the attenuation of the two baluns. The dipole elements are modelled using the Numerical Electromagnetic Code¹⁴ (NEC) to determine an equivalent 2-port network for the site. In the model the ground plane is infinitely large and perfectly conducting. The dipole is modelled as a wire with 31 segments. A 1 V source is applied to the centre of the transmit wire and a 100 Ω load is placed at the centre of the receive wire. The gap at the antenna feed points and the dielectric supports are not included. NEC is used to calculate the input impedance of the transmit antenna and the complex current in the load.



Figure 2. Calculation of theoretical site insertion loss

By measuring the S-parameters of each balun, the scattering matrix of their equivalent 2-ports is calculated. The transmission coefficient (S_{21}) of the cascade combination of the 2-ports for the transmit balun, site and receive balun is calculated. The derivation of the theoretical value for SIL is shown schematically in the diagram, Fig. 2.

The calculable dipole has been miniaturised and the frequency extended to 2 GHz. The trickiest part was to reduce the gap at the feed point of the pair of dipole elements and to reduce the bulk and effect of the dipole supporting material. Modelling showed that the effects of the gap and the dielectric material tended to cancel, which was a bonus¹³. This dipole covers the frequency range 850 MHz to 2.2 GHz which is ideal as a standard for antenna gain and field strength for mobile telecommunications. Above 2 GHz it is more appropriate to use horn antennas as gain standards as their gain can be measured more accurately, see Section 9.9. Certainly below 1 GHz the size of horns can become unwieldy, but their great advantage from a calibration point of view is that they are not omnidirectional, making it easier to reduce reflections from the surroundings.

The useable bandwidth of the calculable dipole antenna can be greatly extended by methods of moments calculations, compared to limitations inherent in the approximations of analytical methods. NEC was used to compute antenna factor across bandwidths exceeding 200% with an increase in antenna factor uncertainty to less than ± 0.3 dB at the ends of the band. The limit on the bandwidth was to avoid excessively high antenna factors at the ends of the bands. In order to cover the frequency range 30 MHz to 1 GHz, four dipole lengths were selected, 60 MHz, 180 MHz, 400 MHz and 1 GHz, whose AFs are shown in Fig. 3.



Antenna factors for 4 dipoles

Figure 3. Antenna factors for set of 4 broadband calculable dipole antennas.

3.3. Electro-optic antennas

Wire antennas, of the size that is convenient to handle and transport, at low frequencies, tend to be omni-directional. The advantage of using a highly directional horn antenna, which is possible at UHF and above, is that it is minimally affected by the antenna mounting arrangement and the waveguide or coaxial cable feeding the antenna. In contrast, these factors, including the nature of the ground, are of major significance for omni-directional antennas. The ground problem can be solved by a high quality ground plane that enables the magnitude and phase of the ground reflection to be known precisely; the antenna supports can be made of foam or thin fibreglass tube to reduce the reflections; but the feed cable is made of metal and can be the largest source of uncertainty. The effect of the cable can be minimised by routing it everywhere orthogonal to the dipole element, which is straightforward when the antenna is horizontally polarised because the cable naturally drops to the ground, ideally to a bulkhead connector so that it continues its journey underground to the receiver. But when the antenna is vertically polarised the cable has to be of the order of 10 m away for its effect to dwindle to a fraction of a decibel. When performing calibrations that require height scanning it is not very practicable to lead the cable away horizontally behind the antenna for several metres before dropping to ground.

A solution is to feed the antenna with an optical fibre to a small battery powered optical-to-RF converter at the input to the antenna. Alternatively the antenna could be entirely electrooptical, for example depositing a small metal dipole on a lithium niobate crystal which is fed by laser light to operate as a Mach-Zehnder interferometer¹⁵. There is much less attenuation down an optical fibre than down a coaxial cable, which is noticeable above 1 GHz for the lengths of cable used on a CALTS. Furthermore an electro-optic antenna can be small and cause minimal disturbance to the field it is measuring. It is suitable for use as a transfer standard to transfer antenna factor obtained against a calculable dipole antenna on a CALTS, to a TEM cell, for comparison with the field in the TEM cell derived from the power input to the cell and the cell dimensions.

An example of an optically driven radiating dipole is given in Ref. 16. There are various RF analogue to optical links on the market which can simply be connected to the antenna output. The electronic box will need to contain a battery, because the purpose of removing wires will be lost if a power lead is required. If the box is not sufficiently compact, a short length of coaxial cable can be inserted in order to displace the box far enough behind the antenna, so as not to influence the field being measured.

4. Uncertainties in the use of antennas

Emission testing on OATS is performed to a prescription laid down by CISPR or other standards body. The receiving antenna is set 10 m away from the equipment under test (EUT), however the *FCC* was the first to accept a distance of 3 m and this has become widespread practice. The antenna is scanned in height in order to avoid destructive interference between the direct and ground reflected signals. The maximum signal in the height scan range 1 m to 4 m is noted. This method was adopted because of the omnidirectional properties of practical antennas for the frequency range 30 MHz to 1 GHz. In this case a well-defined metal ground plane makes the measurement more reproducible, when the alternative might have been reflections from a car park surface or a grass field. However,

from a metrology point of view even a well-defined metal ground plane is the cause of several sources of uncertainty.

If the intention of an emission measurement is to measure the emission directly from an EUT, metrologically the CISPR method¹⁷ introduces two error terms which are not recorded in the uncertainty budget. The more important of these two errors arises from the ground reflection. The antenna has to be scanned in height (usually 1 m to 4 m) to avoid destructive interference of direct ray and the ground reflected ray. When they combine in-phase to give a signal maximum the direct signal is increased by between 4 dB and 5.8 dB (this cannot be classified as an error because it is an artefact of the test method, in contrast to free-space testing in a *FAR* [fully anechoic room] which would not have this increase). However an inconsistency arises at 30 MHz where a signal maximum cannot be obtained in the limited height scan range for horizontal polarisation: whereas over most of the frequency range the signal is approximately 5 dB higher, the reading at 30 MHz is 12 dB down from this general higher value. The second error is that the distance between the EUT and the height-scanned antenna is greater than the specified distance. The distance error is significant for a 3 m site and is elaborated in Section 9.6. These error terms should have been included in the setting of the specification limits, but it is doubtful that they were rigorously taken into account. A reminder of this problem helps the argument for the adoption of FARs.

In order to reduce the uncertainty contribution of the antenna to the measured field strength it is necessary to understand the emission test method and the properties of the antenna. A good introduction to awareness of these issues is given by Dvorak in his paper¹⁸ and at workshops at the EMC Zurich conference from 1989. Strictly, the antenna factor is valid only if the field radiated by the EUT closely duplicates the (usually uniform, linearly horizontally polarized) field in which the antenna has been calibrated. If the field radiated by the EUT on an OATS differs from the calibration field, the antenna factor will be modified. Whether this modification will be substantial or negligible depends on many parameters as well as on the way they mutually interact. Larger deviations may be expected, among others, for non-uniform fields having a vertical component or a component in the direction of propagation, for non-uniformly illuminated high gain (LPDAs) or broadband measuring antennas with poor cross-polar rejection, for conducting antenna cables, etc.

The height above ground of the maximum signal will vary with frequency. There are three properties of the antenna which affect its output voltage, in response to a field at a predefined distance from the EUT. First, the antenna factor varies with height. Second, the radiation pattern affects the proportions of signal received directly from the EUT and via reflection from the ground plane. Also the radiation pattern of some broadband horn antennas are very non-uniform in their upper frequency range. Third, the phase centre of LPDAs changes with frequency.

The output voltage of an antenna, of given characteristic impedance, is related to the strength of the field in which it is immersed by the antenna factor (AF). When calibrating the antenna one could argue that its gain, or antenna factor, should be measured for every height maximum at every frequency and polarisation direction (see *boresight*) because no two antenna factors will be the same. This would enable a systematic correction for these effects, which would result in lower measurement uncertainties. However this could be costly. The more practicable solution is to measure an "optimum" antenna factor and to account for the variations caused by these three properties as uncertainty terms. The optimum AF is a single value to be used regardless of measurement geometry, including phase centre. In practice the

optimum AF is the free-space antenna factor or AFfs. By default EMC antennas are calibrated in their boresight direction. However in the case of height scanning from 1 - 4 m above a ground plane with a separation of 3 m from the EUT, it can be more accurate to use a *geometry specific AF*, measured with the same height scanning and separation geometry, see Section 9.6.

4.1. Antenna related uncertainties in emission testing

Antennas measured by a laboratory accredited to perform antenna calibrations will have the uncertainty of the antenna factor quoted in the certificate of calibration. Additionally there are uncertainties associated with the method used for emission testing. These have been extensively researched at NPL and techniques have been developed for calibrating antennas that give the lowest overall measurement uncertainty for a given method of emission test.

Table 1 shows an example uncertainty budget for a radiated disturbance measurement. This table is based on a table in CISPR 16-4-2⁷, which itself was based on a table in the previous edition of UKAS document LAB34⁵, namely UKAS NIS81. Such uncertainty budgets are popularly used to calculate the uncertainty of an EMC test, but the reader should bear in mind that the unfixed nature of many products being tested, especially those with cables attached, can result in far higher uncertainties, and other factors such as operator experience and mistakes have not been included in Table 1. It is worth reading about EMC Compliance Uncertainty¹⁹ in CISPR 16-4-1, which considers these other influence quantities, for a fuller discussion on the scope of the uncertainties of EMC testing.

Following the guidance in the GUM^4 the measurand *E* is calculated as:

$$\begin{split} E &= V_r + L_c + AF + \\ &+ \delta V_{sw} + \delta V_{pa} + \delta V_{pr} + \delta V_{nf} + \delta A + \delta AF_f + \delta AF_h + \delta A_{dir} + \delta A_{ph} + \delta A_{cp} + \delta A_{bal} + \delta SA + \delta d + \delta h \end{split}$$

where the terms are defined in Table 1 as Input Quantities.

The three variable properties of an antenna, namely mutual coupling, radiation pattern and phase centre, are quantified as uncertainty terms that are additional to the uncertainty given foremost in the calibration certificate for antenna factor. Mutual coupling of the antenna to the ground plane implies a variation of antenna factor with height in the height scan range 1 m to 4 m. This can be as much as 6 dB for a horizontally polarised *resonant dipole* antenna at 30 MHz, see Fig. 4. The effect is much smaller for vertical polarisation and can generally be ignored for heights above 0.4 wavelengths, see Fig. 5. For a horizontally polarised biconical antenna it is typically 1.8 dB above 55 MHz, see Fig. 19. Notice that below about 50 MHz the antenna factor changes very little with height because the antenna has become a short dipole with a high self-impedance, compared with which the mutual impedance is negligible. An LPDA antenna is directive, and the fact of the elements being embedded in an array desensitises them to mutual coupling; even for horizontal polarisation the deviation is less than ± 0.3 dB, for an LPDA with lowest design frequency of 200 MHz.

| Input Quantity | Input Quantity X_i Uncertainty of x_I | | ertainty of x_I | $u(x_i)$ |
|-------------------------------------|---|-----------|-----------------------|----------|
| | | dB | Prob. Dist.; <i>k</i> | dB |
| Receiver reading | V_r | ±0.1 | <i>k</i> =1 | 0.10 |
| Cable attenuation: antenna-receiver | L_c | ±0.1 | <i>k</i> =2 | 0.05 |
| Receiver corrections: | | | | |
| Sine wave voltage | δV_{sw} | ±1.0 | <i>k</i> =2 | 0.50 |
| Pulse amplitude response | δV_{pa} | ±1.5 | rectangular | 0.87 |
| Pulse repetition rate response | δV_{pr} | ±1.5 | rectangular | 0.87 |
| Noise floor proximity | δV_{nf} | ±0.5 | <i>k</i> =2 | 0.25 |
| Mismatch: antenna-receiver | δМ | +0.9/- | U-shaped | 0.67 |
| | | 1.0 | | |
| Log-periodic antenna factor | AF | ±2.0 | <i>k</i> =2 | 1.00 |
| Log-periodic antenna corrections: | | | | |
| AF frequency interpolation | δAF_f | ±0.3 | rectangular | 0.17 |
| AF height deviations | δAF_h | ±0.3 | rectangular | 0.17 |
| Directivity difference | δA_{dir} | +1.0/- | rectangular | 0.29 |
| | | 0.0 | | |
| Phase centre location | δA_{ph} | ±0.3 | rectangular | 0.17 |
| Cross-polarisation | δA_{cp} | ±0.9 | rectangular | 0.52 |
| Balance | δA_{bal} | ±0.0 | | 0.00 |
| Site corrections: | | | | |
| Site imperfections | δSA | ± 4.0 | triangular | 1.63 |
| Separation distance | δd | ±0.1 | rectangular | 0.06 |
| EUT table height | δh | ±0.1 | <i>k</i> =2 | 0.05 |

Table 1 – Example uncertainty budget for EMC radiated emission test from 200 MHz to 1 GHz using a horizontally polarised LPDA antenna at a distance of 10 m.

k=1 and k=2 are normal distributions. The expanded uncertainty is $2u_c(E) = 5.06$ dB, where $u_c(E)$ is the combined standard uncertainty, which is the root sum of squares of the values of $u(x_i)$ in the last column. It has been assumed that the sensitivity coefficient is unity for all the components.

There are other possible error contributions by the measuring receiver, such as input impedance variations, and spurious responses that may become important when measuring harmonics of ISM (industrial, scientific and medical) equipment or broadband sources.



Figure 4. Variation of AF with height for a horizontally polarised resonant dipole antenna



Figure 5. Variation of AF with height for a vertically polarised resonant dipole antenna

The second uncertainty term arises from the directivity of the antenna. Antenna factor is measured for the boresight direction of the antenna. When a vertically polarised dipole or biconical antenna is scanned in height above a ground plane the received signal is reduced because the EUT is no longer in the boresight direction. This is especially the case for an

LPDA (polarisation vertical or horizontal) for which this uncertainty term can be as much as + 2 dB on a 3 m OATS, i.e. the measured signal is 2 dB less than would be measured by a dipole antenna.

The third uncertainty term arises from the variation of phase centre with frequency of a LPDA. The LPDA has been numerically modelled in order to assess this uncertainty component²⁰. This term is more significant when the antenna is used over a ground plane because the signal strength of the combined in-phase direct and reflected rays varies more than for a single ray in free-space. The method for correcting for phase centre in such conditions is given in Appendix 4. In free-space the uncertainty for a LPDA of length 1.4 m is ± 2 dB at the ends of its frequency band on a 3 m range, assuming the phase centre is fixed at the mechanical centre of the antenna. The uncertainty is reduced to ± 0.6 dB for a 10 m range. See Section 9.4 for the calculation of this uncertainty for a 1 m long broadband hybrid antenna – most LPDA antennas (200 MHz – 1 GHz) used for EMC testing are around 0.65m long. The uncertainties are greater when the measurements are made over a ground plane.

LPDA antennas can exhibit cross-polar rejection as low as -14 dB with respect to the co-polar level. This is caused by the quarter wave elements either side of the boom not being co-linear. See section 9.3 for more discussion on this issue. An LPDA illuminated by equal field strengths in horizontal and vertical polarisation (i.e. a field at 45°) will be measuring the co-polar field with an error of 0.9 dB if the cross-polar rejection of the LPDA is 20 dB. Clause 4.4.3 of CISPR 16-1-4¹⁰ states that if the cross-polar rejection is less than 20 dB the uncertainty must be calculated.

There are further uncertainty terms associated with the design of antennas, which can be as great as ± 15 dB. The problem of balun imbalance has caught many operators unaware, wasting man-days of effort, especially when using such antennas for site validation. Some models of tuneable dipole and biconical antennas have poorly balanced baluns. Some LPDAs have uncharacteristic resonances. Some antennas have poor quality input connectors that result in repeatability issues, particularly antennas that are designed to operate to 18 GHz. These problems are described in Section 5.

4.2. Uncertainties in the validation of test sites

An OATS or an anechoic chamber should be as close to its ideal behaviour as possible, to minimise its contribution to the measurement uncertainty of an EMC emissions test. An ideal OATS has an infinite area, is perfectly flat and perfectly conducting and the hemisphere above it has no reflecting obstacles. An ideal FAR is like free-space with no reflections coming from any direction. The validation measurement is a test of how close to the ideal is the actual behaviour of the site. There are two widely used methods for validating sites. The first is the measurement of site attenuation that is normalised by the antenna factors, see section 4.2.1, and the second is the measurement of site attenuation on a *reference test site*, or a CALTS, followed by a repeat measurement on the site-being-validated, see section 4.2.2. The second method is more accurate but relies on access to a reference test site, which is a high quality site of the sort that an NMI or a calibration laboratory might have.

A third and the most accurate method is to measure the site insertion loss (SIL) between a pair of calculable dipole antennas²¹ set at a fixed height, i.e. not using height scanning; see Fig. 6. Because the SIL is calculable, no reference site is required, removing the uncertainty

components associated with this. If an OATS or SAC (semi-anechoic chamber) is being evaluated, the heights of the antennas have to be chosen for a given frequency such that there is not destructive interference between the direct and ground reflected rays. However it is desirable to evaluate a site by swept frequency measurement with small frequency increments. It is possible to get accurate results for the calculable dipole over a broad band with numerical software: the process is accurate for a given fixed height of the two antennas over a bandwidth such that the signal does not drop below the in-phase maximum by more than a few decibels.



Figure 6. Schematic layout for site attenuation measurement

A horizontally polarised (HP) ray undergoes 180° change on reflection, whereas the phase of a vertically polarised (VP) ray is unchanged. In order to locate the signal maximum, one antenna is scanned in height: it will be noticed that the frequency at which a signal minimum occurs for HP is near to the frequency at which a signal maximum occurs for VP, and vice versa.

4.2.1. Normalised site attenuation (NSA)

Clause 5.6 of CISPR 16-1-4 describes¹⁰ the method of validation of a compliance test site (COMTS) by measuring the normalized site attenuation (NSA) for both horizontal and vertical polarizations. Unless otherwise specified by relevant standards, free-space antenna factor shall be employed in the NSA measurements regardless of the polarization and separation distance between the transmitting and receiving antennas. This is appropriate for test sites that are intended to provide a free-space environment, such as a FAR. However if the antennas are not far enough apart for mutual coupling between them to be insignificant, corrections to compensate for mutual coupling have to be made, or else the uncertainty increased according to the estimated magnitude of the mutual coupling. In the case of an

OATS (or SAC), corrections will also be needed to compensate for the effects of mutual coupling of the antennas to their images in the ground plane, or the uncertainty increased appropriately.

The uncertainties of Section 4.1 are potentially doubled when doing NSA measurements because the uncertainties in the antenna factors of <u>two</u> antennas have to be accounted for. A traditional method for validating EMC test sites is the measurement of site attenuation and then the calculation of normalised site attenuation, which is compared with a calculated ideal value. Site attenuation is the measurement of insertion loss between two antennas with a given separation where one antenna is fixed in height above a ground plane and the other is scanned in height between 1 m and 4 m in order to measure the maximum signal at each frequency. This is normalised by subtracting the antenna factors of the two antennas. A theoretical value of NSA is calculated between a pair of half-wave dipole antennas using software that simulates the height scanning. The difference between the measured NSA and the theoretical NSA is compared with a criterion for site acceptability. In CISPR 16 for many years this criterion has been agreement within ± 4 dB.

The method of measuring NSA related to Table 2 of CISPR 16-1-4 requires that one antenna be placed at a fixed height of 1 m and the other is height scanned from 1 m to 4 m. Calibrating the fixed antenna for its particular height and polarisation is possible, as described in Section 9.2.1, so that the uncertainty term for antenna factor variation with height for this antenna is eliminated. For vertically polarised NSA measurements, uncertainty contributions from cable reflections of around ± 1 dB per antenna can be reduced substantially by routing the cable horizontally behind the antenna for several metres before dropping vertically to ground, shown schematically in Fig. 6.

The calibration of antennas by the *SSM (standard site method)*, as used in ANSI C63.5²², using an antenna separation of 3 m is discussed in Section 9.6.4. There are various sources of error if the AF were to be regarded as AFfs, but for the purpose of site evaluation the NSA measurement method is a copy of the calibration method, especially if exactly the same antennas are used, in which case the errors will cancel out. So for the evaluation of a 3 m OATS, it is actually more accurate to use AF as measured by the SSM at 3 m rather than AFfs. The whole process boils down to a direct comparison of the quality of the site on which the antennas were "calibrated" with the quality of the site being evaluated. This means that we can dispense altogether with NSA (and calibration of antennas suitable for NSA measurements) and simply measure the site insertion loss on a good quality site, being the reference between the two measurements. This difference is compared directly to the site evaluation criterion. This method is described in the next Section 4.2.2.

4.2.2. Site attenuation comparison method (SACM)

As introduced in the previous section, the whole process of calibrating antennas by the SSM and using the consequent AFs to derive NSA from a measurement of SA, which itself uses the same height scanning geometry of the SSM, is a circular process. Essentially what is happening is that the site being evaluated is being compared with the site on which the antenna calibrations were performed (the CALTS). So why not make a direct comparison by simply measuring SA on the CALTS, repeating the measurement on the site-being-evaluated, and take the difference between the two results? The difference between these two measurements can be compared to the site acceptance criterion, for example the ± 4 dB used

for the NSA method. This method is termed the "Site attenuation comparison method" or SACM. The SACM does not require the knowledge of the antenna factors of the two antennas, hence the elimination of these uncertainty components. The SACM is more exact than the NSA method, because the calibration of the antennas ignores the fact that one of the antennas in the three-antenna-method is at two different heights for two of the antenna pairings, thereby imparting an error to the AF of all three antennas.

The SACM assumes that SA is measured on the CALTS and the site-being-evaluated using the same pair of antennas. Advocates of the NSA method may counter that they already have calibrated antennas and should not need to send a pair of antennas to a CALTS for a special SA measurement to be made. Adherents also argue that some tests laboratories may not have access to a CALTS, or would rather calibrate the antennas on an OATS, taking the precaution of performing the calibration at more than one position on the OATS and averaging the results in order to reduce the contribution of the site errors. Some people are wedded to the principle of the NSA method, not realising that the fundamental limitation of the accuracy of validation of a site is the quality of the site on which the antennas were calibrated. Once this is understood, the SACM can be appreciated as a viable short-cut, with improved uncertainties.

One problem is that the SACM has not yet been incorporated into the standards. A way of overcoming this was to dress up the SACM to give it the appearance of the NSA method. A method bridging the NSA method and SACM, which improves on the uncertainty of the NSA method, yet does not quite achieve the lower uncertainty of the SACM, was termed the Dual-Antenna Factor method²³. For the Dual-AF method the combined antennas factors, from one SA measurement on a CALTS, are calculated. So now NSA can be derived by subtracting these AFs from the SA measured with the same antennas on the site-being-evaluated, and the NSA is compared to the theoretical values as is done for the standardised NSA method. This achieves lower uncertainties than the standardised NSA method because the Dual-AF is derived from only one SA measurement, and not the three required by the SSM. Ref. 23 takes the ultimate step of dispensing with NSA, further reducing the uncertainty, by describing the Site Reference Method, which is identical to the SACM described above.

5. Problems encountered in calibrations for customers.

It is not uncommon for antennas to develop faults, even the most recently manufactured ones. Current antenna factors no longer apply to the faulty antenna and this can lead to wrong field levels being reported in EMC Compliance Test Reports. A common fault is a loose internal electrical connection. Often the customer can discover this by measuring its reflection coefficient, which would show a glitch; but the calibration laboratory would notice this when the antenna was sent to them for its regular calibration.

5.1. Biconical antennas

Many models of biconical antennas have been found to have a serious balun imbalance. This causes unwanted current to flow on the outer conductor of the input cable, see section 9.2. The radiation from the cable interferes with the field from the EUT being measured. In the worst cases the signal is nearly cancelled, with errors up to 16 dB having been encountered. The effect is most pronounced when the input cable is laid out parallel to and in proximity to the antenna elements. This is a particularly serious fault for tuneable dipole sets, considered

by some operators as their most accurate reference antennas, especially when used with offtuned element lengths for the purpose of broadbanding. Antennas are being redesigned as manufacturers become aware of this fact. Fig. 7 shows the change in signal strength when a vertically polarised biconical antenna, exposed to an incident field, is inverted.

The mechanical tolerances in some balun designs of biconical and broadband hybrid antennas are quite critical, especially the higher power ones and it is quite common for the balun balance to degrade, the cause of which is suspected to be a mechanical knock. The degradation is most common around 30 MHz but has been noticed a lesser extent up to 150 MHz. A regular check of the balun balance is recommended.



Example of Poor Balun Balance

Figure 7. Balun imbalance of a high power biconical antenna

The most common design of biconical antenna is a pair of conical "cages" comprising 6 radial wires and connected to a balun at the centre, with a tip to tip length approximately $1.35 \text{ m} \pm 0.03 \text{ m}$ (depending on balun width) and a broadest diameter of approximately 0.52 m, see Fig. 8a. The standardised dimensions were originally published in MIL-STD- $461^{44,48}$. In the cage design of element, at approximately 287 MHz there is an internal resonance which can increase the antenna factor by up to 8 dB over a narrow bandwidth. This undesirable effect was cured by Schwarzbeck in around 1987 by attaching a bar to one of the six wires on each cage as illustrated in Fig. 8b, which pushes the frequency of the resonance above 300 MHz, the upper operating frequency of a biconical antenna, but it does mean the antenna factor is on a steep gradient as it approaches 300 MHz, the steepness depending on antenna model. Also the radiation pattern changes shape above about 260 MHz, but the deterioration is particularly noticeable at 300 MHz. The antenna gain can change by 0.3 dB depending on the orientation of the cross-bar relative to the direction in which the field is being measured²⁴. One model of antenna has 3 bars on alternating wires, which removes the

287 MHz resonance but unfortunately introduces another resonance, albeit smaller, at 217 MHz. Another model of antenna has hinges at the 3 bends on each of the 6 wires to make it collapsible and therefore portable. The hinges give intermittent contact and can cause unstable narrow band resonances.



Figure 8a. Schematic diagram of skeletal biconical antenna as used for EMC measurements



Figure 8b. Skeletal biconical antenna with bar to suppress resonance at 287 MHz



Figure 8c. Collapsible form of skeletal biconical antenna, with locking screws at apex of cones

The commercial collapsible design shown in Fig. 8c has been found to have a reproducible performance and is recommended²⁴. The electrical performance is very similar to the biconical antenna of Fig. 8b, but the antenna factor is lower and better behaved above 260 MHz. Also being an open structure it does not suffer from an internal resonance in free-space. Fig. 9 compares the antenna factors using the conventional and collapsible biconical elements. There is a small resonance at around 120 MHz when the antenna is horizontally

polarised at a height of 1 m above a ground plane, but this is not a problem as the height is usually greater than 1 m in an emission measurement. The use of collapsible elements is also encouraged for portability and storage. A conventional biconical antenna cage has a large volume for its weight and can be disproportionately more costly to transport than a collapsible biconical antenna.



Antenna Factors in Free Space

Figure 9. Comparison of computed antenna factors for conventional and collapsible 50Ω biconical antennas.

5.2. LPDA antennas

Another antenna design problem leads to large resonances in the response of LPDAs. These resonances are narrowband and can be typically 10 dB in magnitude. They do not usually show on new antennas, but appear unexpectedly during use. They are associated with a breakdown in RF contact between the dipole elements and the boom of the antenna to which they are bolted. The breakdown can occur if the element does not have a tight fit on the shaft, or if a metal oxide film builds up. Black powder deposits have been observed at the contact points of customers' antennas that are made of unpainted aluminium. This has been associated with input powers of tens of watts used in susceptibility testing, causing electrical arcing on a particular make of antenna. It is recommended that such antennas are limited for receiving use only. It is important regularly to check the performance of such antennas for the onset of resonances. The dipole elements can be removed and cleaned, restoring the antenna to its correct performance. Nevertheless, this does not always work and a full reconstruction of the antenna may be needed. There are models of LPDA for which attention has been paid to the mechanical construction, for example with welded elements, which do not suffer from this problem.

LPDA antennas can exhibit poor cross-polar performance. Section 4.1 cites an acceptable cross-polar rejection of 20 dB. A method of measuring cross-polar performance is described in Section 9.3.2. It is tempting for manufacturers to quote antenna factors of LPDAs to as high a frequency as possible, which looks reasonable where the antenna factor appears to continue to increase monotonically with frequency. But if this is taken to the extreme the cross-polar rejection can be zero, in other words the antenna accepts equal amounts of co-and cross-polar signal.

5.3. Monopole antennas

Some models of rod antenna have high gain amplifiers that can be susceptible to static electricity. As an example, the failure of the first amplification stage can lead to an increase in the antenna factor of 10 dB. Often this is not found out until the antenna is sent for its next routine calibration. Some models have switchable gain settings and the customer needs to specify whether they want it calibrated with the factory setting or the setting as sent to the calibration laboratory. Another problem for calibration laboratories is that active rod and loop antennas arrive either with their batteries flat or they need replacing. It is recommended to check that these are in good condition before sending the antenna for calibration.

5.4. Connector problems

The majority of problems that occur when calibrating EMC horn antennas above 1 GHz involve the antenna's connector. Some of these problems are not aided by the relatively poor quality connectors fitted to some antennas, though the problem is not helped by the inadequate care and maintenance that connectors receive at the hands of some EMC test engineers.

The most common problem is connectors for which the centre conductor protrudes beyond the reference plane. It is important to visually inspect and gauge connectors regularly, since a protruding centre conductor can damage any other connector to which it is attached. Apart from any damage that may be caused, measurements made with such a connector must be in doubt, unless the connector to which it is attached has been gauged and found to be recessed by more than the protrusion of the first connector. If not, the centre conductor of one of the two connectors will become displaced, possibly breaking its support bead, but certainly affecting the measured values. While a protruding centre conductor can lead to damage and suspect measurements, an overly recessed conductor will result in a large gap between the shoulder of the male pin and the female spring fingers. The effect of this gap on the mismatch at the junction rises rapidly with frequency. Fig. 10 shows the connector of a double ridged guide horn antenna being gauged.



Figure 10. Gauging the connector of an antenna

It is not uncommon to come across antennas with connectors whose centre conductor is not properly captivated, allowing the conductor to recede as the connection is made. This has been observed several times, particularly on older double-ridged guide horn antennas with Type-N female connectors. It is not possible to make repeatable measurements with such an antenna and the variation observed can be several dB. Thus it is important to check that the centre conductor is securely fixed. Although less common, occasionally one finds that, while the centre conductor is captivated, it can rotate. This is a problem if the launching pin inside the antenna is not quite symmetrical and can lead to very poor repeatability.

It is also very important to keep microwave connectors clean, as contaminants cause the connector to wear more quickly, can affect the mismatch at the junction and lead to poor repeatability. Hewlett-Packard published an excellent microwave connector care manual²⁵ that explains what an engineer needs to know about connector cleaning, gauging, specifications and use.

Good quality Type-N connectors will operate up to 18 or even 20 GHz without over-moding, but this is not true of all Type-N connectors. For antennas operating at higher frequencies, smaller line diameter connectors are required. SMA connectors are commonly used on antennas to operate to 26.5 GHz, but there are several potential problems with such connectors:

- 1) The quality of SMA connectors varies widely and only precision SMA connectors will operate reliably up to 26.5 GHz.
- 2) The repeatability of SMA connectors over many connection-disconnection cycles is poor. One reason is that the female centre conductor spring fingers are very thin and easily damaged during connection.
- 3) The dielectric support bead can start to protrude, resulting in a significant gap between the male pin shoulder and the female contact fingers.

4) No traceability exists for reflection coefficient measurements for SMA connectors.

A good alternative to SMA connectors is the GPC-3.5 connector. It is mechanically more robust, has better repeatability, and traceability for reflection coefficient measurements exists. Since very few antennas are furnished with GPC-3.5 connectors by the manufacturer, one can always attach a 3.5 (m) to 3.5 (f) adaptor onto an SMA (f) connector to act as a connector saver and provide a more repeatable calibration interface. It should be noted, however, that even though SMA and GPC-3.5 connectors can be mated and both have a characteristic impedance of 50 Ω , there is still an electrical discontinuity between them due to the dielectric-air interface. This discontinuity is frequency dependent, becoming more significant at higher frequencies, but is generally small in relation to other EMC measurement uncertainties.

Above 26.5 GHz, over-moding of SMA connectors is likely to occur. For antennas operating up to 40 GHz, the 2.92 mm connector is typically used (also known as the K connector). This is a precision connector for which traceability exist and is mateable with GPC-3.5 and SMA. However, it must be stressed that, while 2.92 mm connectors can be mated to GPC-3.5 and SMA, if used in a measurement system with those connectors, the mode-free operation of the system will be limited by the GPC-3.5 or SMA connectors. It will be possible to make measurements at higher frequencies, but not reliably, since any slight asymmetry caused by stress in the connectors or cables could lead to over-moding and very different results. It is a common mistake for test engineers to fall into the trap of judging the frequency range of the measurement system by the connectors on the measurement ports alone, without checking the cable types and other connectors in the system.

The 2.4 mm connector allows mode-free operation up to 60 GHz, but there are not many antennas currently available with this connector. At these frequencies, cable losses become very high. It is worth noting that the 2.4 mm connector is not mateable with 2.92 mm connectors.

While discussing connectors, it is also worth noting here that the centre conductor of some same sex adaptors are actually made in two parts, which are screwed together with a very fine screw thread. These can gradually become unscrewed due to the body of the adaptor being rotated during disconnection, rather than the coupling nut. There are two lessons to be learned here: a) never rotate the body of the adaptor when connecting or disconnecting; always rotate the coupling nut b) check the S-parameters of the adaptor regularly to ensure it is operating correctly.

Apart from connector problems, horn antennas are sometimes received for calibration with some of the screws holding the various component parts loose. Clearly this is likely to lead to non-repeatability performance of the antenna. It is important to check antennas before use to ensure there is no damage or loose parts. If anything is found to be loose or damaged, repair and re-calibration are essential otherwise all subsequent measurements will be compromised.

6. Self calibration using transfer standards

An EMC test house can designate an antenna as a transfer standard. For traceability to national standards to be claimed, the transfer standard must be calibrated by a laboratory with the appropriate accreditation for antenna calibration, for example UKAS accreditation in the UK. The test house can then calibrate its own antennas by substitution with the transfer standard, in other words by using the *standard antenna method* (SAM). In order for the secondary antennas to be traceable to national standards, two fundamental points need to be observed. The transfer standard should only be used for calibrations, or handled so that its UKAS calibration is not invalidated. The second point is that the method of calibration should be acceptable to an accreditation body and that the additional uncertainties imparted to the secondary antennas shall be agreed by the assessor.

The test house needs to be aware of the problems with performing antenna calibrations. If the *three-antenna method (3AM)* or SSM is used, the measurement site needs to be of a high quality - traceability can be claimed if the site attenuation is measured by a UKAS accredited laboratory. ANSI C63.5 does not recommend calibrations using vertical polarisation because of problems with accuracy²². CISPR 16-1-5 deals comprehensively with the quality of the site required. Detailed methods of calibration are under consideration, to be added to that CISPR standard.

Special care is needed in the calibration of problem antennas described in Section 5. When using the substitution method, accurate calibrations can only be obtained when the transfer antenna is of the same construction as the secondary antenna. For example there would be additional uncertainty terms if a tuneable dipole were used to calibrate a biconical antenna or if a transfer LPDA were not an identical model to the secondary LPDA.

NPL measures free-space antenna factor of biconical antennas above a ground plane using vertical polarisation to reduce mutual coupling to its image, as discussed in Section 9.2.2. Not many test houses can do accurate calibrations using vertical polarisation. If a customer requires an antenna calibrated for transfer purposes employing horizontal polarisation, this should be stated and a horizontally polarised calibration will be provided. If the customer has a 6 m mast, NPL can provide a calibrated in free-space conditions. It is difficult to measure boresight AF of an LPDA at a fixed height above a ground plane, but this is not essential because the antenna factor changes by no more than an estimated \pm 0.4 dB as the antenna is moved upwards in height above 1 m. This applies to a horizontally polarised LPDA with lowest frequency of 200 MHz placed above a ground plane; the changes are less when the antenna is vertically polarised.

In view of all the complications owing to the introduction of the reflecting ground plate and the recent wide availability of fully anechoic chambers it seems be worthwhile to consider a standardisation of the use of chambers with a corresponding modification of field strength limits to free space values.

7. Calibration interval

The calibration interval depends on the quality policy of the organisation and also on the reliability of the antenna. The antenna factor of a well-designed and well-built antenna should not change provided the antenna is handled with care. If repeat calibrations show that the antenna factor does not deviate by more than the uncertainty of measurement, the calibration interval can be extended. The recommended interval for the recalibration of antennas by a UKAS accredited laboratory is 2 years. In between calibrations regular intermediate checks should be performed and if these show the antenna to be in good condition and it is known that the antenna has been handled carefully and not had a high power input, for example, the calibration interval could be reviewed and a longer interval considered.

A mechanical check should be performed each time an antenna is used. This will ensure that the RF connector is in good condition and that the antenna elements are firmly affixed and are not bent out of shape. The user should regularly check the input match and do a functional check using a stable noise source, or measure the insertion loss between two antennas for a defined setup. These measurements should be made immediately after the antenna has been calibrated to provide reference data for subsequent internal checks. The user may prefer to do an internal antenna calibration as a repeatability check to ensure that the antenna performance has not changed since the user last measured it with an identical setup. Two problems to look out for are delicate baluns which can suffer if the antenna experiences a mechanical knock, and LPDA elements that are not welded to the main body of the antenna which can corrode and lose RF conductivity, especially if immunity testing input powers are used.

8. Calibration test sites

In order to be able to achieve low uncertainties of measurement of antenna factor, a test site is needed that is of higher quality than that required for EMC compliance testing. A description of an open area test site, including the recommended size of ground plane is given¹⁰ in clause 5.6 and Annex D of CISPR 16-1-4. Clause 5.6 states that the measurement of the site shall be within ± 4 dB of a specified value. Annex F shows that the intention is that the antenna factor uncertainty shall be better than ± 1 dB. Clearly the site has to be much better than ± 1 dB in order to calibrate antennas to less than this value. CISPR 16-1-5 gives a specification¹¹ for a Calibration Test Site with the criterion that the site has to be within ± 1 dB of a specified value. Unless the site itself is closer to ± 0.5 dB, an antenna calibrated on it, by the three-antenna method, is likely to have an uncertainty greater than ± 1 dB. On the other hand the standard antenna method is less demanding of the quality of the site, because if the standard antenna has similar dimensions to the *antenna under test* (AUT), the unwanted site effects can mostly cancel.

A real problem with calibrating antennas on an outdoor CALTS is ambient RF interference, particularly from radio and television transmissions. In the south east of England these give an ambient level of up to 80 dB μ V/m, in the region of 90 – 230 MHz and at various frequencies within 390 – 600 MHz (TV). The levels are up to 5 dB higher at around 940 MHz, presumably caused by mobile phones. Ideally the level of signal received by the receiving antenna should be at least 20 dB above the ambient level. However it is likely that the transmit level required to achieve this at all frequencies will be higher than that permitted by the Broadcasting Regulatory Authority. It may be possible to get an agreement with the

authority to be allowed to increase the power into the transmitting antenna at frequencies where the ambient level does not permit a valid measurement. This restriction on power led NPL to develop a method for calibrating LPDAs in close proximity, as described in Section 9.3. This involved knowing the phase centre positions accurately. A spin-off was the reduction in the amount of absorber needed on the ground plane between the antennas, to create free-space conditions.



Figure 11. Plot of typical levels of ambient interference in the south-west of London.

The term antenna factor usually implies the free-space value, bearing in mind that AF can change by several decibels when sited above a large ground plane. Consequently an ideal site for the measurement of AF is a free-space site. In practice a good approximation to free-space can be achieved by an anechoic chamber, particularly if the antennas are directional and therefore they "see" less of the chamber walls. The corollary to this is the "omni-directional" dipole antenna, especially at frequencies below 150 MHz where it is not practicable to get far enough away from chamber walls, or the ground if outdoors; the solution for which is to quantify the ground reflection by means of a large, flat, well conducting ground plane.

Another method of obtaining free-space conditions is the use of TEM cells. Plane wave conditions exist between the plates of the cell. Providing the antenna does not couple unduly to the walls of the cell, the TEM cell is a practical solution for calibrating monopole and loop antennas to uncertainties that are acceptable to the majority of users. A symmetrical TEM cell that has a plate separation of 0.915 m has been found to be satisfactory for calibrating loop antennas up to 0.6 m in diameter to uncertainties of better than ± 1 dB, from 20 Hz to 90 MHz. Popular proprietary rod antennas exceed 1m in height, so a larger TEM cell is needed, bearing in mind that the upper usable frequency decreases with increasing cell size, because of the onset of resonances. It has been found that a *GTEM* cell with a maximum plate separation of 1.75 m is suitable for calibrating rod antennas up to 1.25 m in height to

uncertainties of better than ± 1.5 dB, from 100 Hz to 30 MHz. With extra precautions to remove earth loops, it is possible to extend the frequency range downwards to 20 Hz.

8.1. Open area test site (OATS)

For the purposes of antenna calibration, an OATS has a ground plane that is large enough to give an area underneath each antenna such that a sufficient image is formed, and includes the region between the antennas so that the reflection is well defined, i.e. in contrast to a car park surface or soil. The space above the ground plane has to be free of obstacles that could cause reflections, however it is accepted that the antenna or EUT has to be supported, but the table or mast is made of dielectric material that has a minimal effect on the signals.

An example of an OATS is the open-field EMC antenna calibration facility at NPL, which has gained wide acceptance as a national standard open-field site. The ground plane is welded 8 mm thick mild sheet steel of dimensions 60 m by 30 m. 95% of the area is flat to within \pm 6 mm. The site has no weatherproof cover, because this could compromise accuracy. The centre axis is 45 m away from the control building, constructed of brick and timber (albeit containing electrical wiring) and 50 m away from nearest trees, so that unwanted reflections are negligible. As a measure of the quality of the site, the insertion loss between two horizontally polarised resonant dipole antennas differs from theoretical prediction by less than \pm 0.1 dB in the frequency range 30 MHz to 300 MHz. This level of agreement confirms the quality of both the site and the NPL calculable standard dipole antenna²¹. It was estimated that the ground plane contributes less than 0.05 dB to this difference. At 1 GHz the difference increases to \pm 0.3 dB, which is attributed to the uncertainty of the dipole location (especially height) rather than to site imperfections. A receiver with a linearity of \pm 0.01 dB over a dynamic range of 100 dB was used for these measurements.

A comparison of measurement and theory is required by CISPR to validate a site, though the CISPR criterion for calibration site errors¹¹ in NSA is ± 1 dB. NPL's calculable standard dipole antenna is accurate to ± 0.15 dB over the frequency range 20 MHz to 1 GHz. However the measurement uncertainties offered for customer calibrations include other components, such as mismatch. The uncertainties get larger as the frequency increases to 1 GHz, because of the distance tolerances in setting up the antennas, and reflections from supporting structures.

The facility is equipped with a turntable flush with the ground plane, driven by an azimuth positioner beneath it. This, and some special low reflectivity towers, equips the site with the capability for making radiation pattern measurements and for locating the phase centres of antennas. The turntable can be used also for EMC radiated emission testing.

One of the surprises in the course of trying to make the most precise measurements was the effect of temperature on the cables connected to the antennas. Originally black sheathed double-shielded RG214 cable was used, typically a 10 m length from each antenna to the N-type bulkhead connector in the ground plane, and thence underground via low loss heliax cables to the control room. On a day when the clouds were intermittently exposing the sun the signal was noticed to drift by around 0.3 dB. It took some time to work out the cause. The result was replacement of the cables by a type with a white sheath, and attention to minimising the time between the through connection (cables joined) and the antenna measurement.

Another discovery was that a thin layer of ice on the ground plane had very little effect on the measurement, certainly up to 300 MHz (since RAM was used on the ground for LPDAs), in the context of a ± 1 dB uncertainty in AF. The NPL certificate for passive EMC antennas states that the temperature range for the antenna setup can be -5° C to $+35^{\circ}$ C (ice through to direct exposure to the summer sun) and experience indicates that the effect of temperature on the antenna has less than ± 0.1 dB influence. Though not quantified, it is assumed that wetness on the antenna mast and guy ropes can make them partially conducting and therefore potentially reflective, so as a precaution the procedure is to dry off any moisture. A fine mist does not seem to affect the calibration, but as soon as water drops begin to form on the antenna elements, it is time to take them indoors, under the assumption that the slightly conductive water increases the electrical radius of the elements.

8.2. Case study for design of an OATS

NPL built its first OATS in 1988 and studied the conditions¹² for achieving good measurement accuracy. The ground plane was made of wire mesh that was laid on marinegrade plywood placed on levelled sand. Guidance on the area of ground plane needed for measurements in the frequency range 30 MHz to 1 GHz was obtained from the 1987 edition of CISPR Publication 16. This gave the minimum size of the obstruction free area as an ellipse with major axis 2R and minor axis $\sqrt{3R}$, whose foci are the positions of the source and receive antennas and R is the distance between them. However, if there were a metal hanger on this boundary that reflected all the energy between a pair of horizontally polarised dipoles, the uncertainty would be ± 0.5 dB. For a pair of vertically polarised dipoles, which are isotropic in the azimuthal plane (ie H-plane), the possible uncertainty is ± 3.5 or -6 dB; uncertainties due to antenna factors and an imperfect ground plane have to be added to this. An absolute criterion is that the reflected power should be below a specified level and a suggestion is -35 dB, which represents an uncertainty of ± 0.15 dB. If there are trees on the edge of the ellipse it is prudent to calibrate the site in both wet and dry conditions. Time domain techniques could be used to gate out reflections from the site boundary.

The ground should be flat so that the phase of the ground reflection is known. It may deviate from the expected value by a given amount due to undulations or small objects. The Rayleigh criterion (which limits the phase deviation to 90°) is used by CISPR and results in a maximum RMS surface roughness of 0.15λ on an R = 3 m site, i.e. 45 mm at 1 GHz, see Annex D2.2 of Ref. 10. On a site used for measurements of national standards a phase deviation of $\pi/8$ is more appropriate²⁶ giving an RMS roughness of 11 mm at 1 GHz.

The advice in the literature around 1988 was to avoid using earth (e.g. grass field) as a ground plane because a set of measurements would have to be made to determine the complex reflection coefficient each time it was used; instead a conducting ground plane was recommended, either made of sheet metal or wire mesh with a 7 mm grid. All metals practical for this purpose have a sufficiently high conductivity (>10⁷ S/m) to make no significant difference to the NSA value. This should not be covered with a protective layer of significant thickness as it would alter the phase of the reflection, and measurements would not be repeatable on sites with different ground cover. There should be no gap between the earth and the mesh, which could result in resonances. The edges of overlapping panels of wire mesh should be welded at intervals of a fraction of the wavelength at the highest frequency of intended use.
CISPR recommended a maximum mesh size of a tenth of a wavelength at the highest operating frequency. However, this could let through up to 35% of the power²⁷ depending on the measurement geometry, resulting in a measurement error of approximately 0.7 dB; which would be considerably reduced if the mesh were laid on moist earth. The transmission through 7 mm grid mesh is 0.02 dB at 1 GHz, at which frequency the mesh size is $\lambda/50$. ANSI²⁸ recommends that the size of the ground plane for an open-area test site should cover an area incorporating the first Fresnel zone. The major and minor axes of the first Fresnel ellipse²⁶ for a 10 m site at 30 MHz are 16.3 m and 13 m. For an antenna calibration site, the ground plane should extend a distance of at least λ beyond the antenna so that it has a well-defined interaction with its image and this can be theoretically taken into account. Thus, for a 30 m site the ground plane should be a minimum size of 25 m x 50 m. The edge should be well earthed to the surrounding soil and if the soil has good conductivity (i.e. when wet) it forms a good extension to the wire mesh. Measurements using vertically polarised antennas are sensitive to edge diffraction effects of a ground plane and this could be used to confirm whether the size is adequate: a serrated edge can reduce edge diffraction.

Edge diffraction on the NPL 60 m x 30 m ground plane shows up on a plot of SIL vs. frequency for a pair of vertically polarised biconical antennas spaced 20 m apart on the long axis of the ground plane. The ripple on the plot is very noticeable (2 dB peak-peak) in the null region of the SIL plot because the coupling between the antennas has decreased with respect to the edge reflected signal. This ripple has a frequency of approximately 28 MHz and is distinct from the ripple (50 - 200 MHz) caused by reflections from a vertical cable which is much closer to the antenna. Measurements between vertically polarised s along the short axis of the open field site suffer less from edge diffraction than do measurements along the long axis (because of the longer distances to the edges). In practice most measurements are made near the maximum coupling between antennas, when the direct and specularly reflected signals are in phase, and the edge rays caused a ripple on the plot of about ± 0.15 dB. This was reduced by a factor of about 2 when the earth around the perimeter of the ground plane was soaked by heavy rain. The NPL sheet steel ground plane extends by 1 m of wire mesh into the surrounding earth, but this turned out not to be effective at dissipating the edge currents, because the earth is rarely wet enough. Edge diffraction is undesirable, but it would be a major undertaking to modify the ground plane to eliminate edge reflections - one possibility is a serrated edge, but the serrations would have to be very large to be effective at 30 MHz.

8.3. Anechoic chambers

Before the advent of RF (or radar) absorbing material (RAM), the creation of a free-space environment usually meant a remote area outdoors free of reflecting obstacles (such as buildings and trees) and towers to put the antennas well above the ground. If the antenna had insufficient gain to exclude a ground reflected signal, other means such as diffraction fences could be deployed in the path between the source and receiving antennas. The introduction⁴⁹ of RAM in the 1950s made it possible to create a free-space environment indoors. At first this was at frequencies above about 200 MHz, but in the 1980s ferrite tiles were introduced that gave effective absorption in the range 20 MHz to 200 MHz. One attraction of being indoors is control of temperature and freedom from the vagaries of the weather. However high gain antennas may need a range length of the order of 1 km in order to measure the far-field gain. The spur to developing near-field scanning techniques was the advent of computers, and there

was a surge in employment of near-field scanning in the 1970s, enabling high gain antennas to be measured indoors.

8.3.1. Ferrite lined chamber 30 MHz to 1 GHz.

In 1999 NPL commissioned a state-of-the-art ferrite lined chamber in a shielded room of internal dimensions 9 m x 6 m x 6 m. All internal surfaces are lined with 6 mm thick ferrite tiles that were carefully abutted. In the greater part of the central area of the walls (including floor and ceiling) pyramidal absorber was affixed on top of the tiles. This absorber is of the low-density carbon loaded type that makes a hybrid with the tiles in order to optimise the reflectivity across a wide frequency range. Beyond the central area, extending to the interface between the walls, shorter truncated pyramidal absorber was placed. The chamber was designed to meet the CISPR NSA criterion from 30 MHz to 18 GHz, and agreement within ± 2.3 dB was achieved over the frequency range 30 MHz to 1 GHz (see clause 5.8 of CISPR 16-1-4). An agreement of better than ± 2 dB was achieved from 1 GHz to 18 GHz, but this was using a pair of horn antennas, whereas the current CISPR draft method uses an omnidirectional antenna in place of one of the horns. Publication of this CISPR method is expected in 2006.

The chamber is primarily used for the calibration of wire antennas and incorporates a pair of manually adjustable antenna masts. Height scanning is not employed. To enable mobility into the room and between the masts, a corridor was created by replacing some RAM by 100 mm high ferrite pyramids. These were specified to have low reflectivity up to 18 GHz and can be walked upon. Fig. 12 shows the chamber. Some research was carried out to evaluate the effect on the measured field strength of proximity of antennas to the RAM. The conclusion reached was that there is very little effect as long as the antennas are more than 0.5 m from the tips of the pyramidal absorber.



Figure 12. Ferrite lined chamber of dimensions 9 m x 6 m x 6 m.

8.3.2. Pyramidal absorber lined chamber 1 GHz to 40 GHz.

NPL has a fully lined anechoic chamber used for calibrating horns, LPDA and spiral antennas at frequencies above 1 GHz.

Performing antenna measurements outdoors presents several difficulties, such as the unpredictable weather, reflections from nearby objects and the ground, interference from extraneous sources and interference to others. The use of a shielded anechoic chamber eliminates these problems, but is only practical at frequencies where effective RAM of a reasonable size can be employed. Measurements carried out inside an anechoic chamber will give the free-space antenna factor directly (provided there is sufficient separation between the antennas), as there will be no ground reflections. There is no need for height scanning or measurements both horizontally and vertically polarised.

An anechoic chamber 8 m long, 5 m high and 5 m wide lined with 60 cm RAM provides an ideal environment for calibrating most EMC antennas above 1 GHz at separations up to 3 m.

For such a chamber, the centre line is 2.5 m above the floor, which is a convenient height for most personnel if walk-on RAM approximately 73 cm high is used. Temperature control is desirable, but not essential, provided extremes of temperature that would affect the integrity of the measuring equipment can be avoided and the effect of fluctuations is included in the uncertainty budget.

The antennas have to be supported and directed so that they point towards one another along the mid-line of the chamber and it must be possible to readily adjust the separation between them. Often, low reflectivity masts or tripods are used to support the antennas, but these can be time consuming to reposition if measurements have to be made at more than one separation. An alternative is to install a rail system in the floor of the chamber and to use two low reflectivity carriages. This arrangement allows the separation between the antennas to be adjusted, while keeping the alignment constant, which is more important for directive antennas operating above 1 GHz than for dipoles and biconical antennas. A picture of the setup in the NPL microwave EMC facility is shown in Fig. 13. This chamber is used to calibrate antennas from 1 GHz to 40 GHz.

To avoid reflections from the rails in the floor of the chamber, a rail covering system may be installed. By placing the floor RAM on wooden trolleys driven by linear actuators, it is possible to automate the covering and uncovering of the rails and further improve measurement efficiency. This approach also reduces wear of the floor RAM, since it does not need to be handled to re-position it. Such a system has been installed in the author's facility using skylight-type actuators with a control panel in an adjacent room. A schematic diagram of this system is given in Fig. 14.



Figure 13. Anechoic chamber showing wooden support carriages



Figure 14. Rail covering system, trolleys withdrawn

8.4. TEM cells

A Crawford TEM cell²⁹ consists of a coaxial line in the form of a rectangular outer conductor with a centrally placed inner conductor, known as a septum, as shown in Fig. 15. The relative dimensions of the inner and outer conductor are chosen to provide a characteristic impedance close to 50 Ω to match the coaxial transmission lines normally associated with RF metrology. This two-conductor transmission line will support the transverse electric and magnetic (TEM) mode of propagation just like a coaxial cable.

A TEM wave is characterised by orthogonal electric (E) and magnetic (H) fields, which are perpendicular to the direction of propagation along the length of the transmission line. The rectangular structure of the TEM cell produces near uniform fields over a useful cross sectional area between the inner conductor (septum) and the outer conductor, the components of which simulate a near plane-wave field in free space, having an inherent wave impedance of approximately 377 Ω . The losses in the cell are negligible, due to the large conducting area; therefore only the dimensions of the cross section and standing waves set up by imperfect load match determine any longitudinal variations in field strength. Higher-order modes (waveguide modes) of propagation, a degeneration of the TEM mode and which are superimposed on it, can also cause spatial variations in the field strength if the cell is used above certain frequencies known as the cut-off frequencies.

The basic rectangular transmission cell has an upper frequency limit determined by the onset of higher-order modes of propagation. The frequency and dimensions of the outer conductor, as for a waveguide, determine the number of higher order modes. The first significant higher order mode with the lowest cut-off frequency is generally the first order TE mode (TE₁₀) which can exist when the width of the cell is more than half the wavelength of the operating frequency.



Figure 15. 6 foot (1.83m) high TEM cell for calibrating loop antennas

The GTEM (Gigahertz-**TEM**) cell is a version of TEM cell that enables higher frequency operation. It was introduced in 1987 and is used primarily as an alternative facility for EMC testing. Both the input and the output of a Crawford TEM cell taper to coaxial line. Above a certain frequency the higher order modes can propagate and the tapered sections act like short circuits, causing cavity resonances. The higher the desired frequency of operation, the smaller the cell has to be. For example a cell nominally designed to operate to 500 MHz has a plate separation of only 0.14 m. In order to avoid a cavity the GTEM cell has only a tapered input, the other end being a broadband termination. This termination consists of a 50 Ω distributed resistor board for low frequencies and pyramidal absorbers for high frequencies. The absorbers are arranged on a section of a sphere so that the tips of the absorber point towards the apex of the GTEM cell. The septum is placed at the upper third of the cell, allowing for a larger test volume beneath the inner conductor. NPL has written a Good Practice Guide on GTEM cells⁸. A large GTEM cell is pictured in Fig. 16, which is primarily used for the calibration of rod antennas, as described in Section 9.10.



Figure 16. GTEM cell with maximum plate separation of 1.75 m, used for calibrating monopole antennas.

9. Outline of methods for antenna calibration

The main technique for measuring the gain of antennas is the three-antenna method³⁰. Albert Smith published the formulae for determining the antenna factor by 3AM above a ground plane³¹. The other important technique is the standard antenna method, in which the antenna factor of a standard antenna already has to be known, and this is substituted for the antenna-under-test in order to determine the antenna factor of the latter. The main feature of 3AM is that antenna factors do not have to be known beforehand, but the method relies on measurements of the coupling between pairs of antennas separated by a defined distance on a site free of unwanted reflections. The quality of the site will directly impact on the achievable uncertainty of the antenna factor. A standard antenna is usually calibrated by 3AM, or the standard antenna can be calculable. The site has less influence on the uncertainty of the AUT antenna factors determined by the SAM, as elaborated in Section 9.2.1.

CISPR sub-committee A plans to publish a detailed standard on the calibration of EMC antennas by 2006, probably in CISPR 16-1-5¹¹, which will be a valuable resource. The methods described in this Guide have been refined at NPL since 1989 and the low uncertainties achieved for antenna factor have been made possible by the special ground plane and by substantial funding of development of methods. These methods of antenna calibration are fully compatible with the methods recommended by the standards-making

bodies, recognising that the purpose of such calibrations is to minimise the measurement uncertainties of emission tests done in the manner laid down by these standards.

A useful reference for methods of calibrating EMC antennas is published by the American National Standards Institute as ANSI $C63.5^{22}$. This gives detailed guidance on the quality of the test site and practical advice on how to perform the calibration. It is not the intention to replicate this level of detail in this Good Practice Guide, but rather to recommend resources, to put these in context, and to inform readers about alternative methods. ANSI C63.5 focuses on providing free-space antenna factors, or near free-space antenna factors because of the difficulty of realising true free-space conditions; in this Guide the term quasi free-space is also used.

ANSI C63.5 usefully describes in detail the construction of a simple design of tuned dipole antenna. The calculated free-space antenna factor has an uncertainty of less than ± 1 dB, however no guidance is given as to the deviation of antenna factor with height above a ground plane, which, in the extreme case of a horizontally polarised dipole tuned to 30 MHz, could be as much as 4 dB. Currently CISPR 16-1 refers to the tuned dipole as a reference antenna. However, mainly because of the deviations caused by mutual coupling, there is a move away from using a reference antenna as the criterion for establishing whether or not an emission limit has been met. It has been proposed that measured E-field using a calibrated broadband antenna be used as the criterion to decide whether an emission limit is met. The antenna is of the type in normal use by EMC test laboratories and the limit is written in terms of field strength, dBµV/m.

There is a range of possible calibrations for any one antenna. In order for the customer to get the lowest possible uncertainty of measurement, the calibration can be tailored to the way the customer uses the antenna. For example, if the antenna is used horizontally polarised at a fixed height above a ground plane, such as in an NSA measurement, antenna factors can be provided for that height, eliminating the component of uncertainty that takes account of variation of antenna factor with height. As mentioned in Section 9, AFfs is the default value, and that is a good choice where one wants a unique value of AF to cover the changes in antenna factor when the antenna is height scanned, because the variations are roughly cyclical with change in frequency about the free-space value. Because of the difficulty of obtaining free-space conditions for omni-directional antennas below about 150 MHz, quasi-free-space methods are used that may involve larger uncertainties than those achievable at higher frequencies.

Another example of the way the customer uses an antenna, is for EMC testing with the antenna at a horizontal separation of 3 m from the EUT and scanned in height from 1 m to 4 m. The antenna is obviously not always at 3 m distance from the EUT and when the antenna is at 4 m height there is a significant deviation from the boresight direction of the antennas, where it is normally calibrated, to the point of specular reflection in the ground plane. The uncertainties caused by deviation of the distance from 3 m, and of the radiation pattern from boresight gain would be reduced if the antenna were calibrated in this "geometry specific" way³² with separate calibrations for horizontal and vertical polarisation – see Section 9.6.4. If a broadband hybrid antenna is used, there is even more reason to use the geometry specific method because the large potential phase centre error will be reduced – with the proviso that the other two antennas (used in 3AM) must be of a type that have little change in phase centre with frequency.

Another important example is the use of antennas for site evaluation. If the site fails by a fraction of a decibel this can present a big problem for the supplier of the site, therefore it is important for the antenna related uncertainties to be as low as possible. This is best achieved by not using individually calibrated antennas, because antenna uncertainties are normally the biggest contributor to the overall uncertainty. Instead the insertion loss measured between a pair of antennas on a reference site is used, which more than halves the antenna related uncertainties. This point is further addressed in Section 9.7.

9.1. Checks of condition of antenna

The following antenna properties should be checked prior to the antenna calibration

(1) A visual inspection to ensure that the antenna is not mechanically damaged and that the input connector is clean and dry. Carefully straighten any elements that have been bent in a way that departs from the manufacturer's intended design.

(2) Check that the pin depth of the connector is within the required tolerance, using a coaxial connector pin-depth gauge. If the pin is protruding too much, tightening the cable connector can damage the antenna, but if the pin is recessed too much it can affect the match and, in an extreme case, the transmission. When mating connectors, turn only the threaded collar and not the body (for example an attenuator) of the connector, so as not to damage delicate parts of the connectors, such as spring fingers.

(3) Measure the return loss at the output port of the antenna. It should be similar to the return loss supplied by the manufacturer, or documented in the previous calibration. Typically biconical antennas and tuned dipoles should have a return loss of better than 20 dB at their resonant frequencies and an LPDA antenna should have a return loss of better than 10 dB.

(4) Measure the balun imbalance. This applies to dipole and biconical antennas, including the biconical part of hybrid antennas, usually in the frequency range below 200 MHz. A balance test, first devised at NPL in 1990, is described in Clause 4.4.2 of CISPR $16-1-4^{10}$.

(5) Perform a swept site insertion loss measurement with a maximum frequency resolution of 1 MHz between the antenna to be calibrated and a second antenna to see whether narrow band resonances occur, as caused by the antenna to be calibrated. If a resonance cannot be removed, the antenna factor shall be measured at small enough frequency intervals to detect that resonance.

Note: Log-periodic dipole array (LPDA antennas) and broadband hybrid antennas can exhibit these resonances. If a resonance was not there initially, but has appeared since the antenna was procured it can usually be removed by cleaning the join between the dipole elements and the boom. Elements that are welded will not give rise to resonances due to breakdown in RF contact.

Measurement of the radiation pattern of EMC antennas is rarely required. The issues are considered in Section 9.8. However it is taken for granted that the antennas are low gain and have very broad beamwidths. Some manufacturers provide selected radiation pattern information. One should be cautious about using the antenna outside its specified operating frequency range. For example the E-plane pattern of a biconical antenna, with cage-like elements (see Fig 8a), can start malforming above 260 MHz and multi-lobing above 300

MHz because the antenna is longer than a wavelength. The same is true of dipole antennas if used above their tuned frequency. If LPDA antennas are used below their specified lower frequency the back lobe can become large because there are no dipole elements to create an array below the lower frequency.

9.2. Calibration of dipole and biconical antennas

Dipole and biconical antennas can be calibrated over a ground plane using the version of the three-antenna method formulated by Smith³⁰, involving height scanning, which is called the *Standard Site Method* (SSM). The formula assumes the *H-plane* radiation pattern to be uniform, and this is a reliable assumption for these types of antennas. In the case of biconical antennas whose operating frequency range is 30 MHz to 300 MHz, if the antennas are horizontally polarised and separated by a distance of 10 m, with the fixed antenna at a height of 2 m and the other antenna scanned in height from 1 m to 4 m, the antenna factor should be within ± 0.5 dB of the AFfs. This small margin assumes the use of a very high quality test site, qualified staff, a stable signal generator and a receiver with good linearity. Smith's method is incorporated in ANSI C63.5, the latest version²² of which enables antenna factors to be more precisely converted into AFfs. The correction factors take account of mutual coupling between the antennas and between the antennas and their images in the ground plane, enabling the measurement uncertainties to be reduced.

What applies for the biconical antenna also applies to the "short dipole" antenna, which is a fat dipole that covers the frequency range 30 MHz to 80 MHz. It also applies to any other dipole that is used in a broadband fashion, apart from the ANSI correction factors that have been formulated only for the biconical geometry. However correction factors are not needed for the short dipole because, as the frequency is reduced below the resonant frequency, the input impedance increases, thereby desensitising the antenna to its environment. The short dipole is produced by some manufacturers and is cited in CISPR 16-1-4.

A more accurate method of measuring the antenna factor of dipole and biconical antennas is to use the calculable dipole antenna developed by NPL²¹. The SAM is used to compare the antenna-under-test to the calculable dipole. The principles of the SAM are given in ANSI C63.5 under the heading of Reference Antenna Method. The advantage of using the calculable dipole is that its antenna factor can be calculated for free-space and for horizontal and vertical polarisation at any height above a good quality ground plane. A proprietary version³³ of the dipole comes with software whose kernel is the Numerical Electromagnetic Code which enables the antenna to be used in a broadband fashion; the frequency range 30 – 1000 MHz can be covered by 4 dipoles without the antenna factors becoming too large at the end of the bands.

A major source of measurement uncertainty for biconical antennas is unwanted currents set up on the outer conductor (or braid) of the coaxial feed line. This can also apply to tuned dipoles, particularly off resonance where the input impedance may be high. These currents are caused by poorly balanced baluns. Using two such antennas to measure NSA can make it impossible to get the site to meet the CISPR NSA criterion. The problem is that radiation from the braid currents destructively interferes with radiation from the transmitting antenna. By reciprocity the same is true for a receiving antenna. There has arisen the mistaken belief that the antenna has a preferred polarity when vertically polarised, i.e. with one particular biconical element pointing skywards, and that if the antenna is calibrated and then used in this orientation the problem is solved. In fact the problem is caused by the interaction of the antenna with its feed cable and is dependent on the exact cable layout, the main factor being the distance between cable and antenna. It can be reduced by extending the cables at least 10 m behind the antennas before they are allowed to be parallel to the antenna element axis. When the antenna is horizontally polarised the distance can be minimal providing the cable is laid orthogonally to the antenna. Ferrites are commonly used to reduce braid currents, though it is mistakenly thought by some to completely cure the problem. NPL uses clip-on ferrite clamps on the half-inch diameter cable in all cases where balun imbalance is suspected, and it is worth researching the best ferrite for the required frequency range and for a particular size of cable. Antennas are commercially available which have very satisfactory balance and operators are encouraged to use these.

9.2.1. Calibration of tuneable dipole antennas

Tuned dipoles can be calibrated by the SSM, however by its nature the tuned dipole is most sensitive to coupling with its image in the ground plane. There can be errors of typically ± 2 dB in the measurement of antenna factor below 200 MHz. It is also the case that the SSM is not exact, because there are three simultaneous equations to solve for four unknowns: although there are only three antennas, one of the antennas is at different heights for each of the two measurement pairs it is involved in; because of mutual coupling with its image its antenna factor at each of the two heights is not unique. Adapting the method by having fixed heights for all three antennas can solve this. The AUT is placed at the desired height and the height of the other two antennas is chosen to avoid a signal null in the insertion loss measurement for each pair of antennas. The three pairings of three antennas A, B and C, at one of two heights h1 or h2, are shown in Fig. 17.



Figure 17. Antenna heights in fixed-height three-antenna method.

A simpler and probably more accurate way of measuring antenna factor is by substitution with a calculable dipole. The calculable dipole can be the same length as the tuneable dipole so that it is sampling exactly the same field. In this way any small non-uniformity of the field will cancel when the two insertion loss results are subtracted, thereby eliminating this site effect from the measurement uncertainty. At 30 MHz a tuned dipole is approximately 4.8 m long and because of the small rod diameter the tips will droop by typically 16 cm for most commercial antennas, when horizontally polarised. When the 30 MHz dipole is at a height of

1 m above a ground plane this droop causes a reduction in the antenna factor of approximately 0.25 dB. This is the worst case; the value is less for greater heights and at higher frequencies (where the dipoles are shorter and do not droop).

The antenna factor of the calculable dipole can be calculated for any height, so for example in this way a tuned dipole can be calibrated for a height of 2 m above a ground plane. The *opposing*, or source, antenna has to be set at a height so that a null condition is avoided. AFfs can be measured at heights above the ground plane where effects of coupling with the image are negligible. Manageable heights are practicable above 200 MHz. However at lower frequencies the change in antenna factor with heights up to 4 m can be large - 6 dB at 30 MHz. A trick is to set the height to an intermediate value where the antenna factor coincides with the AFfs, as illustrated in Fig. 4 which shows calculated heights where the antenna factors have been produced³⁴ in which antenna factors measured at a height of 2 m can be converted to AFs at other heights. If AFfs is known, Ref. 34 also includes tables to give the AF at other heights for horizontal and vertical polarisations.

The required tolerance on setting the length of a tuned dipole is a percentage of its length. For example it is not necessary to have a tolerance of 2 mm for a 30 MHz dipole. A tolerance of 1% would cause a change in AF of less than 0.1 dB. This means 50 mm at 30 MHz, but only 1.5 mm at 1 GHz. If the tolerance has to be relaxed, this can be accounted for by a higher uncertainty. Because AF is inversely proportional to the effective length of the antenna (see Appendix 1, clause A1.1), for small deviations from the resonant length, the change in AF is given by 20 log (1 + P) where P is the percentage change in length.

9.2.2. Calibration of biconical antennas

As for the tuned dipole antenna, methods for calibrating biconical antennas are the SSM, SAM, and the fixed height 3AM. These last two methods enable precise antenna factors to be measured for use in the NSA measurement of the quality of a test site, in which one of a pair of antennas is placed at a fixed height above a ground plane.

NPL has developed a method for measuring the free-space antenna factor without having to mount the antenna at extraordinary heights, such as 8 m, which is two wavelengths at the typical resonant frequency of a biconical antenna. The method takes advantage of the fact that mutual coupling to its ground plane image is much less for a vertically polarised antenna, than when it is horizontally polarised. The coupling is most sensitive at the resonant frequency of the antenna, which is around 75 MHz for a typical biconical antenna, and where the effect on antenna factor is less than 0.2 dB when the centre of the antenna is at a height of 2 m above the ground plane. The method relies on setting up a uniform wavefront at the location of the antenna and on minimising reflections from the antenna mount and the feed cable. A sufficiently uniform wavefront is achieved by situating a transmitting biconical antenna at a distance of 20 m with the tip of the antenna a few centimetres above the ground plane. This antenna and its cable must not move during the course of the calibration. The large overall size of the NPL ground plane, and its flatness, help to achieve a uniform field, including keeping edge diffraction small. There will be some taper of the field in the vertical plane, as illustrated by Fig. 18, but the same taper is experienced by the standard antenna and there will be some cancellation of its effect.



Figure 18. Field taper for vertically polarised calibration of biconical antenna.

The antenna mast is made of lightweight fibreglass tube and there is a minimum of metal parts, not being motorised. The antenna feed cable is routed horizontally several metres behind the antenna before dropping to ground. A calculable dipole antenna³³ is used as the standard, using an element resonant at 60 MHz to cover the frequency range 20 MHz to 100 MHz and an element resonant at 180 MHz to cover the frequency range 100 MHz to 320 MHz. It is useful to measure a little either side of the antenna operating frequency range because one can see trends in the antenna factor gradient. For example the antenna factor of some biconical antennas climbs sharply at 300 MHz as it approaches a resonant condition.

This method can be used to calibrate a well-conditioned biconical antenna, for example well built, with a 200 Ω balun and with the collapsible element configuration that is free of resonances. This antenna can be adopted as a transfer standard antenna, which in turn is used to calibrate customers' biconical antennas. Because the transfer standard is almost identical in size to the customer's antenna, the effects of minor site imperfections will cancel. NPL calibrates customers' biconical antennas in a fully anechoic room against a transfer biconical antenna, either with collapsible elements, or with the cage design. Uncertainties of less than ± 0.5 dB can be achieved. For the cage design with one cross bar, the asymmetry of the cross bar leads to an asymmetry in the H-plane radiation pattern above 260 MHz, which can lead to a change in antenna factor of 0.3 dB at 300 MHz so it is important that the cross bar orientation is defined²⁴.

An approximate value of AFfs can be obtained by measuring the antenna factor at various heights and then averaging the results. Fig. 19 shows the antenna factor at heights between 1 m and 4 m in steps of 0.5 m. These plots are calculated by NEC, but the measured results are very similar²⁴, especially the relative amplitudes of the plots. The free-space factor measured with the calculable dipole is superimposed and it can be seen that over most of the frequency range it lies in the middle of the other plots. A more accurate estimate of AFfs is to apply a non-linear least-squares method to the measured results in order to estimate the AFfs³⁵. ANSI C63.5 uses computed mutual coupling values to correct the antenna factor measured by the SSM so that it more closely approximates the AFfs²².



Figure 19. Antenna factor of a horizontally polarised biconical antenna at heights of 1m (black), 1.5m (red), 2m (yellow), 2.5m (green), 3m (cyan), 3.5m (blue) and 4m (magenta) above a ground plane, with the AFfs superimposed (dashed line).

9.3. Calibration of LPDA antennas

The types of LPDA antennas used for EMC measurements are commonly designed to operate over the frequency range 200 MHz to 1 GHz, though the upper frequency for some of these LPDA antennas extends as high as 5 GHz. There is another class of LPDA that covers the frequency range 1 GHz to 18 GHz that is not addressed in this section. These antennas have a gain of approximately 7 dB and so it is not difficult to create a good quasi-free-space environment in which to measure the AFfs. The best way to calibrate LPDA antennas is by the three-antenna method. The antenna is illuminated by a similar opposing antenna positioned at a sufficient distance to ensure an adequate approximation to far-field conditions. By carefully accounting for the phase centre position at each frequency, the antenna

separation can be as small as two wavelengths. NPL has achieved uncertainties in the antenna factor of better than ± 0.5 dB by using a small separation, which reduces the effect of ambient interference and also minimises the amount of RF absorber needed. In practice, a fixed separation of 2.5 m is used between the positions of the elements that are resonant at 300 MHz. To obtain the phase centre positions, the dimensions of the LPDA are measured and converted to a theoretical frequency profile of the antenna, which is given in terms of distance from the tip of the log periodic antenna, and then a polynomial is fitted to this data. These calculated phase centres are used in 3AM to derive the antenna factors.

For LPDAs that operate below 150 MHz the fixed height method of calibrating biconical antennas might be used. For example there is a commercial LPDA that covers the frequency range 80 to 1300 MHz, whose length is 1.7 m. Over a ground plane the distance to the opposing antenna should be at least 10 m and the LPDA height should be around 2 m, which is a compromise between minimising the mutual coupling to its ground plane image and ensuring that the reflected ray emanates from near the peak of the mainlobe of the LPDA. The antenna factor is computed using the LPDA phase centre positions.

9.3.1. Phase centre of LPDA antennas

The phase centre positions derived by the method described above have been confirmed by two other methods, the first of which was to model the antennas using the Numerical Electromagnetic Code¹⁴ and noting the position of the maximum currents. The second method was the traditional method of measuring the phase of the signal as the antenna was rotated – if the centre of rotation of the antenna coincided with its phase centre, the phase of the signal was constant with azimuth rotation of the antenna. The phase centre estimated from the mechanical dimensions was generally within 10 mm of the position found by these other methods, though at a few frequencies the NEC model showed high currents on more than one element, sometimes at twice or half the dominant frequency. This could imply that the phase centre is spread over a few centimetres. A well-designed LPDA will have a smooth monotonic increase of antenna factor with frequency and all the input power will be absorbed by the time it reaches the last element. Perhaps for a well-designed antenna the spread of phase centres is confined to a small proportion of the frequencies and the spread is small. If all the power is not absorbed, a good design solution is to place a resistor across the boom beyond the last element. This has the effect of removing some of the "humps" in the antenna factor versus frequency curve and stabilising the antenna performance. See also the last paragraph of Section 9.5 on conical log-spiral antennas.

Knowing the antenna factor at the phase centre of the antenna at a given frequency is necessary for the accurate measurement of field strength when the source is not very distant. The phase centre of a linear dipole is at the centre of the dipole, and for a horn antenna is near to the face of the horn aperture. However the phase centre of an LPDA antenna is clearly distributed along the axis of the antenna with change in frequency, but it is common practice for the antenna factor to be given in relation to the mechanical centre of an LPDA which coincides with the phase centre at only one frequency. The uncertainty is significant for distances of 3 m or less. In an NPL calibration certificate the AF is given for the phase centre at each frequency, but if another calibration laboratory gives it with respect to a single point of reference, one has to be sure that the antenna has been calibrated with a sufficient separation from the source for the phase centre uncertainty term to be negligible. This is in addition to the uncertainty referred to in the previous sentence when the antenna is being used

to make a measurement. If the distances used in the calibration and measurement are identical, the effect cancels and the uncertainty is zero. However this assumes that the calibration was done with another antenna whose phase centre is either constant with frequency, or the antenna separation used in the calculation is corrected for the change in phase centre.

An NPL certificate provides a formula for calculating the distance of the phase centre from the tip of the central column of the antenna, with constants specific to each model of antenna. These phase centre positions can be used to correct for field strength at the measurement distance, e.g. 3 m or 10 m, see Appendix 4. These are only accurate for free-space conditions, i.e. OATS or semi-anechoic rooms with absorber on the floor, because it has been found that phase centre determination has greater uncertainties when a second ray path from the signal source comes via a ground plane. The mechanical condition of LPDAs should be checked, before submitting them for calibration, and any bent elements straightened and all elements screwed tight. Appendix A4.2 warns the reader of the large errors that occur with the misuse of antenna factors measured by the SAE ARP958 method.

9.3.2. Cross-polar performance of LPDA antennas

Most LPDAs are constructed with the two half elements in echelon. This causes a decrease in cross-polar performance that is worst at the higher frequencies where the displacement of the two elements from a straight line is a greater proportion of the element length. The CISPR 16-1-4 criterion of 20 dB for cross-polar discrimination is too severe for some popular models of LPDA. Whilst this level may be achieved below 500 MHz it can range from just on 20 dB to as low as 14 dB at 1 GHz for different models. One manufacturer has designed an LPDA whose elements are all in the same plane: this is achieved by putting a kink in each element near to its join with the boom, and it has achieved an excellent cross polar performance, better than 20 dB up to 1 GHz. Other manufacturers have tapered the boom so that its line separation is smaller at the higher frequency end, thereby reducing the separation between the planes incorporating the elements each side of the boom. In one exceptional case the cross polar signal at 5GHz is higher than the co-polar signal.

It is a fairly elaborate process to measure cross-polar response to better than 20 dB, and therefore costly. It involves finding an opposing antenna which itself has a cross polar response of better than 40 dB. A linear dipole can be used, but it is more practicable to use a horn antenna whose directivity is a big advantage in reducing reflected signals, and it generally covers a wider bandwidth. The setup, including the site and antenna towers, must be free of reflections of the order of 40 dB down. One of the antennas has to be rotatable though greater than 90°.

9.4. Calibration of hybrid biconical / LPDA antennas

Essentially the biconical and LPDA antennas have been stitched together to create a hybrid antenna that covers the frequency range 26 MHz to 2 GHz. The cut-off frequency of the logperiodic section is determined by the longest "log" element. There is one larger broadband dipole element just behind this, which is attached to the boom via a balun, and this works down to the lowest frequency of operation. Computer optimisation was used for some models to remove sharp resonances at the frequency interface between the two sections and to simplify the design of the large elements, and to taper the element lengths up to the highest frequency. The hybrid antenna enables EMC test laboratories to cover the existing radiated emission frequency range 30 MHz to 1000 MHz in one sweep, which is very convenient.

A word of caution about the use of the three-antenna method for the calibration of LPDA antennas that uses a single reference for the phase centre – a fixed position approximately mid-way between the mechanical extremities of the antenna is commonly used, and some manufacturers place a mark on the antenna. This caution is especially relevant to the longer hybrid antenna and for an antenna separation of only 3 m used in the calibration. It is assumed that the calibrated antenna is to be used to measure a source (e.g. an EMC emission) at a defined distance. However the opposing antenna (e.g. another hybrid antenna) used in the calibration has that distance defined only at one point, i.e. one frequency. At lower frequencies the field strength will be measured at a greater distance, and at higher frequencies it will be measured at a smaller distance. These changes in distance are not accounted for in the determination of the antenna factor of the AUT. So for example, if the hybrid antenna is 1 m long and has a reference point in the centre, and the emission measurement is set up for a distance to the EUT of 3 m, the field strength at the upper frequency, corresponding to the tip of the antenna, could be underestimated by 20 $\log(2.5/3)$ or 1.6 dB.

To counter the potential for this error, at NPL broadband hybrid antennas are calibrated in two stages using the methods of Section 9.2 and 9.3. Instead of using a broadband hybrid as the antenna opposing the AUT, over the frequency range 30 MHz to 200 MHz a biconical antenna is used (SAM) and over the range 200 MHz to 2 GHz an LPDA is used (3AM). An antenna separation of greater than 10 m is used, but for the upper part of the range it is also possible to use the separation of 2.5 m as described in Section 9.3 providing the antenna factor is corrected for variation in phase centre at each frequency.

9.5. Calibration of conical-log-spiral antennas

The types of conical log-spiral antennas used for EMC measurements are commonly designed to operate over the frequency range 200 MHz to 1 GHz, though some commercial spirals start at 100 MHz and some operate over the range 1 GHz to 10 GHz. Spirals are usually seen in military standards. Spirals are not called for in CISPR standards, where the preference is for two separate measurements using vertical and horizontal polarisation - with a single measurement using a spiral it is possible that a signal could be missed if the source happened to emit the opposite sense of polarisation. This is the same reason that a single measurement is not made with a linear dipole inclined at a 45° to the vertical, because the source could possibly be emitting at -45° .

The AFfs of spirals can be measured either on OATS or in a fully anechoic room by the three-antenna method. The reflection from the ground plane changes to the opposite sense of circular polarisation and, if the antennas have an *ellipticity* near to unity, the receiving antenna does not "see" the reflected signal, so effectively the measurement has been performed in free-space and the free-space simultaneous equations can be used to convert the three SIL readings into antenna factors. The uncertainty of AFfs depends on the quality of the ground plane, but it is possible for this uncertainty component to be very small - for example, the uncertainty of the NPL ground plane is estimated to be less than ± 0.05 dB up to 1 GHz, from comparisons of measured and predicted SIL between two calculable dipole antennas.

The phase centre position can be derived from a simple formula based on axial length of the antenna incorporating the ends of the wire spiral at which the radiation is strongest at each end of the operating frequency range, with a slight increase in uncertainty compared with the more precise phase centre determination of LPDA antennas. The uncertainty in the antenna factor due to the error in predicting the correct phase centre is reduced the greater the separation of the antennas in 3AM. Standard calibration geometry is to set the spirals 10 m apart at a height of 2 m above the ground plane. At a height of 2 m there is negligible influence of mutual coupling to the ground plane image. A spiral calibrated in this way can be designated as a transfer standard antenna and used to calibrate similar spirals by the standard antenna method in a fully anechoic chamber, with the opposing spiral antenna approximately 3 m away. An extra \pm 0.5 dB uncertainty is usually added to the uncertainty of the antenna factor for a spiral calibrated in this way, to account for the imperfect field uniformity in the chamber, for the smaller antenna separation and any differences between the transfer antenna and the AUT.

The sense of polarisation of the antenna is right-handed if the electric vector transmitted by the antenna rotates clockwise for an observer looking in the direction of propagation for a given position in space. It is important not to confuse this definition with the definition used in optics, where the observer looks towards the source, and results in the opposite sense.

A simple test of the ellipticity of spiral antennas is done by taking the difference of the responses to a linearly polarised LPDA when oriented horizontally and vertically polarised. The LPDA should have a cross-polar rejection of more than 20 dB. The antenna factor is obtained in a circularly polarised field; however when using spiral antennas for emission testing, it is assumed that the field from the equipment under test is linearly polarised, in which case a 3 dB correction has to be added to the antenna factor in order to simulate what might be measured by a linear antenna. An additional uncertainty must be included to take account of departures from true circular polarisation, and a typical value is ± 0.5 dB.

There is a desire by all parties to keep antennas as compact as possible. However log-periodic antennas, by their nature, work best in the region of the antenna that is bounded by log-periodic radiators. Especially at the lower end of the frequency band if the antenna is cut off at the point where it is giving a reasonable antenna factor, the behaviour of the antenna may suffer. For a conical log-spiral designed to operate from 200 MHz upwards, if the antenna factor rises sharply just below 200 MHz, it means the antenna has been cut off too soon, which results in poor ellipticity between 200 MHz and about 250 MHz. For satisfactory performance at 200 MHz it would be better to use an antenna that has been designed to start at 150 MHz, or to use the design already available that is cut off at 100 MHz. The radiation pattern also suffers and this applies also to LPDA antennas; which is a reason to increase the uncertainty of the antenna factor in the lowest part of the operating frequency range.

9.6. Calibration of antennas for specified measurement geometries

Life would be simple if the emission were being measured in free-space conditions, solely in the boresight direction of the antenna, in the far-field of both the antenna and the emitter and if the emission had sufficient magnitude to be well above the noise floor of the receiver, in a well matched system, and there were no other unwanted signals present. However it is rarely possible to meet all of these conditions simultaneously, especially in EMC measurements. So

we end up with different standards being written, notably by CISPR, the automotive industry and the military. In turn these lead to a variety of methods of calibrating antennas in order to get the most useful measurement, in terms of *accuracy* and/or reproducibility.

Two measurement geometries need special attention for accuracy and reproducibility to be meaningful. The most obvious one is the measurement of emissions at frequencies as low as 30 MHz at a distance of 1 m and a consequent standard that calls for antennas to be calibrated using a separation of 1 m. Of course, very precise field measurements can be made if the field sensor is small enough and is not coupling with the emitter or perturbing the field, but there will be large uncertainties in the prescribed method of using a 1.4 m long and 0.5 m wide biconical antenna up to 200 MHz for such a measurement.

The other example is the 3 m OATS geometry in which the antenna is scanned in height through the range 1 m to 4 m. When the antenna is at a height of 4 m the distance between the EUT and the antenna is no longer the intended 3 m but is more than 4 m and also the EUT will no longer be in the boresight direction of the antenna. Actually, for the greater part of an EMC test, a maximum signal is located below 2.5 m and some antennas have a mechanism to point down as the height is increased, nevertheless these extremes will be encountered at some combination of frequency, polarisation and fixed pointing antenna. Generic or product standards specify emission limits for a distance of 3 m, which are clearly being infringed: the fall off of field strength from 3 m to 4 m is 2.5 dB. Most test houses do not use a pointing mechanism for their antennas, so for an LPDA antenna with 7 dB gain there will be significantly less field being picked up, especially from the ground reflection, when the antenna is higher than the EUT. There is also the issue that the phase centre error is quite small at a range of 10 m, but becomes of the order of 1 dB at a range of 3 m, and is much more for a hybrid antenna which is typically 1 m long (in the direction of the EUT). An antenna calibration method has been devised that mitigates some of these effects and gives a result that is closer to the intention of the standard, in this case knowing the field strength at a distance of 3 m from the EUT. This method is described in Section 9.6.4 as a geometryspecific method.

9.6.1. Measurement distance

Historically EMC radiated emissions were to be measured at a distance of 30 m. This gave a separation of three wavelengths at the lowest frequency of 30 MHz, and the field taper across a 4.8 m long tuned dipole antenna would be insignificant, giving acceptable far-field conditions. When a metal ground plane was mandated there was an objection to using a 30 m distance because a very large ground plane was needed, but primarily 30 m fell into disuse because the field strength was too small to measure and/or ambient interference was swamping it. So 10 m became the accepted test distance. The US Army Electronics laboratory at Fort Monmouth, NJ, invented the biconical antenna which was only 1.4 m long and a suitable replacement for the tuned dipole between 30 MHz and 200 MHz. In view of the large uncertainties associated with EMC measurements, the small uncertainty associated with making a far-field measurement at a distance of one wavelength (10 m at 30 MHz) was no problem.

Later on a distance of 3 m became acceptable, technically because the field strength limits were so low that it was not sufficiently above the noise floor of the receiver at 10 m distance. However 3 m was liked by test labs because they could use a much smaller ground plane and a smaller piece of real-estate to set up an OATS. When semi and fully anechoic rooms appeared on the horizon 3 m was favoured because a *SAC* or FAR accommodating a 3 m site

was less than a third of the cost of one specified as a 10 m site. However it is debatable whether results of emission testing at 3 m and 10 m are comparable, even allowing for a correction of field strength with distance³⁶. A further point is that knowledge of the field strength at 3 m is more meaningful, than at 10 m, for a lot of products which need to coexist in a close environment, such as a home. Sometimes one does not want to know the far-field emission of the product, but its effect on other products in its near-field, so it is more appropriate to measure at a closer distance, which is an issue to be addressed by standards committees.

In early *EMI* testing the test setup simulated the actual equipment environment in so far as was practical, see first paragraph of Section 9.10. In the late 1940s came the military use of the 1 m test distance. The close measurement distance better simulated the real-world situation in military environments and became a habit in military standards in the 1950s to the 1970s. Another factor was that in some cases a very low emission limit was required and a distance of 1 m was adopted to get a better signal to noise ratio. The biconical antenna was more realistic to use in the shielded enclosures than the tuned dipole, certainly below 100 MHz where the tuned dipole is longer than the biconical antenna.

The aerospace contingent of the International Society of Automotive Engineers wrote procedure³⁷ ARP958 for the calibration of antennas with a separation of 1 m. These aerospace standards were developed particularly in support of the military EMC requirements. Many of them were adopted by the land-vehicle committee later on. As time went on, the land-vehicle committee took several of the ARPs and rewrote them for land vehicle use. Sometimes this included a greater measurement distance.

Some of the first measurements that the FCC did were to measure the strength and pattern of AM radio broadcast stations from about 540 to 1605 kHz. For this, the FCC used a shielded loop antenna a mile or so from the radio broadcast antenna. The plane of the loop was vertical and its edge was pointed at the broadcast antenna. This caused it to respond to the near field component that would become the magnetic field component in the far field. The shielded loop was preferred because it was not very sensitive to its surroundings. During a measurement, one could walk up to the loop and put one's hand on it without changing the reading on the instrument.

In the measurement of some types of *ISM* devices, the FCC required using a "balanced doublet" antenna at frequencies above 26 MHz, and a tuned dipole above 30 MHz. These measurements were usually made at the boundary of the property on which the ISM device was located, so the measuring distance could be anywhere from a few hundred feet to a mile or so. As the FCC began to regulate more RF devices, it stayed with the loop antenna for measurements up to 30 MHz and the tuned dipole from 26 MHz up. This was a problem for industrial equipment test personnel, since the measurements using the two kinds of antennas in the range from 26 - 30 MHz seldom were in agreement.

9.6.2. Uncertainty of measurement in the near-field

There is a common misunderstanding of the requirement to make a measurement in the farfield of the emitter, and how to deal with the uncertainties when not in the far-field. Just because the measurement is made in the near-field of the emitter does not mean that there is a large uncertainty associated with that measurement. The measurement can be made very precisely, but it is a question as to what one intends to do with the result. If one is happy with knowing the electric field strength at that distance, the uncertainty can be very low, but the point that many people overlook is what happens when attempting to extrapolate that result to another distance: this is what increases the uncertainty, avoidance of which, one may have to measure the phase and amplitude of both the electric and magnetic fields in three orthogonal polarisations and possibly make many measurements over a grid of points.

Having said that, when the source is known to be a small electric dipole and the sensor is small and does not significantly perturb the field, electric field strength can be extrapolated from a near-field measurement to a greater distance with fair precision using the expression for distance in A1.8 of Appendix 1. If the source were a magnetic dipole the error could be very large, but in this case the magnetic field could be measured precisely with a small loop antenna. As a rough guide, the distance has to be greater than $\lambda/2\pi$ in order for the error not to be large when measuring electric field from a magnetic source and vice-versa. The nature of the source in an EUT can be roughly determined by the behaviour in the fall-off of field with distance up to half a wavelength from the source: if a small electric dipole is used to measure the EUT and the field strength falls off according to the expression in A1.8, then the EUT predominantly comprises an E-field source. Similarly, if a small loop antenna is used and the field falls off as predicted (magnetic version of equation in A1.8), the EUT must be mainly acting as a magnetic dipole source.

9.6.3. Customer expectations regarding calibration distance

Section 9.6.1 highlighted that EMC standards cite measurement distances of 30 m, 10 m, 3 m and 1 m. A common request to a calibration laboratory is for a calibration of an EMC antenna at 10 m, the customer believing that the calibration method has to involve a separation from the opposing antenna of 10 m. For the purposes of clarity assume that the request is for the measurement of AFfs, which is the default parameter for an antenna and which is measured in the far-field. In the case of an LPDA antenna that has been designed for the frequency range 200 MHz to 1 GHz, a sufficient accuracy for a far-field measurement occurs at a minimum distance of two wavelengths, which implies a separation of 3 m between the longest dipole elements of a pair of such antennas. There will be no significant difference in this value if the measurement were done with any greater separation; it is a misunderstanding to assume that the antenna has to be measured at a distance of 10 m, and it means that the laboratory can choose a more convenient distance.

Some antenna manufacturers distinguish, in their certificates of calibration, between antenna factors for measurements at 3 m distance and at 10 m distance. This distinction is based on the premise that measurements made in the near-field of an EUT require an antenna that has been calibrated in the near-field of a source antenna. In contrast CISPR 16-1-4 states that the appropriate calibration of antennas for emission testing is the free-space antenna factor and no distinction is made between its use at a distance of 3 m or 10 m. This is the position taken by NPL.

Sometimes it does make a difference whether the antenna is calibrated using a separation equal to the EMC test distance, or at some other distance. Taking an extreme example, if the customer is using a tuned dipole at a distance of 3 m, the dipole length is 4.8 m at 30 MHz and normally a calibration laboratory would use a separation to the opposing antenna of 20 m or more. However the far-field antenna factor may not be the appropriate value to use for an EMC test at 3 m because the wavefront from the EUT across the antenna will be tapered. The appropriate calibration depends on the size of EUT to be tested. Say the EUT were no more than 1 m wide, it would be appropriate to calibrate the 4.8 m long dipole using an opposing antenna that was 1 m long. If the opposing antenna were another 4.8 m dipole there would be

significant mutual coupling and a different amplitude and phase taper compared with that experienced with the EUT. Calibration of a 4.8 m dipole with an opposing 1 m dipole is a straightforward task using calculable dipole antennas. Similar issues apply for the calibration and use of antennas with a separation of 1 m.

There are laboratories that use tuned dipoles on 3 m test sites, but the majority of laboratories use biconical antennas, whose length of 1.4 m is not a problem. In the main, the advice that can be offered is that radiated emission testing at a distance of 10 m is relatively free of antenna related uncertainties, even with large hybrid antennas, and a calibration laboratory can use their favourite methods for obtaining AFfs, such as calibrating an LPDA using a separation of 3 m and a 30 MHz tuned dipole using a separation of 20 m. By and large for EMC testing at 3 m the uncertainties can be kept small if a conventional biconical antenna is used from 30 MHz to 250 MHz and an LPDA antenna is used from 250 MHz to 1 GHz. The reason for this choice of transition frequency is that above 250 MHz the biconical radiation pattern starts departing from its cardioid shape and below 250 MHz the length of the LPDA antenna contributes significant phase centre errors. Virtually all the relevant EUT emission results will be made with the antenna below a height of 2.5 m, thereby containing the distance and radiation pattern errors. At 30 MHz a distance of 3 m is a third of a wavelength; contrary to what is popularly believed there is not a large near-field error, it is less than 1 dB when using a biconical antenna. If the larger hybrid antenna is used, it may be more appropriate to use a geometry-specific calibration as described in the next section.

9.6.4. Geometry-specific calibration for 3 m OATS measurement

As discussed in the previous section, where the measurement distance or the antenna size may give rise to significant measurement uncertainties, it may be beneficial to calibrate the antenna by a method that resembles the way the antenna is going to be used. This implies using the SSM with careful consideration of the type of opposing antenna and its fixed height chosen relative to the height of the EUT in an emission test. The use of a hybrid antenna on a 3 m OATS is a pertinent example. There are various models of hybrid antenna on the market that range from 0.8 m to 1.5 m separation between the smallest and largest dipole elements: a typical length of 1 m between the elements excited at 30 MHz and 1 GHz is assumed here for estimates of uncertainties.

The rationale for a geometry-specific calibration is that the antenna factor is being measured in a way that is very similar to the way the antenna is being used in an EMC test. This should take care of three main potential sources of uncertainty: firstly the variation of phase centre with frequency, secondly the variation in the distance between the EUT and the antenna as the latter is height scanned, and thirdly the deviation of antenna factor from the boresight value as the antenna is raised and the antenna factor decreases according to the shape of the radiation pattern. Another lesser factor is that for the lower frequencies where there is a small near field error, this will also be taken into account by this method, as long as it is not intended to extrapolate the result to another distance.

For clarity, assume that the EUT is a point source, set at a height of 1 m above the ground plane. The SSM will yield the correct antenna factor across the frequency range if the opposing antenna is a small antenna likewise set at a height of 1 m and at a distance of 3 m from the reference point on the hybrid antenna. A candidate small antenna is a mini biconical antenna that is less than 0.35 m long and that operates over the frequency range 30 MHz to 1 GHz. This method will eliminate the four sources of uncertainty mentioned, but there will still remain the usual small uncertainties in setting the distance of 3 m and the fixed height of

1 m, and the height scanning from 1 m to 4 m. This method achieves an antenna calibration that gives the E-field corrected to 3 m distance from a small EUT at 1 m height.

However, calibration laboratories are assumed to take the natural course, which is to use another hybrid antenna as the opposing antenna, and yet another to make up the three antenna pairs. This has serious consequences on the uncertainty of the antenna factor because it introduces phase centre and radiation pattern errors that do not exist on the small EUT. For a pair of hybrid antennas 1 m long on a 3 m range, the antenna factor is underestimated by 1.6 dB at 1 GHz and overestimated by 1.3 dB at 30 MHz, these figures being reduced the nearer the phase centre is to the reference point on the antenna. If the height scan maximum were 2.5 m, the antenna factor would be underestimated because of the reduced directivity in the direction of the ground signal, by, say, 0.5 dB depending on the radiation pattern at a particular frequency. A big improvement on this would be to use separate biconical and LPDA antennas as opposing antennas instead of a hybrid, but best of all would be to use the mini biconical antenna. One drawback of the mini biconical is its large antenna factor, greater than 50 dBm⁻¹, at the low end of the frequency range.

Another problem is endemic to the SSM method, which is that a real EUT does not normally have a defined centre of radiation and it could be at any point along the height of the EUT. In other words the SSM is tailored to the point of radiation of the chosen opposing antenna at a particular height. So the antenna factor is only strictly valid if the radiation is coming from the EUT from only one point and at that height. For any other point there will be an uncertainty associated with the antenna factor. But for an EUT of 0.4 m height on a table of 0.8 m height this uncertainty will be small, because at the height where the receive antenna identifies the maximum signal, the maximum signal value will not be very different from that if the height of the source were 1 m. This is really a caution as to the uncertainty in using geometry-specific antenna factors when measuring tall EUTs.

9.6.5. Calibration of antennas using 1 m separation - comment on the ARP958 calibration method

SAE³⁷ ARP958 was originally written in 1968 for the calibration of conical log spiral antennas only, from 200 MHz to 1 GHz, also see Appendix 2, section A2.2. It is called up in many standards as the required method of antenna calibration. The antenna separation of 1 m is based on an Aerospace/Automotive requirement. The method involves measuring the insertion loss between two identical antennas with a separation of 1 m from their tips. The antenna factors are calculated by using a 1 m distance in the formula for calculating the gain product from the measured SIL (see Friis formula in Section 9.9). The intention is that the antenna factor will enable a reading of field strength at a distance of 1 m from the source, and this could be the case if the opposing antenna were a flat spiral antenna, but, as argued in the preceding section, the phase centre variation of an opposing similar antenna will add an uncertainty to the field reading. In this case the distance is measured from the tip of the antenna, so the uncertainty will be negligible at the upper end of the frequency range, but it will be a maximum at 200 MHz of 0.7 dB for an antenna of length 0.63 m, i.e. the antenna factor will be too low, so the field strength will be underestimated by 0.7 dB.

Furthermore, if the antenna were to be used incorrectly to measure an EUT at 10 m distance, the antenna factor at 200 MHz would be too high by 3 dB. The same argument applies to LPDA antennas included in a later version of the standard. The presence of these substantial uncertainties excludes this procedure for general measurement of field strength, but the SAE would argue that their method achieves reproducible results. If an antenna is to be used for

both 1 m and 10 m measurements, it is important that the correct antenna factors are used, in other words ARP958 AF and free-space AF respectively. For log periodic designs (including spirals) it is possible to derive a 1 m AF from AFfs using the phase centre information.

A later version of the standard included biconical antennas over the frequency range 20 MHz to 200 MHz and resonant dipoles, which are commonly supplied for the range 25 MHz to 1 GHz. It would be a strange measurement, subject to large uncertainties, if a pair of dipoles 5.8 m long at 25 MHz were placed 1 m apart in order for one of the dipoles to then be used to make a measurement of something entirely unlike a 5.8 m resonant dipole. In practice most people use the biconical antenna, but this has a diameter of 0.5 m, so the extremities of two biconicals are only 0.5 m apart when their centres are 1 m apart, leading to significant mutual coupling. Also at 20 MHz there is a strong near-field effect; the AFfs and the 1 m antenna factor of a typical 50 Ω biconical antenna are compared in Fig. 20, with differences of over 4 dB.



Figure 20 Comparison of free space AF and ARP958 AF_{1m} (upper plot) of a typical biconical antenna with a 50 Ω balun. Note 4 dB difference at 20 MHz.

NPL provides a service for measuring according to ARP958. However with its knowledge of calibrating LPDAs with small separation, a short cut is to calculate AF_{1mARP} from AFfs using the phase centre information derived from the mechanical dimensions of the antenna (this works very well for well designed antennas that have a smooth monotonic increase in AF with frequency). The formula for doing this is given in Appendix 4, clause A4.4.1.

There is another type of 1 m AF that is to be distinguished from AF_{1mARP} . Some standards call for the field from an EUT to be measured at a distance of 1 m. For an LPDA the field strength that it measured is at the location corresponding to the resonant element at a given frequency. At frequencies above 200 MHz it is valid to extrapolate this field value to position of the tip of the LPDA. If these corrections are added to AFfs, an AF_{1m} has been created,

solely to be used for this purpose. The formula for doing this is given in Appendix 4, clause A4.4.2.

9.7. Calibration of antennas for site evaluation

The methods of site evaluation are described in Section 4.2. This section describes ways to calibrate antennas to give the lowest uncertainty for site evaluation by each method. The very lowest uncertainty is achieved by using calculable dipole antennas, which by their nature do not need calibrating in the normal sense. They just need checking mechanically to ensure that they are in working order, the S-parameters of the baluns have to be measured, and it is advisable to perform an SIL test to ensure there is no breakdown of electrical continuity.

For the NSA procedure the antennas should be measured by the SSM using the same antenna geometry as for the NSA measurement. In particular, because the antenna factor uncertainties can be large on a 3 m site, the antenna factors should be measured by the SSM using 3 m separation. These AFs will be optimum for an NSA measurement for a 3 m site; most of the errors will cancel because the same height-scanning geometry is being used for both the SSM and for the measurement of NSA; see Section 9.6.4 for further elucidation.

The NSA procedure involves measuring SA between a pair of antennas, one of which is height-scanned and the other set at a fixed height. Normally the AFs of both antennas are obtained via the SSM without distinguishing the scanned or fixed height, but the uncertainty can be reduced by using the SSM for the height scanned antenna and a special fixed-height calibration for the fixed antenna. The uncertainty can be reduced further by using the three-antenna method with both antennas at fixed heights, to avoid the errors associated with calculating three AFs from four unknowns, as explained in Section 9.2.1; however several combinations of fixed heights are needed to cover a broad frequency range whilst ensuring the signal is near a maximum, i.e. well clear of a null.

To obtain the dual antenna factor, just one SA measurement is needed using the SSM. The product of the antenna factors is calculated²³, since this is what is needed to calculate NSA.

The beauty of the SACM, or the Site Reference Method, is that no antenna calibration is required, with a consequent minimisation of uncertainties. The SA between a pair of antennas is measured on a reference site (or CALTS) and the measurement is simply repeated in an identical manner on the site-being-evaluated.

9.8. Considerations on the measurement of radiation patterns of low gain antennas

In general radiation patterns of EMC antennas are not needed to be known with high accuracy. But it is useful to know the frequency limit when the expected pattern starts degrading or breaking up. For example many models of biconical antenna show a marked change in pattern shape around and above 300 MHz. The front-lobe to back-lobe ratio of LPDA antennas decreases at frequencies below the half-wave resonant frequency of the longest element, where they are no longer behaving in a log-periodic manner. Facilities to measure sidelobe levels of satellite antennas 50 dB down from the mainlobe are very costly,

but there is rarely a need to know accurately the sidelobe or null levels of antennas for EMC applications more than 20 dB down from the mainlobe level.

An environment is required in which the magnitude of the direct signal between the transmit and receive antennas is at least 25 dB greater than any indirect signals arriving from reflections. A single 20 dB reflected ray gives an error of ~1 dB on the direct ray; one has to allow for multiple reflected rays adding in phase or anti-phase with the direct ray. Because the transmit antenna has low gain, it is possible to measure the pattern with a small separation between it and the receive antenna. Say this separation were 2 m, even a perfect reflector only has to be ~18 m away for the reflection to be 25 dB down; the reflector can be nearer if the receiving antenna has greater directivity (assuming the antenna under test, AUT, is transmitting). The facility could be either an open area site with a low reflection tower, or an anechoic chamber that is lined with high quality radio frequency absorber. The room would have to be large for biconical antennas, but much smaller for LPDA antennas. The size is dependent on frequency and directivity of the AUT and on the reflection coefficient of the absorber. Probably biconical antennas are best measured on an OATS with towers, whereas an LPDA or horn antenna could be measured in a chamber with dimensions of at least 5 m x 5 m x 5 m. This would also be suitable for a small biconical antenna designed to operate in the frequency range 1 GHz to 18 GHz.

To validate this environment, perform a standing wave measurement along three orthogonal axes. The centre of the three axes coincides with the position taken by the transmit antenna when its radiation pattern is being measured, i.e. about the axis of rotation of the azimuth positioner. There is a compromise between keeping the antenna separation as small as possible to limit the effect of reflections, and doing a valid standing wave measurement in which the extent of travel of the antenna lies within the flat part of the main beam of the receive antenna. The standing wave measurement shall be done at a sufficient number of intervals across the required frequency range and the extent of travel shall ensure that at least one full wave is recorded from which to take the peak-peak reading. This method needs to be tried to test whether 2 m is sufficient separation: the beamwidth of the receive antenna reduces with increase in frequency, so does the extent of travel required to produce one standing wave, so these factors work favourably together. The magnitude of the sum of reflections can be found from the peak-peak value using the expression²⁶:

$$S(dB) = 20 \log_{10} \left[\frac{\frac{p}{-1 + e^{20}}}{\frac{p}{1 + e^{20}}} \right]$$

where S is the level of the reflected signals below the direct signal, and p is the peak-peak magnitude in dB. For example, where the peak-peak ripple is p = 1 dB, the reflected signal is suppressed by S = -25 dB.

The CISPR standard¹⁷ for EMC emission measurements above 1 GHz requires the 3 dB beamwidth of the antenna to be measured. If an uncertainty of less than \pm 1 dB is required for the location of the 3 dB points on the mainlobe, the rough guideline above on room size may need to be reassessed.

9.9. Calibration of horn antennas

One of the challenges for an antenna calibration laboratory is the wide variety of antennas presented for calibration, with respect to frequency of operation, mounting arrangements, polarisation and also feed types. The common frequency coverage for horn antennas used for EMC testing is 1 GHz to 40 GHz. It is desirable to adopt a measurement system that is flexible and readily reconfigured; otherwise too much time is spent changing equipment setups to meet the varying requirements.

The use of an automated network analyser with a frequency converter test set enables measurements to be made over the entire frequency range of interest in one sweep and the analyser can be readily reconfigured to make reflection coefficient measurements. Depending upon the test set and source available, measurements can be made up to 18 GHz or even 26.5 GHz. As an alternative, a spectrum analyser with tracking generator could be used, but it would only be possible to measure the magnitude of the reflection coefficients of the antenna in that case. At higher frequencies, the loss in the coaxial cables and variation in that loss with cable flexure make the use of remote mixers or diode power sensors advantageous.

The aim of the measurements is to obtain a value for the realised gain of the antenna when terminated in 50 Ω , so it is helpful to include matching pads in the measurement circuit, so that any mismatch correction is kept to a minimum. It is important to assess the match of the pads used, since not all pads have a very good match and this will lead to a mismatch error. The amount of attenuation should be chosen to optimise the dynamic range of the measurement system. One must ensure that the received signal, b_2 , remains below the compression level during the direct connection, but that the reference signal, a_1 , is sufficiently high to retain phase lock. A useful feature on some receiver systems is the ability to apply a slope to the source power. This can be used to good effect to help to compensate for the increased cable loss as the frequency increases.

Since most antennas used for EMC measurements employ sexed connectors, an adaptor is required to connect the transmit and receive cables with their matching pads together. An example layout is shown in Fig. 21 below, in which the matching pads on the transmit and receive ends have been connected by an adaptor. The purpose of this setup is to obtain a reference measurement, so that the insertion loss can be measured when the antennas are connected later. This "direct-connection" measurement characterises the losses in the connecting cables and pads, the coupling factor of the directional coupler and the conversion loss of the frequency converter test set.

We can define the direct-connection ratio, k(f), as:

$$k(f) = \left|\frac{b_2}{a_1}\right|^2$$



Figure 21. Schematic of the direct connection measurement

Note that k(f) is a function of frequency, so it must be measured across the frequency range of the antenna. For a frequency range of 1 to 18 GHz, a frequency step of 100 MHz is typically used, giving a total of 171 points. It is very important that nothing in the measurement system changes between making the direct connection measurement and performing the measurements which follow. If the measurement system is unstable, perhaps due to worn out cables or poor connections, then the subsequent measurements will be erroneous.



Figure 22. Photograph of the direct-connection measurement

Fig. 22 shows the Tx (transmit) and Rx (receive) matching pads connected via a Type-N male to male adaptor for the direct-connection measurement. The Tx carriage is on the right and the directional coupler is clearly visible with the reference cable attached to the coupled arm.

Having measured the direct-connection ratio, the next step is to remove the adaptor and connect the transmit and receive antennas and adjust their separation to the required range (typically 1 m or 3 m). It is important that the antennas are aligned so that their principal axes are collinear, since horn antennas are far more directive than biconical or dipole antennas. It is also important to ensure that the antennas are reasonably polarisation matched, though the sensitivity of the received signal to polarisation mismatch is quite small. Fig. 23 shows the schematic setup for the transmission measurements, with the antennas separated by distance R, while Fig. 24 shows a photograph of the same thing.



Figure 23. Setup for transmission measurements

If the realised gains of the transmitting and receiving antennas are Ga and Gb respectively, then, ignoring the residual mismatch losses and the loss in the adaptor used during the direct connection,

$$\left|\frac{b_2}{a_1}\right|^2 = k(f) \operatorname{GaGb}\left(\frac{\lambda}{4\pi R}\right)^2$$



Figure 24. Transmission measurements with ridged waveguide horns set 1 m apart

This measurement must be repeated, having replaced the receiving antenna by a third antenna with gain Gc. Finally, the transmitting antenna is replaced by the antenna that originally received, to give the last possible combination of antenna pairs in the three antenna method. The measurements are summarised in the Table 2 below:

| • | interna comonations when canorating meanly polarised anternas | | | | | | | |
|---|---|------------|------------|--------------|--|--|--|--|
| | Measurement | Tx antenna | Rx Antenna | Gain Product | | | | |
| | 1 | А | В | Gab | | | | |
| | 2 | А | С | Gac | | | | |
| | 3 | В | C | Gbc | | | | |

Table 2. Antenna combinations when calibrating linearly polarised antennas

From these 3 measurements, the gains of each of the antenna may be obtained, for example:

$$Gc = \left(\frac{GacGbc}{Gab}\right)^{\frac{1}{2}}$$

or, expressed in dB:

$$Gc = 5 \operatorname{Log}_{10} \left(\frac{GacGbc}{Gab} \right) \quad [dB]$$

Once the three transmission measurements have been made, it is important to repeat the direct-connection measurement to ensure that there has been no significant drift in the measurement system and to demonstrate that the measurement is repeatable. As a rule of thumb, it should be possible to achieve repeatability of ± 0.03 dB or better, provided good quality connectors and cables are used.

By performing a full three-antenna calibration rather than a substitution measurement, one has some check on the validity of the measurements, provided at least one of the antennas has been measured before. This is valuable in helping to avoid blunders that can easily occur if, for instance, one of the connections was poorly made, or the wrong power level used, resulting in compression. However, it is important to note that, just because the gains of the two antennas provided by the calibration laboratory agree with historical data, this does not, in itself, prove that the values obtained for the antenna under test are correct. It could happen that the connection to this antenna was faulty, which would have applied to both measurements in which it was used. Provided the connections to antennas A and B were good then the error will only be apparent in the gain of antenna C, but if this is unknown, it would have to be a gross error to be noticed.

The measurement system described above is suitable up to about 26.5 GHz. Above this frequency, cable losses become unacceptably high and it may become difficult to maintain phase lock from the reference signal at the higher frequencies as the power dispersion increases. For measurements in coax up to 50 GHz, the receiver system can be replaced by diode power sensors, but their speed of operation is quite slow by comparison.

At still higher frequencies, remote harmonic mixers may be employed, with a mm-wave source module to provide the test signal. The advantage of this arrangement is that the LO (local oscillator) drive to the mixers is provided at a relatively low frequency, between about 3 and 6 GHz, with the IF (intermediate frequency) at 20 MHz. This dramatically reduces the loss in the system. Great care is needed to ensure that the mixers are operated at the optimum level to avoid compression while still maintaining sufficient dynamic range for the measurement. By comparison to the coaxial system described above, these measurements are far more time consuming to perform, mainly due to the increased setting up time required.

Fig. 25 shows the realised gain as a function of frequency for a double ridged guide horn antenna measured at 1 m from the aperture using the method described above.



Figure 25. Graph showing the realised gain at 1 m for a DRG (double ridged waveguide) horn antenna

To measure an antenna that is nominally circularly polarised (CP), the approach recommended by SAE ARP958 is to use the two identical antenna technique. However, as noted below, it is not always possible for a calibration laboratory to obtain a second antenna that is sufficiently identical; proving that it is identical is likely to be more time consuming than performing a full three antenna calibration in any case.

There is a further problem with calibrating circularly polarised antennas: they come in leftand right-handed varieties. This doubles the number of antennas the calibration laboratory requires to be able to deal will all antenna varieties using this method. There is an alternative approach, however. One can use two linearly polarised antennas in conjunction with the nominally circularly polarised antenna under test to perform an extended three-antenna technique calibration. The transmission measurement for the two linearly polarised antennas is exactly as described earlier, but when the CP antenna is used, two transmission measurements must be made for each antenna combination, one with the linearly polarised antenna vertical and the other with it horizontal (in fact any two orientations that are 90° apart will do, provided exactly the same orientations are used for the measurements involving antennas B & C that were used for A & C). The five transmission measurements are summarised in the Table 3 below:

| Measurement | Tx antenna | Rx Antenna | Gain Product | Polarisation |
|-------------|------------|------------|--------------|--------------|
| 1 | A | В | Gab | Vert. |
| 2 | A | С | Gacv | Vert. |
| 3 | A | С | Gach | Horiz. |
| 4 | В | С | Gbcv | Vert. |
| 5 | В | С | Gbch | Horiz. |

Table 3 Antenna combinations and polarisations when calibrating a CP antenna

The gain of antenna C is given by the sum of its two partial gain components:

$$Gc = \left(\frac{GacvGbcv}{Gab}\right)^{\frac{1}{2}} + \left(\frac{GachGbch}{Gab}\right)^{\frac{1}{2}}$$

The gain Gc so obtained is the polarisation matched gain for the antenna. Provided the polarisation of the antenna is reasonably close to circular, there is little difference between the polarisation matched gain and the gain of the antenna when receiving CP waves of the correct handedness.

The reflection coefficients, and therefore the match, of typical wideband EMC antennas tend to vary significantly across the operating band, sometimes exhibiting peaks as high as $\Gamma = 0.5$ (i.e., VSWR = 3.0) and more. There are three good reasons for measuring the complex reflection coefficients of microwave EMC antennas:

- 1) It is a good diagnostic to show whether the antenna is operating correctly.
- 2) It allows mismatch corrections to be applied.
- 3) It allows mismatch correction uncertainties to be evaluated.

Even though matching pads are used during the measurements, these are not perfect and the resulting mismatch can become significant for a very poorly matched antenna. For example, a good Type-N matching pad will have a reflection coefficient $\Gamma \leq 0.07$ units (return loss > 23 dB). If this is connected to an antenna with $\Gamma = 0.5$, a Z₀ mismatch loss of up to 0.32 dB could result. Pads with smaller coaxial connectors, such as SMA, 2.92 mm or 2.4 mm are likely to have worse reflection coefficients and hence larger errors can be expected. The match of these pads can be expected to deteriorate over time as they become worn through repeated connection and disconnection, which makes regular verification essential.

In addition to the mismatch, one must account for the loss of the adaptor used during the direct connection. In general, good quality Type-N adaptors can be obtained with precision connectors, whose characteristics meet those given in Table 4 over the frequency range 1 to 18 GHz.

| Parameter | Value | in dB | |
|--|----------------|--------------------------|--|
| $ \mathbf{S}_{11} $ or $ \mathbf{S}_{22} $ | < 0.050 units | Return Loss >26.0 dB | |
| S12 or S21 | > 0.9943 units | Insertion Loss < 0.05 dB | |

Table 4. Typical Type-N adaptor characteristics

The effect of such an adaptor can readily be treated as an uncertainty, since its effect will be small compared to other uncertainties. However, it must be stressed that unless the S-parameters of the adaptor are evaluated periodically, it is impossible to determine the magnitude of the uncertainty, and as adaptors wear through use, their characteristics can change markedly.

Engineers who are used to dealing with antennas, cables and connectors below 1 GHz should be aware that losses and mismatches increase rapidly with frequency above 1 GHz and this is especially true of smaller connectors and line sizes, such as 2.92 mm and 2.4 mm.

9.9.1. Comment on the ANSI and ARP958 calibration methods

ANSI C63.5 indicates that horn antennas may be calibrated from 1 GHz to 40 GHz for a 3 m separation over a ground plane provided the antennas are at least 2 m from the ground. It also indicates that due to the inherent directivity of horn antennas, the ground reflection will be negligible and therefore the free-space antenna factor will be obtained.

The document SAE ARP958 states that the measurements may be carried out inside an anechoic chamber, with no RAM covering on the floor. Contrary to ANSI C63.5, SAE ARP958 indicates that ground reflections will influence the results and a standard height of 3 m should be used. At NPL a fully anechoic chamber is used for these measurements, thus ensuring any reflections from the surroundings are reduced to insignificant levels, which results in the free-space antenna factor being measured.

SAE ARP958 also recommends using two identical antennas set 1 m apart, and using the usual Friis transmission formula:

$$P_{\rm R} = \frac{P_{\rm T}G_1}{4\pi {\rm r}^2} \times \frac{\lambda^2 G_2}{4\pi}$$

There are 5 main problems with this approach that must be kept in mind when considering the uncertainties associated with this method:

- 1) The Friis transmission formula assumes far-field conditions, but is being employed in a near-field situation.
- 2) The effect of multiple reflections between the antennas will depend on the antennas used and, therefore, so will the result.
- 3) Antennas manufactured to be identical are generally not. Differences of over 1 dB are common for the type of wideband antennas employed in EMC measurements.
- 4) The calibration laboratory may not have an antenna identical to the one submitted for calibration.
- 5) The calibration of the antenna is valid when the EUT (Equipment Under Test) is an "identical antenna" (as in the calibration conditions), but in practice the EUT will be the product being tested for its EMC compliance, which will have a different effect on the antenna performance. This is because of the close proximity of the antenna and the EUT.

Despite these difficulties, the method may still be used, and problems 3) and 4) can be avoided altogether by using a three antenna method, rather than assuming the antennas are identical.

9.10. Calibration of rod antennas

First, some background on why many models of rod antenna are 41 inches long. In the mid-1940s the military first encountered EMI in situations in which the receiver and transmitter were close to each other. In cargo airplanes the radio operator's position had a desk area that consisted of a flat metal plate large enough to hold a receiver and provide a writing surface and a surface for a telegraph key. The antenna lead-in to the receiver was about 41 inches long, coming from the top of the fuselage down to the receiver. Only about a foot away from the operator's desk was the transmitting equipment operating in the LF, MF and HF bands. The length of a rod antenna and its diameter simulated the lead-in wire in those radio operator's positions. It was used only 12 to 20 inches away from the device under test, during the EMI tests, to simulate the real-world situation.

Rod antennas are used for radiated emission measurements most commonly in the frequency range 10 kHz to 30 MHz. Most models have a telescopic monopole element, which can be set to a length of 41 inches because this is the usual length specified in the testing standards. The most widely used method of calibrating rod antennas is by a closed circuit (i.e. conducted) measurement, in which the element is substituted by a capacitor of about 12 pF. This equivalent capacitor substitution method (ECSM) is based on the fact that the capacitor simulates the low frequency impedance of the actual element, and there are expressions described in both the CISPR and ANSI standards, which are used to find the correct value of capacitance for any given dimension of element. Unfortunately, among the various literature sources there are two slightly different forms of the expression used to calculate the capacitance, one of which will give an incorrect value. The 'incorrect' expression will give a value near 10 pF for a 41 inch element, and if this were used for a calibration the potential error in antenna factor is up to 1 dB. The correct expression is given in Annex B of CISPR 16-1-4¹⁰.

NPL has developed a technique for the calibration of rod antennas up to 1.75 m in length, in *free-field* conditions, from 100 Hz to 100 MHz. The base of the rod antenna unit is placed in electrical contact with the 60 m x 30 m metal ground plane on the NPL site, and it is illuminated by a field from a transmitting monopole antenna approximately 20 m away. The antenna is substituted by a calculable monopole whose antenna factor is know to better than $\pm 0.3 \text{ dB}^{38}$. By this method it is possible to achieve uncertainties of less than $\pm 0.5 \text{ dB}$ for a well-behaved rod antenna of the same height as the calculable standard rod. Because the input impedance increases as the monopole gets electrically shorter, the transmitted field strength is very small below 5 MHz, so a calibration method has been developed which uses a large GTEM cell to illuminate monopoles by an approximately plane wave in the frequency range from 100 Hz to 30 MHz. In both configurations the antenna factor is determined by substitution with a standard passive rod antenna whose antenna factors are calculated by numerical simulation¹⁴. The free-field method described here achieves antenna factor uncertainties less than $\pm 1.5 \text{ dB}$ (*k*=2).

The GTEM calibration method has been compared to the ECSM in a recent NPL project³⁹. The main conclusion was that the two methods agree within measurement uncertainty, but

with some stipulated conditions. Assuming that the ECSM has been set up with the correct value of capacitance which, when mounted in an input adaptor, accurately simulates the monopole element, then the measured antenna factor will correspond to the case where the zero potential ground reference is at the base of the element (physically on top of the base unit).

This equivalent capacitance antenna factor will be higher than the value measured in the GTEM because in this case the ground reference is under the base unit, and the element feed point is raised up. Depending on the height of the base unit, this effect causes from 0.8 dB to 2 dB difference in antenna factor. Neither result is wrong, rather it is just a case of choosing which is appropriate for the testing configuration being considered. For example, many monopoles have a small ground plane which may be bonded to a conducting test bench, and this ground plane may either be attached to the top of the base unit or underneath it. The ECSM more closely matches the case with the ground plane on top, and the open-site and GTEM measurements match the other situation. Often, within the context of EMC measurements, these magnitudes of uncertainty are not considered to be significant, but, if required, approximate corrections can be derived to account for the height of any base unit, so that both calibration methods are equally valid.

Above 15 MHz, which is in the region approaching the resonant frequency of the monopole element, free-field techniques, such as the GTEM method, provide more reliable antenna factors simply because the ECSM assumes a purely reactive element impedance which is not the case near resonance. The ECSM result will be in error for 41 inch rods above 20 MHz and significantly in error above 30 MHz. However the field in the GTEM cell is not sufficiently uniform or calculable above 30 MHz, so for frequencies above 30 MHz the antenna range has to be used.

Occasionally monopole antennas are used on tripods, putting the antenna about 1 m above ground, and with a small ground plane attached to the base unit. The ground planes are about $0.6 \text{ m} \times 0.6 \text{ m}$ in size so they are obviously much too small to simulate an effective ground plane, which ought to be at least two wavelengths in extent for a reasonable image of the rod element to be formed. Every case like this requires a unique free-field calibration. If the coaxial cable carries any common mode current, its layout can affect the measured signal. The layout of the vertically hanging section of feed cable should be moved with respect to the rod element and if there is a significant change in reading, the cable should be wound through a ferrite torroid one or two turns.

Antenna factors derived by the above methods can be used to give the E-field strength in free-field conditions, and as argued in Appendix 2 will accurately measure the field vector at a near-field distance, such as 1 m. However most applications of rod antennas are in screened rooms, whose size dictates that there are no longer free-field conditions, in other words the room is operating below its cut-off frequency. In this case E-field magnitude as conventionally understood is not being measured and special steps are needed to derive E-field magnitude which can be related to free-field emissions. Significant progress has been made in quantifying this relationship and is discussed in Appendix 5.
9.10.1. The equivalent capacitor substitution method

In the ECSM the monopole element is replaced by a capacitor that simulates the reactive part of the monopole element at low frequencies, when the monopole is at 0 Volt potential with reference to an infinite ground plane. The real part of the impedance is tiny in comparison to the reactive part, so it is ignored for this substitution: in fact, for a 1 m element the real part is less than 0.05% of the reactive part below 10 MHz.

Since an infinite ground plane is impossible to realise, most EMC testing standards which use monopoles are designed to give an approximation to this ideal within practical limits. A typical OATS usually gives a good approximation to an infinite ground plane, as long as the edges are well grounded to earth; also there should be no buildings nearby which cause significant reflections. Many commercial monopoles are supplied with a small ground plate (usually $0.6 \text{ m} \times 0.6 \text{ m}$), or radial wires, which should ideally be referenced to the common ground of the test environment of the antenna. For this reason, standards for 1 m emission measurements in screened rooms, such as MIL-STD-461/462, require the ground plate of the monopole to be bonded to the conducting surface of the table on which EUTs are placed.

CISPR 16-1-4 gives the following expression for calculating the value of capacitance to use for the ECSM. Here h is the height of the monopole element, and 'a' is the average radius of the element.

$$Ca = \frac{55.6 \times h}{\left(Ln\left(\frac{h}{a}\right) - 1\right)} \cdot \left[\frac{Tan\left(\frac{2\pi h}{\lambda}\right)}{\frac{2\pi h}{\lambda}}\right]$$

When *h* is a small fraction of λ the second term, in square brackets, becomes unity and very often you will find just the first part of the expression is quoted. For a 1.04 m element (41 inches), with 3 mm radius, this expression gives $C_a = 11.9 \text{ pF}$, hence the commonly quoted figure of 12 pF.

The ECSM makes use of an adaptor box that contains the capacitor. The box has a suitable adaptor to connect the capacitor directly onto the input terminal of the monopole being calibrated, i.e. where the element would normally be attached. The box is supplied with RF signal via a BNC coaxial connector, and this signal then passes through the capacitor and directly into the input terminal. Fig. 26 shows a schematic of a typical configuration for the ECSM.



Figure 26. Schematic of ECSM calibration system.

The coaxial T-piece is used because the capacitor presents a high impedance to the source, and potentially this could cause large mismatch errors if the source were connected directly to the capacitor. With a T-piece the voltage, V_{D_i} can be measured with a 50 Ω load on the output of the monopole and then, when the output voltage, V_R , is measured, the 50 Ω load is connected to the output from the T-piece so that the source always sees the same impedance; this ensures that the input voltage to the capacitor is stable. Once these two voltages are measured, the antenna factor of the monopole is given by the following expression:

$$AF(dB(m^{-1})) = 20 \cdot Log_{10}\left(\frac{V_D}{V_R}\right) - 20 \cdot Log_{10}(h_E)$$

where h_E is the effective height of the element, which is given by the following standard expression:

$$h_E = \frac{\lambda}{2\pi} \cdot Tan\left(\frac{\pi h}{\lambda}\right)$$

It can be seen that the expression for effective height reduces to h/2 when h is a small fraction of λ , and therefore the effective height of a 1 m element is 0.5 m at low frequencies, which gives a logarithmic correction of + 6 dB [i.e., -20log₁₀(h_E)]. In some places the expression for antenna factor is quoted with just + 6 dB included for the effective height correction term, which can be misleading if the monopole being measured does not have a 1 m element.

9.11. Calibration of loop antennas

Loop antennas are widely used in applications up to the microwave bands (\approx 3 GHz). They are often used as electromagnetic (EM) field probes and can be designed to primarily respond to the magnetic field component of the electromagnetic wave. EMC standards ANSI C63-4 and CISPR16-1-4 specify the use of loop antennas for the measurement of magnetic field emission between 9 kHz to 30 MHz. MIL-STD-461E requires magnetic field measurements from 30 Hz to 100 kHz. Typically, loop antennas have low efficiency, and are usually deployed as receiving antennas for radio direction finding, low noise LW and MW receivers, or where it is required to measure the magnetic component of a radiated field.

9.11.1. Passive loop antennas

Small loops have a far-field pattern very similar to that of a small dipole (when oriented normal to the plane of the loop) and have a gain the same as a dipole (gain = 1.5). As receiving antennas, loops are essentially sensitive to magnetic fields normal to the plane of the loop. The magnetic field induces a circulating current proportional to the magnetic flux density averaged over the area contained within the loop. The open circuit voltage across the loop is given by

$$Voc = 2 \pi AN f B$$
Where
$$A = \text{Area of Loop (m}^2)$$

$$N = \text{Number of turns}$$

$$f = \text{Frequency (Hz)}$$

$$B = \text{Magnetic Flux Density (Tesla)}$$

In its simplified form the equivalent circuit of a loop antenna consists of an inductor in series with the load resistor. In practice there is also a small resistance due to the wire and a small parallel capacitance. When the load impedance is much greater than the reactance of the loop $(Z_L \gg \omega L)$, the voltage developed across the load is close to the open circuit voltage, *Voc* defined above. The loop output voltage is directly proportional to frequency for a constant magnetic flux density. As the frequency increases and the loop impedance becomes comparable with the load impedance, the voltage across the load reduces due to the increasing source impedance of the loop. Eventually when the loop impedance is much greater than the load impedance ($Z_L \ll \omega L$), the increase in the induced voltage is exactly offset by the increase in the loop impedance so the voltage delivered across the load remains constant with frequency.

Magnetic antenna factors can relate the magnetic field to the voltage delivered to the load. The magnetic field strength can be expressed in terms of field strength, H in A/m or magnetic flux density, B in Tesla, which depends on the medium, as in the expression:

$B = \mu_0 \mu_R H$

For air the relative permeability μ_R is 1 and $\mu_0 = 4\pi \ x \ 10^{-7}$.

Magnetic antenna factors (AF_H) are therefore in units of A/m per V commonly expressed as S/m [where S (conductance) = A/V], or $pT/\mu V$ (pico tesla per micro volt) because the unit Tesla is very large.

Sometimes loop antenna factors are expressed in terms of plane-wave-equivalent electric fields providing an estimate of the electric field if the antenna is in far-field conditions, usually taken to be a distance of greater than $\lambda/2\pi$ from the source.

Antenna factors are almost always expressed as logarithmic ratios (in decibels) referenced to an absolute reference field strength. This helps to make the wide numerical range of antenna factors more practical and easy to use. Magnetic antenna factors are then expressed as dB[S/m] (0 dB = 1 S/m) or $dB[pT/\mu V]$ (0 dB = 1 pT/ μV), depending upon whether one is relating the output voltage to the Magnetic Field Strength (H) or the Magnetic Flux Density (B). See the conversions between units below. Because antenna factors are in units of voltage and current we must multiply the logarithm by 20 to convert to the decibel ratio.

Conversions: $AF_H dB[pT/\mu V] = AF_H (dB[S/m]) + 1.984$ $AF_E dB[1/m] = AF_H (dB[S/m]) + 51.53$

where AF_E is the antenna factor for electric field strength.

Fig. 27 shows a typical graph of antenna factor against frequency. It is convenient to plot the frequency on a logarithmic axis to cover several decades of frequency; this also serves to linearise the part of the graph where the antenna output is proportional to frequency. Doubling the frequency doubles the voltage output and results in a 6 dB decrease in antenna factor. We can say that the antenna factor changes 6 dB per octave in this region. This is also equivalent to 20 dB per decade of frequency.



Figure 27. Graph of antenna factor against frequency for a loop antenna

9.11.2. Active loop antennas

Active loop antennas exploit the region where $Z_L <<\omega L$ and the antenna factor remains constant with frequency. By using amplifiers having very low input impedances, the effective load impedance seen by the loop can be kept lower than the loop reactance at the lowest frequency of operation. An amplifier can then be used to adjust the sensitivity to a suitable level for the application and present a suitable source match to the coaxial connecting cable. A schematic circuit is shown in Fig. 28.

Active antennas require a power supply which can be built into the receiver. The power supply must be accurate and stable to prevent variations in the gain of the amplifier. The amplifier must be low noise and have low distortion and low intermodulation products. Using this technique, antenna factors flat to within ± 1 dB can be achieved from 9 kHz to 30 MHz using a 60 cm diameter loop.

As the frequency increases the current in the loop becomes non-uniform and residual capacitances within the loop cause resonances and spurious responses to occur.



Figure 28. Schematic diagram of an active loop antenna

A loop antenna can respond to the electric field unless carefully designed to prevent it acting as a folded dipole, as illustrated below in Fig. 29:



Figure 29. Morphing of a dipole into a loop antenna.

If the loop antenna is being deployed to measure the magnetic component of the electromagnetic field then it is important to know that the loop is insensitive to the electric field. Typically the H-Field sensitivity of a well-designed loop antenna in plane wave impedance will be between 30 dB and 60 dB greater than that of the corresponding E-Field.

Dipole currents can be minimised by making the loop diameter less than 1/12 of the wavelength of the highest operating frequency; however, unless an effective balun is used at the feed point, the cables form part of the antenna and easily overshoot the diameter criterion. By orienting the loop antenna so that a diameter from the feed point is parallel to the electric field, the sensitivity to the electric field can be minimised irrespective of the diameter, provided a suitably balance connection is made to the feed point. A convenient and highly successful way of connecting to the feed point is to use a transmission line within the loop which can lead to a suitable coaxial connector and hence to a receiver.

Fig. 30 shows a typical configuration of the shielded loop antenna:

The load Z_L is presented across the gap in the loop regardless of the actual position of the load or the length of the transmission line. It is assumed that the wall of the loop is thick enough and sufficiently highly conducting so that if the gap were closed the currents within the loop shield would be negligible. If this is the case then the outer surface of the coaxial shield is the conducting ring of the antenna and the inner surface is one of the conductors of the coaxial line. The voltage induced by the electric field is identical in both halves of the outer loop causing negligible voltage across the gap. The magnetic field, however, causes circulating currents in the outer loop which presents a voltage across the gap. It is good practice to reduce induced currents in the outer braid of the coaxial cable using ferrite clamps as the currents can cause errors due to imperfect screening within receivers.





The calibration of loop antennas requires the generation of a calculable uniform magnetic field. As loops are used at relatively low frequencies, an open area site is not usually large enough for far-field operation.

9.11.3. Induction field method

One method given in IEEE Standard 291-1991⁴⁰ and referenced in ARP958³⁷ involves the induction field between two close coaxially aligned loops. An antenna separation of 1 m to 2 m is generally used. The method requires the measurement of the circulating current in the transmitting loop and this can be done with a thermocouple ammeter or a vacuum thermocouple, which is introduced into the loop circuit. No estimates as to the uncertainties of this method are given in the standard. The significant contributions will be from the measurement of the circulating current, measurement of separation distance, measurement of the diameters of both loop antennas and the estimate of the effective area of the transmitting loop. The precise alignment of the antennas may also be required, and also a calibrated

receiver to measure the output voltage from the receive antenna. Since the induction field decreases as the inverse of the distance cubed this places a sensitivity factor of 3 on all of the measured lengths, however the distance to the nearest reflecting object need be only two or three times the antenna separation distance. It is expected that uncertainties in antenna factor using this method would vary between 0.5 to 2 dB depending on the frequency and type of antenna.

9.11.4. TEM cell method

The TEM cell is described in Section 8.4. A known magnetic field is generated in a Crawford type cell, calculated from the power into the cell, the separation distance between the septum and the lower (or upper) plate of the cell and the characteristic impedance. The loop antenna is placed in the lower (or upper) part of the cell, with the plane of the loop in line with the axis between the cell input and output, as shown in Fig. 31. In general a loop antenna has much smaller mutual coupling effects to its surroundings than does an electric dipole antenna, which makes it possible to calibrated loops of diameters getting close to the cell plate separation. For example a standard loop of 0.6 m diameter can be calibrated to an uncertainty of better than \pm 0.5 dB in a cell with a plate separation of 0.915 m.



Figure 31. Equipment setup for calibration of loops in a TEM cell.

10. Uncertainties

Antennas are calibrated by various methods. In this section an example is taken to illustrate the method of assessing the antenna factor uncertainty. Low gain VHF antennas derive their traceability from measurements on an open-field site. In simple terms, the antennas are precisely positioned above the site ground plane in an electromagnetic field of known field strength. The response of that antenna to the field is measured as the voltage at the antenna output. The required parameter is the antenna factor, which is the ratio of the field strength to the output voltage. The antenna factor is dependent on the height and orientation of the antenna above the ground plane, which introduces mutual coupling effects, thereby altering the input impedance of the antenna. The two methods used to derive antenna factor are the three-antenna method and the standard antenna method and though most of the uncertainty treatment applies to both, there are some distinctions. Antennas are also calibrated in the nonideal field uniformity of fully anechoic rooms, but an advantage is that mutual coupling to images is insignificant.

The sources of uncertainty are receiver linearity and drift and level of signal above noise floor, source drift, antenna positioning, repeatability of cable connection to the standard and customer's antennas, effect of temperature on attenuation of outdoor cables, reflections from the ground plane and unwanted reflections from the site, masts, etc. Modern instrumentation has very accurate frequency settings and the error is insignificant for this application. Guidelines for the assessment of uncertainties are given in Refs. 4 and 5. To give an idea of what uncertainties are achievable and offered on a commercial basis, the uncertainty offered by NPL for the antenna factor of a dipole antennas is $\pm^{\circ}0.35$ dB, but the best achievable uncertainty is ± 0.15 dB. The best achievable uncertainty for horn antennas is ± 0.04 dB. The typical uncertainty often requested by customers for broadband antennas is ± 1 dB which allows a speedier and lower priced service. Broadband hybrid antennas are more complex and the standard uncertainty offered is ± 1.2 dB, though a lower uncertainty service of ± 0.7 dB is offered. Monopole antennas cover many octaves in frequency and, as generally for all antennas, the worst-case uncertainty across the whole frequency band is guoted in the certificate of calibration. It would be possible to divide up the frequency band and quote best uncertainties for each part.

An uncertainty budget for the calibration of a dipole antenna above a ground plane by the standard antenna method using the calculable dipole is given in Ref. 6. An example uncertainty budget follows in Tables 2 and 3 for the calibration of a biconical antenna on the open-field site by the standard antenna method.

10.1. Uncertainties budget for a biconical antenna calibrated on an OATS

Table 5 contains the common components which apply to all linear ratio type measurements done with the HP8753D ANA. This includes each site insertion loss measurement in the three-antenna method, and also the difference (in dB) which is calculated during the standard antenna method. In the tables Ci is the sensitivity coefficient, Ui is the standard uncertainty, Dist. is Probability Distribution and Div. is Divisor.

| 10010 | | | | | 01002 | |
|-------|--|---------------|----------|------------|-------|----------------|
| Note | Source of uncertainty | Value (dB) | Dist. | Div. | Ci | Ci x Ui |
| A | Source amplitude stability | 0.02 | Rect | √3 | 1 | 0.012 0.012 |
| В | Receiver noise, see Section 10.2 | 0.02 | Rect | √3 | 1 | 0.012 0.012 |
| C | Connector repeatability Freq ≤ 300 MHz | 0.1 (0.05) | Nor m | 1 | 1 | 0.1 (0.05) |
| D | Cable attenuation variation due to temperature | 0.2 | Rect | √3 | 1 | 0.115 |
| Е | Resolution of stored data (2 decimal places) | 0.005 | Rect | √3 | 1 | 0.003 0.003 |
| F | Ambient (Received power at least 20dB higher) | 0.04 | Rect | √3 | 1 | 0.023 0.023 |
| G | HP8753D linearity error | 0.14 | Rect | $\sqrt{3}$ | 1 | 0.081 |
| | | | | | | |
| | Combined Uncertainty, Uc | | | | | 0.177 |
| | For frequencies ≤ 300 MHz | | | | | (0.155) |

Table 5. Table of common components for ratio measurements on the HP8753D

Table 6 contains the uncertainty components for the calibration of biconical and broadband hybrid antennas by the NPL quasi-free space method described in Section 9.2.2, using the standard antenna method. The measurand AF is calculated as:

AF = Std(VP2m) - Cust(VP2m) + StdAF(FreeSpace)

| Note | Source of uncertainty | Value (dB) | Dist | Div | Ci | Ci x Ui |
|------|---|---------------|-----------|------------|----|------------------|
| А | Component from Table 5 (for frequencies ≤ 300 MHz) | 0.155 | Norm | 1 | 1 | 0.155 |
| В | AF of VHBB9124 standard | 0.35 | 95% cl | 2 | 1 | 0.175 |
| С | Reflections from mast & cables. This is relatively small because correlation in (Std - Cust) helps to reduce error. | 0.05 | Rect | √3 | 1 | 0.029 |
| D | Error in free-space assumption with the biconical antenna at VP2m | 0.15 | Rect | √3 | 1 | 0.087 |
| Е | Positioning error, 2cm/20m | 0.009 | Rect | $\sqrt{3}$ | 1 | 0.005 |
| F | Mismatch error | 0.12 | U | $\sqrt{2}$ | 1 | 0.085 |
| G | Additional component for Bilog calibration up to 250 MHz. This includes error due to difference in structure and misalignment. | {0.3} | Rect | √3 | 1 | {0.173} |
| | Combined uncertainty Uc 95% confidence limit | | | k=2 | | 0.265 0.5 |
| | Bilog combined uncertainty Uc 95% confidence limit | | | k=2 | | {0.317} {0.6} |

Table 6. Uncertainty for quasi free-space calibration of biconical antennas.

10.2. Uncertainty due to receiver noise

The presence of noise in a microwave measurement receiver results in an error in the measured magnitude of a *CW* (*continuous wave*) signal. If the receiver is a vector receiver, there will be a corresponding error in the measured phase. The error is small if the magnitude of the signal is well above the noise floor of the receiver but gets larger as the signal gets smaller. The size of this error is of interest in a wide range of RF and microwave measurement applications.

In order to investigate the error, a simple mathematical model is introduced⁴¹ in which the CW signal is represented by a constant phasor, the noise is represented by a normally

distributed random phasor and the indication of the receiver is given by the sum⁴² of the constant phasor and the random phasor. Monte Carlo Simulation can then be used to predict, from the model, how the error in magnitude varies with the signal to noise ratio (s/n). The s/n is the ratio of the magnitude of the signal to the time averaged magnitude of the receiver noise. The error due to noise predicted by the model is then compared with the error predicted by the coherent and incoherent addition of signal and noise and also with the error observed in practice for a spectrum analyser.

In order to measure the error due to noise as a function of the s/n for a spectrum analyser, the indicated magnitude is observed for several applied signals of known magnitude. The known signal levels are achieved using a signal generator and a calibrated switched attenuator. The amplitude of the signal generator is set in order that, with the calibrated attenuator switched to zero attenuation, the signal level is sufficiently far above the noise floor of the spectrum analyser to be accurately measured. Increasing the attenuation of the attenuator in known steps allows signals of several known amplitudes to be applied to the spectrum analyser. The error is obtained from the known true magnitude and the magnitude indicated by the spectrum analyser. The level of the noise floor is measured with no applied signal and is used to calculate the s/n. All the spectrum analyser readings are obtained as the average of a number of repeat measurements.

A straightforward and intuitive mathematical model of a CW microwave signal in the presence of noise has been constructed. Monte Carlo Simulation was used in conjunction with the model to predict the variation of the error in indicated magnitude with the s/n in a receiver. The error curve thus obtained shows good agreement with that obtained by the quadrature addition of signal and noise and also with the observed behaviour of a spectrum analyser. Table 7 shows some error magnitudes for selected s/n ratios. According to the model, when the noise level is equal in magnitude to the wanted signal (i.e. s/n 0 dB) the error is 2.7 dB, compared with an error of 6 dB in the case where both the wanted and unwanted signals are CW. For the error to be less than 0.1 dB the s/n should be at least 15 dB. The magnitude of error for other s/n ratios can be determined from Fig. 32.

| Signal to noise ratio | Error in measured | | |
|-----------------------|-----------------------|--|--|
| (dB) | signal magnitude (dB) | | |
| -6 | 6.8 | | |
| 0 | 2.7 | | |
| 3 | 1.4 | | |
| 6 | 0.71 | | |
| 10 | 0.28 | | |
| 15 | 0.089 | | |
| 20 | 0.028 | | |
| 40 | 0.00035 | | |

Table 7. Predicted error in the measured magnitude due to presence of noise in the receiver for different values of the s/n



Figure 32. Predicted error in magnitude due to noise as a function of the signal to noise ratio

10.3. Uncertainties in chambers

An example uncertainty budget is given in Table 8 for a typical Double Ridged Guide (DRG) horn antenna measured at 1 m.

| | | Value | Probability | | Ui |
|--------|------------------------------------|--------|------------------|---------|--------|
| Symbol | Source of Uncertainty | +/- dB | Distribution | Divisor | +/- dB |
| U1 | Neglecting loss of adaptor and RCg | 0.102 | U-Shaped | 1.41 | 0.072 |
| U2 | Multiple Reflections | 0.410 | U-Shaped | 1.41 | 0.290 |
| U3 | Mismatch error (assuming RCg=0) | 0.123 | U-Shaped | 1.41 | 0.087 |
| U4 | Polarisation Mismatch | 0.030 | Normal | 1 | 0.030 |
| U5 | Cable flexing | 0.170 | Normal | 1 | 0.170 |
| U6 | Receiver non-linearity | 0.123 | Rectangular | 1.73 | 0.071 |
| U7 | Receiver noise | 0.070 | Normal | 3 | 0.023 |
| U8 | Separation Measurement | 0.060 | Normal | 1 | 0.060 |
| U9 | RAM reflections | 0.085 | U-Shaped | 1.41 | 0.060 |
| U10 | Antenna alignment | 0.100 | Normal | 1 | 0.100 |
| Uc | Combined uncertainty | | normal | | 0.387 |
| U | Expanded uncertainty | | normal ($k=2$) | | 0.8 |

Table 8 Example uncertainty budget for DRG horn antenna measured at 1 m

The expanded uncertainty for this example is ± 0.8 dB and is based on a standard uncertainty multiplied by a coverage factor k = 2, providing a level of confidence of approximately 95%. The budget assumes a sensitivity factor of unity and infinite degrees of freedom for every component. Since treatment of the calibrated antenna after its return is beyond the control of

the calibration laboratory and its intrinsic stability is not normally assessed, a statement similar to the following is normally made in the calibration certificate: "The uncertainty applies only to the measured values and gives no indication of the long term stability of the antenna".

As can be seen from the table, the most significant source of uncertainty is due to multiple reflections between the antennas, which are separated by only 1 m. This uncertainty was estimated from the results of a series of measurements made between a pair of DRG horns as the separation between them was varied about 1 m, and also from measurements of octave bandwidth waveguide horns with adaptors to Type-N calibrated against similar antennas and against DRG horn antennas. Example data for the realised gain of a WG18 horn antenna with adaptor to Type-N calibrated against other similar WG18 horns and also against a pair of DRG horns is given in Fig. 33. The increased effect of multiple reflections between the antennas when the DRG horns were used for the calibration is very evident.



Figure 33. Gain values at 1 m measured against WG18 horns and DRG horns

In the example given above, partial mismatch correction was performed, to correct for the most significant terms in the mismatch formula only. The loss through the adaptor and the mismatch terms arising from connections to the Tx matching pad were treated as uncertainties, which were evaluated from periodically measured data.

Since the Tx and Rx matching pads must be connected via an adaptor at the start of the measurement and then one of the antenna carriages is moved back to achieve the required separation, the RF connecting cables are unavoidably flexed during the calibration. While it is possible to obtain coaxial cables with very good repeatability after flexing, all cables gradually deteriorate and it would be optimistic to use cable flexing data from a new set of cables when drawing up one's uncertainty budget. For this reason the value given above is consistent with a set of cables that are reaching the end of their useful life. It should also be

noted that cable-flexing effects increase with frequency and in general are more significant for horn calibrations (1 to 18 GHz) than they are for wire antenna measurements (below 1 GHz).

The effect of reflections from the RAM within the chamber was estimated from reflectivity measurements on the RAM, the measured directivity of the antennas, and the geometry of the measurement setup. The receiver non-linearity was evaluated using a calibrated step attenuator.



Figure 34. Deviation from the mean for 19 separate calibrations of a DRG horn.

An indication of the overall repeatability of the measurement is given in Fig. 34. This shows the deviation from the mean value for 19 separate calibrations of a double-ridged guide horn antenna over a period of a year. This is the same antenna as used for Fig. 33. The increased spread above 14 GHz corresponds to the large peak in the gain. It is in this region that the multiple reflections between the antennas are most pronounced.

10.4. Uncertainties budget for calibration of rod antennas in a GTEM cell

The estimated uncertainty for the calibration of a 41-inch monopole is shown in Table 9. Refer to Section 9.10 to learn about the different methods for the following frequency ranges:

| Range $A = 100 \text{ Hz}$ to 10kHz | Derived AF in GTEM |
|---|--------------------|
| Range $B = 10$ kHz to 30 MHz | SAM in GTEM |
| Range $C = 30$ MHz to 100 MHz | SAM on OATS |

| Source of uncertainty (dB) | Div | Range A | Range B | Range C |
|---|------------|----------|----------|----------|
| ANA linearity: 0.1 dB | $\sqrt{3}$ | 0.058 | 0.058 | 0.058 |
| Connector repeatability (BNC): 0.05 dB | 1 | 0.05 | 0.05 | 0.05 |
| Mismatch in SIL: 0.49 dB | $\sqrt{3}$ | 0.283 | | |
| $\Gamma_{ANA} = 25 \text{ dB}, \ \Gamma_{GTEM} = 20 \text{ dB}, \\ S11_{AUT} = 10 \text{ dB}$ | | | | |
| Error in H due to sloping septum: 0.231 dB | $\sqrt{3}$ | 0.134 | | |
| 0.04m change over AUT base area. Multiplied by 8.68/1.5m (coverage factor) | | | | |
| Error in derived correction C: 0.2 dB | $\sqrt{3}$ | 0.116 | | |
| Curvature of E-field: 0.3 dB | $\sqrt{3}$ | 0.173 | | |
| E-field calculation assumes a plane wave. | | | | |
| Mismatch in SAM: 0.50 dB (STD) | $\sqrt{2}$ | | 0.354 | 0.354 |
| 0.14 dB (AUT) | | | 0.099 | 0.099 |
| $\Gamma_{ANA} = 25 \text{ dB}, \text{ S11}_{STD} = 0 \text{ dB},$ S11 _{AUT} = 10 dB | | | | |
| SAM error due to different heights: | | | | |
| 0.3 dB (GTEM) | $\sqrt{3}$ | | 0.173 | |
| 0.1 dB (OATS) | | | | 0.058 |
| AF of passive standards: 0.2 dB | 1 | | 0.2 | 0.2 |
| Coupling with GTEM walls: 0.2 dB | $\sqrt{3}$ | Neg | 0.116 | |
| Only significant above 10 kHz | | | | |
| | | | | |
| Combined uncertainty | | 0.384 dB | 0.474 dB | 0.429 dB |
| Expanded uncertainty, <i>k</i> =2 | | 0.77 dB | 0.95 dB | 0.86 dB |

Table 9: Uncertainty budget for calibration of 41" rod antenna in GTEM cell and OATS.

10.5. Interpretation of uncertainties in NPL EMC antenna certificates

EMC testing is notorious for the large uncertainties associated with radiated emissions measurements. Instead of being a small percentage of the result, they can be a factor of ten. The main reason for this is that the device being tested is not reproducible. In large part this is caused by the variability of layout of the power and data cables connected to the device, which act like aerials, radiating any common mode current picked up from the electronics or chassis of the device. Because of the potential for large uncertainties the quality of test facilities specified in the standards is not very onerous, derived from the view that "what is the point of having very precise instrumentation when the reproducibility of an emission test is very poor?" This argument can be countered with the retort "Not all devices have cables

and some EMC tests can be performed precisely. Besides if one is to discover the source of uncertainties, one should pin down those whose accuracy can be improved, so that the real cause of the variability stands out from the other (reduced) uncertainty components. One is then in a better position to do something about it, like recommend a change in the test method."

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12. Glossary

The first use of terms in this document are italicised and included in this glossary.

- *3AM* see three-antenna method
- *accuracy* refers to the closeness of agreement between the result and the true value, whereas uncertainty refers to the range of values within which the true value is believed to lie. To quote accuracy the result must be known, whereas an uncertainty of ± 1 dB means the measurement could lie anywhere within this range.
- AF antenna factor is the ratio of the electrical field strength of an incident plane wave at a specified point of the antenna to the voltage induced across a specified the load (typically 50 ohm) connected to the antenna. Usually, the antenna factor is defined for the plane wave incident from the direction corresponding with the maximum gain of the antenna. The antenna factor has the physical dimension of inverse meters (m-1), and measured data are normally expressed in decibels (dB/m). Antenna Factor is proportional to frequency and inversely proportional to the square root of realised Antenna Gain.
- In radiated emission measurements, if an antenna factor of an antenna AF is known, the strength of an incident field, E, can be estimated from a reading V of a measuring receiver connected to the antenna by the formula E (dBmV/m) = V (dBmV) + AF (dB/m).
- AFfs see free-space antenna factor
- AUT antenna under test, i.e. the antenna being calibrated.
- *balun* is a passive electrical network for the transformation from a balanced to an unbalanced transmission line or device or vice versa
- *balun imbalance* when the balun itself if unbalanced, common mode current can be introduced to the outer conductor and this can radiate, interfering with the radiation of the antenna. This can be severe effect causing errors of greater than 5 dB in the measurement of field strength. A balance test is described in Clause 4.4.2 of CISPR $16-1-4^{10}$.
- bilog antenna "bilog" is a registered name, by Schaffner Ltd, who first manufactured (1994) an antenna to cover the frequency range 30 MHz to 1 GHz by combining a biconical and a LPDA antenna. A more generic name is "hybrid broadband antenna" or "hybrid biconical-log antenna".
- *boresight* the boresight direction of an antenna is mechanically defined and is generally the direction of maximum gain. It would be possible to calibrate antenna factor for other polar angles so that the effects of directivity could be compensated by calculation. However this would require measurement of amplitude and phase so that the direct and ground reflected signal could be correctly combined. It is not considered practicable to do this so only boresight antenna factor is measured.

- *broadband hybrid antenna* a combination of biconical antenna and LPDA antenna originally designed to cover the frequency range 30 MHz to 1 GHz. The biconical element is replaced by a bow tie in most models.
- *calculable antenna* antenna of which the electromagnetic characteristics, such as radiation pattern, antenna factor, and input impedance, can be calculated using either analytical or numerical techniques based on the dimensions and geometrical parameters. Effects of the balun are typically accounted for by s-parameter measurements of the balun network. If the balun structure is amenable to numerical modelling, a balun structure can be embedded into a numerical model.
- *cardioid* description of the shape of radiation pattern which is similar to that of an elemental dipole antenna, whose H-plane is uniform in magnitude and the power in the E-plane has a $\cos^2\theta$ shape. Popularly known as a doughnut shape.
- *CALTS* or CALibration Test Site, a term coined by CISPR in standard CISPR 16-1-5 for an OATS of higher quality that is suitable for calibrating antennas to be used for EMC testing. It has a tightly specified site attenuation performance in horizontal electric field polarization.
- *civil EMC* an example of a specification for civil EMC radiated emission measurements is European Standard EN 50081-1 "EMC - Generic emission standard Part 1: Residential, commercial and light industry".
- CISPR Comité International Spécial des Pertubations Radioélectriques, or International Special Committee on Radio Interference. This IEC sub-committee issues documents with which, in most part, European EMC Standards, or EuroNorms, conform. www.iec.ch
- *CW* (*continuous wave*) a CW signal in this context has a sinusoidal waveform.
- *DRG horn* Double Ridged waveGuide horn antenna
- *E-plane* the plane of the electric field radiated by an electric dipole antenna, the dipole lying in this plane. The E-plane pattern is the radiation pattern shape in the plane of the dipole.
- *ellipticity* an elliptically polarised field vector, is a field vector at a point in space, whose extremity describes an ellipse as a function of time. When the magnitude for one polarisation is the same as that for its orthogonal polarisation the ellipse is a circle, and the antenna that produces the field has perfect circular polarisisation.
- *EMC* ElectroMagnetic Compatibility. For further understanding refer to Ref. 2.
- *EMI* electro-magnetic interference. The term RFI, or radio frequency interference, is also used. But more generally one sees the term EMC which includes interference and immunity measures.

- *EUT* equipment under test, or an electrical/electronic product placed on the market that requires validation of EMC compliance so that the CE mark (for import into Europe) can be affixed.
- *FAR Fully Anechoic Room* has RF absorber covering all internal walls, including ceiling and floor. This is in contrast to a semi-anechoic chamber where the floor is deliberately left uncovered to simulate the ground plane of an open-field site
- *FCC* Federal Communications Commission, USA. Testing to show compliance with FCC rules is only allowable if the test site has been listed with the FCC.
- *free-field* is an unbounded field, in contrast to the field in a transmission line such as a waveguide or coaxial line.
- *free-space* is the condition where there are no reflections that return to the antenna and distort the reading of field strength. It is the condition in which the environment has no effect on the antenna, for example there is no effect of mutual coupling to anything in the environment.
- *free-space antenna factor (AFfs)* is the antenna factor of the antenna when it is removed so far from other conducting surfaces that mutual coupling of the antenna to these surfaces is negligible.
- *geometry specific AF* the antenna factor is measured in a way that is very similar to the way the antenna is being used in an EMC test, therefore including corrections associated with the site geometry, antenna pattern and antenna phase centre, and in extreme cases the near-field of the antenna. The site geometry comprises the heights of the antenna and EUT above the ground plane and the actual distance between them, compared to the separation intended in the standard which is the horizontal distance between their projections onto the ground plane (e.g. 3 m).
- *GTEM* Gigahertz TEM cell, distinguished from the Crawford TEM cell in that one end of the cell is terminated in a load, so that a resonant cavity is not formed, implying that higher frequencies can be used.
- *H-plane* the plane of the magnetic field of an electric dipole antenna, which is orthogonal to the E-plane. The H-plane pattern lies in the plane perpendicular to the plane in which the dipole antenna lies.
- *height max.* The height maximum is the height at which the maximum signal is recorded when the receive antenna is scanned in height from 1 m to 4 m, according to CISPR 16-1. This is to avoid destructive interference between direct and ground reflected rays from the source to the antenna.
- *ideal site* The ideal test site is a mathematical half-space, free of obstacles and bounded by a perfectly flat and perfectly conducting ground plane of infinite extent, acting as an ideal reflector of electromagnetic waves in the frequency range under consideration.
- *ISM* Industrial, Scientific and Medical equipment, distinct from residential/consumer products. A classification for IEC standards.

- LPDA Log-Periodic Dipole Array antenna A series of half-wave dipole elements tuned to increasing frequencies are spaced logarithmically along a boom to create a broadband antenna. The bandwidth is determined by the lengths of the shortest and longest elements.
- *NMI National Measurement Institute* or a National Standards Laboratory, for example the NPL in the UK.
- NSA normalised site attenuation is site attenuation (in decibels) minus the antenna factors (dBm⁻¹) of the transmit and receiving antennas.
- *OATS open area test site* is a flat open area site which has a large flat metallic ground. Potential scatterers such as buildings, trees, power lines and fences are sufficiently far away to cause insignificant change to the intended signal
- *opposing antenna* measurement of the gain or antenna factor of an antenna involves the use of another antenna, except in the rarely used Purcell's mirror method. The antenna factor is derived from the site insertion loss involving a pair of antennas. The antenna of interest is the antenna-under-test, or AUT, and the second antenna is often of a similar type, but at least has to give sufficient signal of the desired polarisation over the required frequency range. Often it is assumed that the AUT is the receiving antenna and that the opposing antenna is the transmitting, or source, antenna, but by the principal of reciprocity it makes no difference to the measurement outcome if the transmitter is connected to the AUT and the opposing antenna is connected to the receiver. Hence the term "opposing" rather than "source" has been used in this Guide.
- *phase centre* is the position on the antenna about which the antenna is rotated and results in the least change of electrical phase of the signal. See Appendix 4.2.
- *plane wave* a field with a wave impedance of 377Ω that is uniform in amplitude and phase across the aperture of the antenna.
- *resonant dipole* see *tuneable dipole*
- SACM Site attenuation comparison method
- *SAC semi-anechoic chamber* i.e. has a metal ground plane, but absorber on walls and ceiling, intended as an indoor version of an OATS.
- source antenna see opposing antenna
- SAM standard antenna method in which a the AUT is substituted by a standard antenna whose antenna factor is known. Unless the antennas are measured in a perfect plane wave in ideal free-space conditions, measurement uncertainties can be reduced if the standard antenna has similar dimensions to the AUT.
- *SA site attenuation* is defined as the minimum site insertion loss measured between two polarization-matched antennas located above a flat reflecting ground plane when one antenna is moved vertically over a specified height range.

- *SIL site insertion loss* is the loss between a pair of antennas on a test site, when a direct electrical connection between the generator output and receiver input is replaced by transmitting and receiving antennas placed at specified positions and with a specified polarisation.
- *SSM standard site method* this is the three-antenna method implemented above a ground plane in which height scanning is necessary to avoid signal nulls, and the maximum signal is found. It was defined by Smith³⁰.
- *reference test site* is a physical site that is as close as possible to the ideal site. It may be an especially large and flat ground plane at a National Measurement Institute or at a Calibration Laboratory. It is similar to or better than the CALTS. In CISPR jargon a reference site has been validated for both horizontal and vertical polarisation, in contrast to a CALTS which is defined for horizontal polarisation only. There is no physical reference for a FAR because the reference is the theoretical free-space attenuation.
- *three-antenna method (3AM)* is the principal method for calibrating antennas. The coupling between two antennas is measured in free-space conditions, with the antenna separation placing them in each other's far-field. This is repeated twice more with two other pairs of antennas selected from the three combinations of pairs of three antennas. The antenna factor of each antenna is calculated using three simultaneous equations including the three coupling measurements. The SSM is essentially the same, but with the antennas placed over a ground plane, and the antenna separation does not necessarily place them in the far-field.
- *tuneable dipole antenna* or tuned dipole antenna, is a thin wire antenna consisting of two straight co-linear conductors of equal length, placed end to end, separated by a small gap, whose length is a little less than half a wavelength and it is tuned for a resonant condition; at the specified frequency the input impedance of the wire antenna measured across the gap is pure real when the dipole is located in the free-space The resistance, or real part of the impedance, is approximately 72 Ω . Practical antennas commonly have telescopic elements for operation over the frequency range 30 MHz to 1000 MHz.
- *UHF* ultra high frequency, or 300 MHz to 3 GHz. VHF is very high frequency from 30 MHz to 300 MHz.
- UKAS is the United Kingdom Accreditation Service Address: UKAS, www.ukas.co.uk
- *uncertainty*⁴ see *accuracy*. An uncertainty of ± 1 dB in field strength or antenna factor is approximately equal to an uncertainty of $\pm 12.5\%$.

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Appendices

Appendix 1 Formulae

A1.1 Antenna factor of half wave dipole

From antenna theory, the following relations apply to a thin half wave dipole:

$$h = \frac{\lambda}{\pi}$$
 and $E = \frac{V_{OC}}{h}$ therefore $AF = \frac{\pi}{\lambda}$

where *h* is the effective length, VOC is the open circuit voltage at the antenna terminals and *E* is the incident field strength. VR is the voltage measured at the input of the receiver. The above formula gives *AF* of a very thin half wave dipole into a high impedance receiver. *AF* is conditional on the characteristic impedance, in this case a 50 Ω system, in which case *AF* is approximately given by (not including any balun losses):

$$AF = \frac{2.6\,\pi}{\lambda}$$

In decibels, an approximate expression for the antenna factor of a tuned dipole antenna is given by

$$AF = 20 \log f_M - 31.4$$

where f_M is the frequency in MHz. Approximately 0.5 dB can be added for the balun loss of a quarter wave coaxial balun. Note that a tuned dipole is slightly shorter than half a wavelength, because, for an antenna in a free-space environment, the reactance is tuned to zero. More shortening is required as the dipole radius increases.

A1.2 Antenna factor

$$AF = \frac{E}{V}$$
 in units of m⁻¹

In decibels $AF_{dB} = 20 \log_{10} AF$

where E is the electric field strength, in volts per meter, of a plane wave polarisation matched to the antenna, and V is the voltage measured at the input of the receiver, allowing for the attenuation of the cable connecting the antenna to the receiver. In decibels this is given as:

$$AF (dBm^{-1}) = E dBV/m - V dBV + A_C dB$$

where A_C is the attenuation of the cable in decibels and $V dBV = 20 \log_{10} V$.

A1.3 Conversion of realised gain to antenna factor

Antenna factor is directly related to gain. Note that antenna factor is proportional to frequency, so whereas the gain of an LPDA antenna may be fairly constant with rise in frequency, antenna factor increases linearly with frequency:

$$AF^{2} = \frac{4\pi Z_{F}}{G\lambda^{2} Z_{0}} \qquad \text{unit } \mathrm{m}^{-1}$$

where G is realised gain, λ is the wavelength in metres, Z_F is free-space impedance (~377 Ω) and Z_0 is the characteristic impedance of the antenna input transmission line. The significance of realised gain is that the mismatch loss of the antenna is included and does not have to be added to the antenna factor.

A1.4 Calculation of gain by the three-antenna method

The Friis formula relates the insertion loss between two antennas to the product of their gains:

$$P_R = P_T G_T G_R \left(\frac{\lambda}{4\pi R}\right)^2$$

where P_R and P_T are the powers received and transmitted, and G_R and G_T are the gains of the receive and transmit antennas. *R* is the separation distance between the antennas and λ is the wavelength, both in the same unit. This is repeated for the three combinations of pairs of three antennas and the gain solved by simultaneous equations.

A1.5 Calculation of antenna factor by the standard site method

Over a ground plane the ground ray has to be removed and this is a function of E_D^{max} (see Ref.16 by Smith for the definition). The far-field is assumed. Three insertion losses, *A*, are measured between 3 pair combinations of 3 antennas, and calculated using 3 simultaneous equations:

$$AF_{1} = 10\log_{10} f_{M} - 24.46 + \frac{1}{2} \left[E_{D}^{MAX} \left(dB\mu V / m \right) + A_{1} + A_{2} - A_{3} \right] \quad dB / m$$

where f_M is the frequency in MHz.

A1.6 Calculation of antenna factor by the standard antenna method

The antenna factor of the AUT is calculated using the following formula:

$$AF_{AUT} = AF_{STD} - (P_{AUT} - P_{STD})$$

where:

 AF_{AUT} = antenna factor (dB/m) of AUT at the measurement height.

 AF_{STD} = antenna factor (dB/m) of standard antenna at the measurement height for a given polarisation.

 P_{AUT} = receiver reading (dBm) with AUT in place.

 P_{STD} = receiver reading (dBm) with standard antenna in place.

A1.7 Site insertion loss

Site insertion loss (SIL) is defined as:

$$SIL = 10\log_{10}\frac{P_1}{P_2}dB$$

where P_1 is the power accepted by the receiver when the coaxial cables including the matching pads are connected directly, and P_2 is the power accepted when the signal is transmitted from one antenna to the other across a measurement site.

A1.8 E-field from an elemental dipole

The equation for the E-field component of an elemental, or Hertzian, dipole is reproduced here in order to show the relationship of E-field strength, E_{θ} , to distance, d, from the dipole:

$$E_{\theta} = \frac{j\eta\beta^2}{4\pi} M \left[\frac{1}{(\beta d)} - \frac{j}{(\beta d)^2} - \frac{1}{(\beta d)^3} \right] e^{-j\beta d} \cos\theta$$

where $\beta = 2\pi/\lambda$, η is the characteristic impedance of free-space, *M* is the dipole moment and θ is the angle from the centre of the transmit antenna, on the axis between the transmit and receive antennas, to the location of the receiving point (producing a cardioid radiation pattern).

The first term, the radiated field, shows that the E field is inversely proportional to distance. The second term shows that the field falls off as the square of the increasing distance; this is known as the near-field, or induction field, and only really becomes significant for an elemental dipole at distances of less than half a wavelength. The field in the third term falls off as the cube of the distance and is sometimes known as the electrostatic term; the measurement point has to be extremely close to the source for this term to become relevant, and this term does not normally enter EMC considerations. In item A1.9 the expression for D

in terms of the first two terms can be used to accurately calculate the field strength in the near-field of an short dipole. In the context of EMC uncertainties a short dipole can be considered to be less than a quarter of a wavelength. This calculation cannot be used to predict the E-field strength from a source that comprises a magnetic dipole, as mentioned in Section 9.6.2.

The ratio of the apertures of a half-wave and a Hertzian dipole is only 1.64 to 1.5. This means that an infinitely small dipole draws energy from nearly the same area like a half-wave dipole and may thus be sensitive to field inhomogeneities in an area exceeding its physical dimensions.

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A1.9 Theoretical value of normalised site attenuation in free-space

NSA =
$$20 \log_{10} \left(\frac{5Z_0 D}{2\pi} \right) - 20 \log_{10} f_M$$
 dB

where
$$D = \frac{d}{\sqrt{1 - \frac{1}{(\beta d)^2} + \frac{1}{(\beta d)^4}}}$$

Wavelength λ and separation distance *d* are in metres. f_M is the frequency in MHz and $\beta = \frac{2\pi}{\lambda}$, $Z_0 = 50\Omega$. For large distances the denominator reduces to unity and D = d. This expression for *D* is valid for antennas or field sources whose largest dimension is less than about a quarter of a wavelength.

A1.10 Derivation of H-field from a measurement of E-field

In plane wave conditions, i.e. in the far-field of the source, the H-field strength is directly related to E-field strength. Assuming that the medium is air, the H-field, in amps per metre, is given by:

H = E/377 or in decibels $H dB(Am^{-1}) = E dB(Vm^{-1}) - 51.5$

Appendix 2 Measurement of field-strength in the near-field of an EUT

A2.1 Measurement of absolute field strength

When assessing the uncertainty of measurement in the near-field it is important to know what the aim of the measurement is. Is it to measure the output voltage of the antenna as a means of comparing the performance of different EUTs, or is it to measure absolute field strength radiated by a specific EUT? In the latter case the method of calibration of the antenna contributes to the uncertainty of absolute field strength and, if the measurement is made in the near-field of the EUT, there will be an additional uncertainty in the extrapolation of the field to greater distances.

EMC dipole-like antennas have dimensions that are not much larger than half a wavelength, λ . Near-field effects can dominate the reading for distances, r, less than $\lambda/2\pi$ from a radiating source and for uncertainty budget purposes they should be quantified for distances of less than $\lambda/2$ from the source. In simple terms the near-field effects can be described as: (1) field strength variation by a $1/r^2$ term in addition to the 1/r distance term, (2) an amplitude and phase taper of the emitted field throughout the volume of the antenna and (3) the effect of the proximity of the EUT on the current distribution on the antenna elements. The latter is a mutual coupling effect. There is also the possibility that the proximity of the antenna has an effect on the field emitted by the EUT.

If the field were to be extrapolated from a distance of 3 m (from the EUT) to the far-field there would be an error of 1 dB at 30 MHz, dropping to 0.2 dB at 75 MHz.

There are two problems with the ANSI method of calibration, described in Section 4.2.1, which give rise to an uncertainty of the order of ± 0.5 dB in the field strength of a small source at 3 m distance, namely the ANSI formula does not include near-field terms, and in the derivation of AF one is trying to solve for 4 unknowns from 3 measurements. This error was found by modelling, comparing the E-field at a point in space to the field registered by the biconical antenna. Studies at NPL have shown that in the case of the biconical antenna a free-space calibration gives a similar order of uncertainty. A free-space calibration is preferred because only one set of AFs is needed for all distances greater than 3 m. The use of free-space antenna factors is taken as the default value in CISPR 16-1-4.

Note: the fact that the ANSI formula does not include near-field terms is not a problem when the antennas are used for NSA measurements, because the NSA formula also does not include near-field terms so the omission cancels out. It is only a problem for antenna factor, which is meant to give a reading of absolute field strength.

A2.2 Calibration with 1 m separation according to SAE ARP958

In Section 9.6 and Appendix 5 the background is given to the choice of a 1 m separation, between the antenna and the EUT in order to ensure better reproducibility of the measured field strength in an adverse measurement environment. At best the measurement was a

prescription to compare the emission of different EUTs in a repeatable manner so that they could be passed or failed against a given specification limit. This was seen as a more important criterion than achieving an accurate measure of absolute field strength. However it resulted in the adoption of the ARP958³⁷ method for measuring the "1 m gain" of an antenna, whose equivalent for the purposes of this document is labelled "1 m AF_{ARP}". This procedure is called up in standards such as MIL-STD-461⁴⁴.

In 1968 the International Society of Automotive Engineers produced document SAE ARP958 that gave a method for calibrating conical log-spiral antennas operating above 200 MHz for testing EUTs at a distance of 1 m. The method did not take the variation of phase centre with frequency into account, and could result in an error in the measurement of absolute field strength of up to 1 dB for an antenna 0.8 m in length at 200 MHz. The error decreased to zero at 1 GHz corresponding to the reference plane at the tip of the antenna. This assumed the measurement was made in free-space conditions, so the error could increase if there was a ground reflection. The document states that the calibration is of "1 m AF_{ARP} " which is not intended to give absolute field strength. In 1992 the document was revised to include the calibration of biconical antennas from 20 MHz upwards.

The problem with this procedure is that it uses the same calibration method and equations as for a calibration in the far-field, which clearly results in errors at 30 MHz for which 1 m is a tenth of a wavelength (0.1 λ). The formula for deriving gain takes no account of the fact that at 20 MHz an antenna separation of 1 m is 0.07 λ and is therefore truly in the near-field. "1 m AF_{ARP}" is approximately 2 dB higher than the AFfs above 50 MHz, 2 dB lower at 30 MHz and 5 dB lower at 20 MHz which implies that the uncertainty in the measurement of absolute field strength could be 5 dB. Also the antenna under test will have built into its antenna factors the effects of coupling to the source antenna. When the source antenna is replaced by an EUT the coupling will be different and the antenna factor will not be wholly valid.

This is an example of a calibration procedure that results in antenna factors that are intended to be used for only one type of setup, i.e. when making measurements with the tip of the antenna placed 1 m from the EUT. If the "1 m AF_{ARP} " of a 0.8 m long spiral antenna were mistakenly used to measure an EUT at a distance of 10 m (in free-space) the field strength could be in error by 4 dB. If an antenna has to be used at different distances it would be sensible to use just the free-space antenna factors and correct the distance by the phase centre value at each frequency. The other point to note about free-space antenna factor is that it is a unique property of the antenna, giving the E-field strength of a plane wave, and not dependent on its proximity to other conducting bodies.

NPL research on the reduction of uncertainties of 1 m measurements in screened rooms showed that uncertainties of typically \pm 30 dB can be reduced to the target of \pm 6 dB for frequencies above 30 MHz by partially lining the room with RF absorbing material. The EUT size was limited to 1 m³ and the distance to the antenna increased to 2 m. Under these conditions the correct antenna factor to use was AFfs. Much of the outcome of this research has been adopted in UK Defence Standard 59-41, in particular Part 5 Section 5. There is therefore less incentive to research the uncertainties in measuring absolute field strength using a biconical antenna 1 m away from an EUT.

Free-space antenna factor is the preferred factor for converting the output voltage of an antenna into the absolute field strength of the field in which the antenna is immersed. AFfs is a well-defined property of the antenna, independent of its surroundings. When the antenna is

used to make a measurement, AFfs is the best starting point to quantify the field, for which the uncertainties of deviations from the free-space assumption can be assessed. There is little advantage in uncertainty terms in using AF_{3m} compared with using AFfs at 3 m.

Appendix 3 Information on uncertainties in using biconical antennas.

This information can accompany a certificate of calibration.

A3.1 Antenna Factor

Where the antenna factor has been given for a specific configuration above a ground plane (including free-space), the associated uncertainties only apply when the antenna support structure, including the input cable, does not cause significant reflections which would affect the received signal. If there are any significant sources of reflection the user should assess the resulting uncertainty and treat it as an additional uncertainty term. For calibration purposes the free-space condition is achieved by mounting the antenna vertically polarised at a height above the ground plane at which mutual coupling is negligible.

Where there is a sharp resonance in the antenna factor the uncertainty given in the certificate does not apply. At the frequency where the resonance causes a deviation of greater than 1 dB from the overall trend of the data, the magnitude of the increased uncertainty can be estimated from the height of the spike on the antenna factor graph. The affected range can be taken as ± 1.5 % of the centre frequency. Because the data is sampled at discrete points the maximum error may be much larger than that shown in the antenna factor graph.

The antenna factors are valid at the measurement height for any separation distance from the source exceeding one wavelength. For shorter distances the change in antenna factor with distance becomes significant and additional uncertainty would therefore be introduced. When the antenna is used for emission testing at a distance of 3 m from an equipment under test, whose size does not exceed that of the biconical antenna, there is an estimated increase in uncertainty of ± 0.3 dB in the range 55 MHz to 100 MHz, which is caused by mutual coupling of the antenna to the EUT. Below 100 MHz the antenna is in the near-field of the EUT and though the field magnitude will be correctly measured there will be additional uncertainty if the field strength were extrapolated to a greater distance. For extrapolation to a distance of 10 m, which is effectively in the far-field, this uncertainty is estimated to be ± 0.2 dB at 100 MHz increasing to ± 1 dB at 30 MHz.

In order to measure the absolute E-field at different heights and polarisations above the ground plane it is necessary to know the antenna factor at each height and polarisation. However, since many calibrations would be required, a viable alternative is to use a single calibration of free-space antenna factor, AFfs. Because AFfs is approximately the average of AF at all heights in the range 1 - 4 m, the uncertainty in using AFfs is less than using AF measured at a fixed height. For example, biconical antennas are often calibrated horizontally polarised at a height of 2 m above a ground plane. The additional uncertainty is caused by coupling of the antenna with its image in the ground plane which results in a change in the input impedance. For vertical polarisation there is no additional uncertainty for heights above 1.5 m, but between 1 m and 1.5 m the additional uncertainty is ± 0.7 dB in the range 55 MHz to 100 MHz. For horizontal polarisation, at heights above 1 m, the antenna factor may differ from the quoted values by up to ± 0.5 dB in the range 20 MHz to 300 MHz. The values for horizontal polarisation can be reduced by 0.5 dB for antenna heights above 2 m. The

above variations are representative; the exact variation will vary slightly according to each antenna design.

If the antenna is used in an unlined screened room the use of these antenna factors may not give the absolute value of field strengths, but a calibration provides an essential check that the antenna is working properly. The antenna factors can be used to compare measurements made in an identical setup using a different antenna of the same type.

During height scans, with the antenna vertically polarised, there will be an additional uncertainty caused by the directivity of the vertical radiation pattern. In normal use, signal maxima on a 10 m range occur for antenna heights below 2.5 m and the error here will be negligible. However, for a 3 m range the received signal could decrease by more than 1 dB.

A3.2 Balance Test

The balance of the antenna balun may be tested by mounting the vertically polarised antenna in a uniform vertically polarised electric field, and observing the difference in received signal when the antenna is inverted. Any change greater than 0.5 dB is caused by common mode current on the cable which is caused by an unbalance of the balun. It is important for this test that the cable hangs vertically behind the antenna in the usual manner. For this test there should be a horizontal distance of between 0.5 m and 2 m from the antenna element to the point at which the cable drops vertically. The cable should not move during the course of the measurements. An antenna is considered to have a good balun balance when the observed difference is less than \pm 0.5 dB.

The inversion test is a qualitative measurement which reveals imbalance of the balun which, for some models of biconical antenna, can cause a large uncertainty in the measured field when the output cable is aligned parallel to the antenna elements. It is recommended that the user conducts tests of their own to quantify this effect in each particular measurement configuration. For antenna models with significant balun imbalance it is recommended that ferrite clamps are placed on the cable near the antenna input when the antenna is used for emission testing. Ferrite clamps on the output cable only provide a partial reduction of the braid current; a better solution is to use a perfectly balanced balun. The uncertainty of Antenna Factors is increased by the magnitude of balun imbalance.

A3.3 Return Loss

The quoted antenna factors apply when the mismatch between the antenna and the receiver is attenuated. A well-matched 10 dB attenuator is recommended. If no attenuator is used (and the receiver front-end attenuation is set to zero), the antenna factor can change by \pm 1.4 dB at 30 MHz, assuming a receiver return loss of greater than 14 dB, an antenna return loss of 1 dB and a cable loss of 1 dB.

A3.4 ARP958 Antenna Factor

Measurements at 1 m distance from an emitter is called for in MIL-STD-461D⁴⁴, which stipulates that procedure $ARP958^{37}$ is to be used for 1 m calibrations. It is necessary to distinguish between AF_{1m} and conventional AF which enables absolute E-field strength to be obtained from the voltage output of the antenna. ARP958 describes AF_{1m} as "apparent"

antenna factor because it is derived from equations which do not take near-field terms into account. When AF_{1m} is used to measure absolute field strength an additional uncertainty term of $\pm 2 \text{ dB}$ must be included. This only applies to the frequencies above 30 MHz, The uncertainty increases to $\pm 5 \text{ dB}$ as the frequency is reduced from 30 MHz to 20 MHz.

A3.5 ANSI Height Scan Method

The ANSI C63.5²² procedure describes how the antenna factor may be measured over a ground plane by a height scanning three-antenna method. For each measurement pair, one antenna is at a fixed height and polarisation, and the other is height scanned. The receiver is set to record the maximum measured signal during the scan. In the three pairings each antenna is measured twice, and if the customer supplies two antennas then one of the antennas is always allocated to the height scanning mount, and the other to the fixed mount. An NPL antenna is used for the third antenna which height scans for one pair and is fixed for the other pair. If the customer supplies one antenna it will be placed at the fixed height.

Where standards call for an ANSI calibration (e.g. for NSA measurements), NPL recommends the use of free-space antenna factors for the scanned antenna for measurements at 10 m separation because they agree well with 10 m ANSI antenna factors. However, at 3 m separation the ANSI antenna factors differ significantly from the free-space values, and therefore only the ANSI antenna factors should be used in order to comply fully with the NSA method described in ANSI C63.4:1992 and CISPR 16-1:1998.

Appendix 4 information on uncertainties in using LPDA antennas.

This information can accompany a certificate of calibration.

A4.1 Antenna Factor

The antenna factors are valid for any separation distance from the source exceeding one wavelength. For shorter distances the change in antenna factor with distance becomes significant and additional uncertainty would therefore be introduced.

Where there is a sharp resonance in the antenna factor the uncertainty given in the certificate does not apply. At the frequency where the resonance causes a deviation of greater than 1 dB from the overall trend of the data, the magnitude of the increased uncertainty can be estimated from the height of the spike on the antenna factor graph. The affected range can be taken as ± 1.5 % of the centre frequency. Because the data is sampled at discrete points the maximum error may be much larger than that shown in the antenna factor graph.

If the antenna is used horizontally polarised during a height scan from 1 m to 4 m above a ground plane, the antenna factors may differ from the values quoted by up to ± 0.5 dB. This is because the input impedance of the antenna changes due to coupling with its image in the ground plane. This coupling is greatest at the lower frequencies where the wavelength is a larger fraction of the height above the ground plane. When the antenna is used vertically polarised, there is no significant coupling with the ground plane, but the cable should extend horizontally behind the antenna for at least 2 m before dropping to ground in order to minimise parasitic reflections, particularly at the lowest frequency of operation of the antenna.

If the antenna is used in an unlined screened room the use of these antenna factors may not give the absolute value of field strengths, but a calibration provides an essential check that the antenna is working properly. The antenna factors can be used to compare measurements made in an identical setup using a different antenna of the same type.

There is a further error arising from the directive nature of the antenna radiation, which is greater at the higher frequencies. In a normal height scan up to 4 m, on a 10 m range, the signal maximum can be reduced by up to 0.5 dB compared with that for a uniform radiation pattern. For a 3 m range this error could be up to 2 dB (given that the signal maximum is normally achieved at a height of less than 2.5 m).

A4.2 Phase Centre

When a LPDA is receiving E-field radiation the phase centre is the active part of the antenna at any given frequency. The active part of the antenna corresponds approximately to the position of the element whose length is equal to that of the equivalent resonant half wave dipole for the received frequency.
The quoted uncertainty in antenna factor is only valid when the phase centre is placed at the point at which the field is required to be measured. If the antenna position is not adjusted with frequency to make this condition true, a correction should be made to the measured field (at the phase centre position). This is valid in free-space conditions but there is additional uncertainty when applied to a LPDA above a ground plane. For distances of greater than one wavelength from the antenna a reduction of the field proportional to the inverse of the distance can be assumed, which means that in an anechoic environment a linear extrapolation may be used to adjust the field strength. The adjustment of antenna factor to a fixed reference point on the antenna is described later in the annex. For measurements made over a ground plane this correction has to be calculated using the difference in E_{Dmax} (see Ref. 31).

The NPL certificate contains an expression that allows the phase centre at any frequency to be calculated. This approximation is derived from some equations that govern LPDA antennas with triangular profiles (i.e. where the element tips form a straight line). Hence larger errors in the predicted phase centre will occur when these expressions are used for tapered antennas. The values for the constants, which are given in the NPL certificate are derived from the following equations:-

$$\delta = \frac{X_L \cdot L_H - X_H \cdot L_L}{L_L - L_H} \qquad Tan\alpha = \frac{L_L}{2*(X_L + \delta)}$$
$$X_F = \frac{71.2}{Tan\alpha} \cdot \frac{1}{F_{_{MHz}}} - \delta$$

Where:

 L_L and $L_H =$ The lengths of two well spaced elements which are designed to be resonant at the Low and High frequency ends of the LPDA respectively. X_L and $X_H =$ The distance from the tip to the same two elements.

If the use of the above corrections is not feasible, an alternative strategy is available. This method, which may be applied in an anechoic chamber or near signal maxima during a height scan, uses a fixed phase centre, whose position is chosen in order to weight the incurred error evenly at either end of the operating frequency band. The fixed phase centre, X_{FIX} , is given by :-

$$X_{FIX} = \frac{1}{2} \cdot \left[X_{LOW} + X_{HIGH} \right]$$

The error incurred, U_E , at either end of the operating band is given by :-

$$U_{E} = \pm 20 * Log_{10} \left\{ \frac{R - \left[\frac{X_{LOW} - X_{HIGH}}{2} \right]}{R} \right\}$$

Where :- X_{LOW} = The phase centre of the low frequency operating limit. X_{HIGH} = The phase centre of the high frequency operating limit. R = The required separation to the EUT (i.e. 10 m or 3 m).

A4.2.1 Assessment of uncertainty arising from phase centre position on LPDAs

The antenna factor for an LPDA should be given at the phase centre position at each frequency. The uncertainty for AF in the Calibration Certificate only applies when the phase centre is positioned at the required separation distance, R, from the source (eg 10 m) at each frequency. If the antenna position is not adjusted with frequency, a correction should be made to the recorded field strength (at the phase centre position) to give its value at distance R from the source. For distances of greater than one wavelength from the antenna a fall-off of the field proportional to the inverse of the distance can be assumed. The correction factor, C_R dB, is added to the field strength, equation 1.

 $C_R = 20 \log [(R + P - d)/R]$

E-field strength is given by equation 2:

 $E_f = V_f + AF_{FS(f)} + CR$

where:

$$\begin{split} &f=\text{frequency, (MHz).} \\ &R=\text{the required separation from the source to the reference point, (m).} \\ &P=\text{phase centre position as a function of frequency, (m from tip).} \\ &d=\text{distance of the reference point from the antenna tip, (m).} \\ &E_f=\text{E-field at distance R from source (dBV/m).} \\ &V_f=\text{voltage at output of antenna at frequency f (dBV).} \\ &AF_{FS(f)}=\text{antenna factor (free space) for E-field at the phase centre, (dB/m).} \end{split}$$



Figure A4.1 The distance, R, from a source to the reference point on an LPDA

Table A4.1Example LPDA antenna factors. Correction assumes reference is 0.3 m fromtip

| Frequency | Antenna Factor at | phase centre, | Correction factor for |
|-----------|--------------------|----------------------|-----------------------|
| MHz | phase centre, dB/m | distance from tip, m | 3 m distance, dB |
| 200 | 11.1 | 0.6 | 0.83 |
| 400 | 15.2 | 0.3 | 0 |
| 1000 | 24.0 | 0.1 | -0.6 |

2

1

For measurements made over a ground plane this correction has to be calculated using the difference in E_{Dmax} (see Ref. 31) for the distances to the phase and mechanical centres of the antenna. If this procedure is not adopted a fixed phase centre can be chosen as the centre of the antenna. This will weight the error evenly at the ends of the operating frequency band. The uncertainty calculated in this way applies to fully anechoic chambers or above a ground plane at heights giving signal maxima. The uncertainty is greater at other heights, especially near signal minima. The fixed phase centre may be chosen as the point half way between the positions on the antenna which correspond to the phase centres for the two ends of the antenna's operating frequency band. For example for an LPDA of length 0.6 m and an operating frequency range of 200 MHz to 1 GHz, this method gives a fixed phase centre for this antenna at 0.35 m from the antenna's tip, and the uncertainty in the antenna factor at the ends of the band is \pm 0.2 dB for a separation of R = 10 m from the source. For R = 3 m the uncertainty increases to \pm 0.8 dB, as shown in Table A4.1.

A4.3 Return Loss

The antenna factors quoted apply when the mismatch between the antenna and the receiver is attenuated. A well matched 6 dB attenuator is recommended. For example, if no attenuator is used and the receiver front-end attenuation is set to zero, the antenna factor can change by typically ± 0.3 dB, assuming a receiver return loss of greater than 14 dB, an antenna return loss of 10 dB and a cable loss of 3 dB.

A4.4 Adjusted Antenna Factor

For LPDA antennas it is possible to calculate (A4.4.1), the result of an ARP958 1 m measurement rather than actually perform the measurement. We can do this because the LPDAs are in the far-field of each other. This calculated value does not take account of the small amount of coupling between the antennas which would occur during an actual ARP958 measurement, but this effect is of the order of 0.1 dB and is included in the stated uncertainty.

We can also calculate (A4.4.2), an adjustment to the antenna factor, which extrapolates the field measured at the phase centre of the antenna to a defined reference point. The separation to the EUT has to be specified and the reference point on the LPDA is sometimes specified as the tip. This type of adjustment is not the same as the first type, because the ARP method includes the unwanted phase centre spread of the opposing antenna as explained in Section 9.6.5, but roughly similar results are generated (using a 1 m EUT to LPDA separation, and setting the reference point at the tip). However, for large LPDA antennas the difference between the two adjustments can be in the region of 1 dB.

A4.4.1 ARP958 calculation

$$AF_{1m} = AF_{FS} + 10*Log_{10}\left[\frac{R + (2*X_F)}{R}\right]$$

A4.4.2 Reference point adjustment

$$AF_{REF} = AF_{FS} + 20*Log_{10}\left[\frac{R + X_F - X_{REF}}{R}\right]$$

Where :-

 AF_{lm} = ARP958 antenna factor, usually with R = 1 m.

 AF_{FS} = Measured free space antenna factor.

 AF_{REF} = Antenna factor referenced to defined point on LPDA.

- R = Separation either from tip to tip (A) or from EUT to reference point on LPDA (B).
- X_F = Position of phase centre from LPDA tip.

 X_{REF} = Position of defined reference point from LPDA tip.

Note

For R = 1 m and $X_{REF} = 0$ m, both expressions give similar answers for small values of X_F .



Adjustment of AF to give field strength at reference point of LPDA

Figure A4.4 Graph to demonstrate the error that could result if an LPDA calibrated for AFfs were used with its fixed reference point at a distance of 3 m from the EUT.

A4.5 Use of ARP958 Antenna Factor

Measurement at 1 m distance from an emitter is called for in MIL-STD-461D⁴⁴, which stipulates that procedure ARP958³⁷ is to be used for 1 m calibrations. It is necessary to distinguish between AF_{1m} and conventional AF which enables absolute E-field strength to be obtained from the voltage output of the antenna. ARP958 describes AF_{1m} as "apparent" antenna factor because it is derived from equations which do not take phase centre into account. When AF_{1m} is used to measure absolute field strength (at position of the active element at the frequency of measurement, i.e. the phase centre) an additional uncertainty term of ± 4 dB must be included at 200 MHz, and this diminishes to ± 0.5 dB at 1 GHz. This is because AF_{1m} extrapolates the field strength from the position it is measured by the active

element, to a distance of 1 m from the emitter. The extrapolation assumes a fall off in field inversely proportional to distance and does not take into account an imperfect measurement environment, such as a partially lined screened room, in which the field may not fall off linearly with distance.

A4.6 ANSI Height Scan Method

The ANSI C63.5²² procedure describes how the antenna factor may be measured over a ground plane by a height scanning three antenna method. For each measurement pair, one antenna is at a fixed height and polarisation, and the other is height scanned. The receiver is set to record the maximum measured signal during the scan. In the three pairings each antenna is measured twice, and if the customer supplies two antennas then one of the antennas is always allocated to the height scanning mount, and the other to the fixed mount. An NPL antenna is used for the third antenna which height scans for one pair and is fixed for the other pair. If the customer supplies one antenna it will be placed at the fixed height.

Where standards call for an ANSI calibration (e.g. for NSA measurements), NPL recommends the use of free-space antenna factors for measurements at 10 m separation because they agree well with 10 m ANSI antenna factors. However, at 3 m separation the ANSI antenna factors differ significantly from the free-space values, so for an emission measurement made by height scanning the uncertainties may be less using the ANSI 3 m AFs. Where the NSA of a 3 m site is being measured by the NSA method described in ANSI C63.4:1992 and CISPR 16-1-4¹⁰, the ANSI 3 m AFs should be used for best results (the ANSI method ignores near-field terms, but this cancels out if the AFs and NSA are both calculated using the formulas given in the ANSI standards).

A4.7 Uncharacteristic resonances.

Some LPDA antennas with screwed on elements sometimes develop one or several sharp resonances. Sometimes an oxide layer is apparent at the join of the element to the boom, and cleaning away this oxide can cure the resonance. This phenomenon is noticed more when the antenna has been used with higher powers, such as used in immunity testing. This phenomenon has not been noticed on LPDA antennas whose elements are welded to the boom. Some LPDA models have sharp resonances that are inherent in their unfortunate design and show up on the brand new antenna.

A4.8 Mutual coupling between large horn antennas 1 m apart.

The ARP958 method calls for two identical antennas to be mounted with their faces 1 m apart. Horn antennas specified to cover the frequency range 200 MHz to 2 GHz have aperture dimensions of typically 0.94 m x 0.68 m and there is substantial mutual coupling between them, which shows as an oscillation in a plot of antenna factor against frequency. When one of these antennas is used for EMC testing, the object being tested is unlikely to present the same mutual coupling. Approximately 2 dB of additional uncertainty should be allowed for.

Appendix 5 Emission measurements in screened rooms

In the early days of EMC testing, a lot of the measurements were carried out in screened rooms which acted like resonant cavities. Because of the high Q-factor of the rooms the uncertainties in measured field strength could be of the order of \pm 30 dB. At best the measurement served to identify the frequency of emission and to show up differences between one EUT and another measured in the same way. A displacement of the antenna of the order of 10 cm could cause the received signal to vary by 20 dB. In this scenario it did not matter much if the uncertainty in the antenna factor (AF) was of the order of \pm 3 dB. One advantage of the unlined room is that it is more like a reverberation chamber, ensuring that emissions not directed at the antenna, e.g from the top and back of the EUT, have a chance of being detected.

These days better quality control demands that some sort of calibration is performed, if only to prove that the antenna is working as the manufacturer intended. The trend is to use absorbing material to lower the Q-factor of screened rooms⁴⁶. In this case antenna factor uncertainty can be an important contribution to the overall uncertainty budget. Calibrations of antennas for use at distances of 1 m from the EUT and associated uncertainties are discussed in Section 9.6.5.

Screened rooms of approximate size 6 m x 4 m and 2.5 m high are commonly used for emission measurements to Defence Standards⁴⁵. For frequencies below 30 MHz any absorber present may not be very effective so the room may considered to be undamped. Also rooms of this size will be operating below the cavity cut-off frequency. This means that the radiated field from the EUT is not propagating and is likely to differ in magnitude from the field measured at the same distance in free-space.

MIL-STD-461⁴⁴ specifies the ARP958 method³⁷ of calibrating antennas for use at a distance of 1 m. However this method involves calibrating an antenna in open space, for subsequent use in a screened room. If one makes the assumption that a meaningful measurement of an EUT is one that gives the magnitude of radiated field strength when the EUT is operating in open space, a special calibration of the antenna for use in a screened room is required.

An appropriate method of calibration was investigated. In this investigation a distance of 2 m was used between the antenna and the EUT, but the method equally applies for 1, m. A rod antenna was used to measure a 10 cm diameter spherical dipole emitter, both on an OATS and in a screened room. The difference between the measurements was used to correct a subsequent measurement of an EUT in a screened room with the intention of giving the field that would have been obtained if the EUT had been measured on an OATS. It was necessary to also use a loop antenna to measure the spherical dipole. Furthermore it was necessary to replace the spherical dipole with an emitting loop and to measure its field with a rod and a loop antenna in the screened room and on the OATS. The correction factor that applies to the EUT depends on whether the EUT is primarily an E-field source or an H-field source.

The full procedure is described in Refs. 45, 46 and 47. It was found that the field radiated by the EUT in open space differed by about 10 dB from the field measured in a screened room operating below its cut-off frequency, when simply using the free-space antenna factors of the rod and loop antennas. When the above procedure was followed, the field measured in the room agreed, after correction, to within about 2 dB of the field measured on an OATS.

Appendix 6 Antenna calibration services offered by NPL

As a National Measurement Institute, NPL has studied extensively the calibration of antennas and the uses to which they are put in order to assist industry to improve their measurement uncertainty. NPL has put in a lot of effort since 1988 to improve traditional calibration methods and to develop new methods, also addressing the measurement uncertainties associated with antennas in EMC testing of products. It is one thing to have a value of uncertainty in the calibration certificate for the measurement of free-space antenna factor, but another factor that must be addressed is the uncertainties associated with the use of the antenna; for example the uncertainties arising from the proximity of the antenna to the ground plane, from near-field effects and from the combination of measurement geometry and radiation pattern. Where appropriate, NPL issues customers an Annex to the calibration certificate that addresses some of these additional sources of uncertainty.

Staff at NPL have expert knowledge of the performance of most types of EMC antennas on the market. Some 28 different models of biconical antenna and 28 models of LPDAs have been measured to date, plus 10 models of hybrid broadband antennas, 24 models of dipoles, 13 models of rod antennas, 3 models of spiral antennas and 5 models of EMC horn antennas. Unexpected performance characteristics on some models of antenna are known and the staff are alerted to identify these and to advise the customer accordingly.

The choice of calibration is governed by the final measurement uncertainty that the user of the antenna expects to achieve. Different laboratories accredited for antenna calibration offer different uncertainties. For some models of antenna NPL offers two levels of uncertainty, standard or low uncertainty. These are typically ± 1 dB and ± 0.5 dB respectively, though in some cases ± 0.3 dB can be achieved for wire antennas, and ± 0.04 dB can be achieved for standard gain horn antennas. The standard uncertainty services are more economical so the price is lower for the customer. To achieve lower uncertainties more sophisticated measurement procedures are adopted and sometimes verification is made using an independent method. It is important that the customer should confirm the best measurement uncertainty when inquiring about antenna calibration.

Calibration of antennas by the procedure of ANSI C63.5 is popular. However this is a laborious procedure because it involves height scanning of 3 pairs of antennas and strictly it should be done on a near-ideal ground plane. However in popular usage the ground plane is often not ideal and this increases the uncertainty of the measured antenna factor. For these reasons NPL recommend the use of AFfs. For antennas which are used for emission testing on a 10 m OATS or in free-space conditions as in a fully anechoic chamber, the use of AFfs will result in lower uncertainties of measurement than the use of AF_{ANSI}. For a 3 m OATS (or semi-anechoic chamber) the uncertainties will be similar.

The measurement method can be tailored to ensure that the uncertainties associated with the use of the antenna are minimised. A classic case is the use of antenna for site evaluation, where validation of a costly EMC facility is at stake and precision measurements are expected. The antenna that remains at a fixed height for the NSA test of an OATS can be calibrated at that fixed height, which removes the uncertainty component associated with mutual coupling to the ground plane, which can be ± 1 dB for a biconical antenna. NPL is able to provide this type of calibration.

At first approach the customer can find out from the NPL web site what measurement services are available. The NPL website address is:

www.npl.co.uk

For antenna calibrations, follow the link for Measurement Services, then RF and Microwave Free Field Standards. More information is obtained by clicking on the title of each section of the table. There are further antenna calibration services, for example of parabolic dish antennas, that are listed on the website but not referred to in this guide.

Nearly all calibration services are UKAS accredited and the best measurement uncertainties can be seen on the UKAS schedule. On the Measurement Services pages there is a link to UKAS accreditation, on which there is a link entitled 0478 Laboratory Scope. Clicking this link opens up the UKAS web address where one can view the full Schedule for NPL. The uncertainties for antenna calibration are in the sections headed Antenna Gain and Antenna Factor. The UKAS home page is <u>www.ukas.org</u>.

The uncertainty is based on a standard uncertainty multiplied by a coverage factor of k = 2, providing a level of confidence of approximately 95%. All calibrations include measurement of return loss. Biconical antennas, tuneable dipoles and hybrid broadband antennas are tested for balun imbalance if the balance for that model is known to be poor. Broadband antennas are calibrated by stepped frequency technique using an interval of 1 MHz for biconical antennas and 2 MHz for LPDA antennas.

The uncertainties may be increased (a) for antennas which show significant balun imbalance, (b) in the region of sharp antenna factor resonances and (c) where the polarisation of spiral antennas is not perfectly circular. Antenna factors for log-periodic dipole arrays (LPDA) are referred to the phase centre. Additionally antenna factors can be supplied with reference to the mechanical centre of the LPDA. Also antenna factors can be supplied which give the corrected field strength at a distance of 1 m from the tip of the antenna, or other distance on request. Alternatively antennas can be calibrated to the SAE ARP958 procedure on request.

In some cases it is possible to achieve lower uncertainties, confidence for which is obtained by NPL's participation in international intercomparisons. Such services are not yet UKAS accredited and it may be worth the customer enquiring about them.

The Customer Service Executive for antennas can be contacted on tel +44 (0)20 8943 6796

or via the NPL helpline:

enquiry@npl.co.uk

at the address: National Physical Laboratory Hampton Road Teddington Middlesex United Kingdom TW 11 OLW

