

NPL REPORT

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**COMPARISON OF
CALIBRATION METHODS
FOR MONOPOLE
ANTENNAS, WITH SOME
ANALYSIS OF THE
CAPACITANCE
SUBSTITUTION METHOD.**

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M J Alexander**

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COMPARISON OF CALIBRATION METHODS FOR MONOPOLE ANTENNAS, WITH SOME ANALYSIS OF THE CAPACITANCE SUBSTITUTION METHOD.

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Division of Enabling Metrology

ABSTRACT

This report describes an investigation into the theoretical basis for the equivalent capacitance substitution method (ECSM) of monopole calibration, and the uncertainty in applying this method. An important part of this work was to compare the results obtained by ECSM to the free-field calibration technique developed using a MEB1750 GTEM cell at NPL, which replicates a plane wave environment. The project resolved the main sources of error in applying the ECSM, and found good agreement between the two methods of calibration.

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Approved on behalf of the Managing Director, NPL
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1 INTRODUCTION

1.1 Background to monopole calibrations

For historical reasons many monopoles used for EMC emission measurements have elements with length given in imperial units (41 inch), which at first seems like an odd choice. The origin of this length goes back to the receiving antennas used on certain types of US aircraft during WWII. The long wire antennas were attached to the outer skin of the aircraft and they were connected, through a hole in the fuselage, to the radio receiver by a length of insulated wire. This wire happened to be 41 inches long, and the particular gauge of wire which was used meant that the effective capacitance was around 10 pF at frequencies below 20 MHz.

In order to measure the sensitivity of the radio system to other interference signals inside the fuselage a 41 inch monopole was used to simulate the short section of connecting wire. In a form of early EMI testing, the 41 inch monopole could be used to measure the interference from other pieces of equipment inside the fuselage, and this would give an indication of the sensitivity of the radio to this interference. They did not worry about the external antenna because it was shielded from the equipment inside, and the apparent impedance at the feed-point (i.e. the hole where it entered the aircraft) was very high.

This value of 10 pF as a substitute for 41 inch monopoles has found its way into the ANSI C63.5-1998 standard which describes the equivalent capacitance method for calibrating monopoles. Other sources suggest that 12 pF is a more accurate value to use, and this can be confusing. In the early days of EMC a few decibels of error in the monopole calibration was insignificant when compared to the overall uncertainty, but these days EMC measurements are more accurate and there is a need to reduce the calibration uncertainties.

Calibration of monopole antennas using free-field propagation is problematic because of the large wavelengths at frequencies below 1 MHz and the fact that an 'infinite' ground plane is required as reference. However, there are modern techniques which can achieve reasonable accuracy for free-field antenna factor.

1.2 Summary of Equivalent Capacitance Substitution Method (ECSM)

The essential idea behind the ECSM is that the monopole element is replaced by a capacitor which simulates the reactive part of the monopole element at low frequencies, when the monopole operates with reference to 0 Volt potential on 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.

In practice an 'infinite' ground plane is impossible to realise, however most EMC testing standards which use monopoles are designed to give an approximation to this ideal within practical limits. A typical Open Area Test Site (OATS) usually gives a good approximation to an 'infinite' ground plane, as long as the edges are well grounded to earth and there are no buildings near by. Many commercial monopoles are supplied with a small ground plane (usually 0.6 m × 0.6 m), or radial wires, which should ideally be referenced to the common ground of the relevant 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 (Ref [1]) gives the following expression for calculating the correct 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.

$$C_a = \frac{55.6 \times h}{\left(\ln\left(\frac{h}{a}\right) - 1 \right)} \cdot \left[\frac{\tan\left(\frac{2\pi h}{\lambda}\right)}{\frac{2\pi h}{\lambda}} \right]$$

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 $C_a = 11.9$ pF, hence the commonly quoted figure of 12 pF.

ANSI C63.5-1998 gives the same expression except that the fraction inside the natural logarithm is $(2h/a)$. For the same 1 m element described above, this ANSI expression gives a capacitance value of 10.4 pF, which is the reason why you also find 10 pF quoted as the capacitance value for a 41 inch element. Later in this report there is a theoretical derivation which shows that the CISPR 16-1-4 expression is correct. It is possible that the ANSI expression might originally have been written in terms of diameter, and the $(2h/a)$ remained even after a was later redefined as radius.

The ECSM requires an adaptor box to be constructed. The correct value of capacitance is installed inside this box, which 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 coax connector, and this signal then passes through the capacitor and directly into the input terminal. Figure 1-1 shows a schematic of a typical configuration for the ECSM.

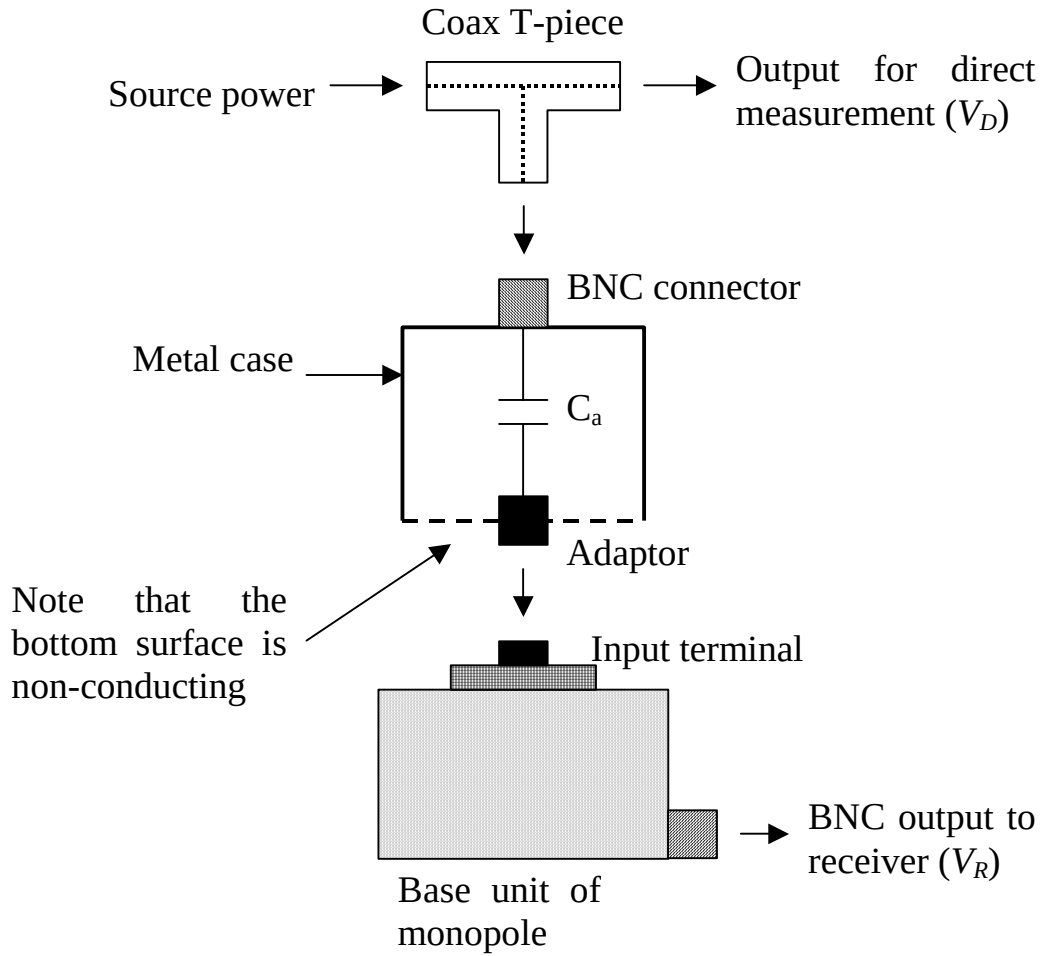


Figure 1-1: Schematic of ECSM calibration system.

The coax 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 we can measure the 'direct' voltage (V_D) with a $50\ \Omega$ load on the output of the monopole and then, when we measure the output voltage (V_R), the $50\ \Omega$ load is connected to the output from the T-piece so that the source always sees the same impedance, and 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. $-20 \cdot \log_{10}[h_E]$). Occasionally, the expression for antenna factor is quoted with just +6dB included for the effective height

correction term, which can be misleading if the monopole being measured does not have a 1 m element.

Figure 1-2 shows the equivalent circuit for a receiving monopole, and from this representation the derivation of the expression for antenna factor is straight forward because we know that the open circuit voltage is given by h_E multiplied by E-field strength.

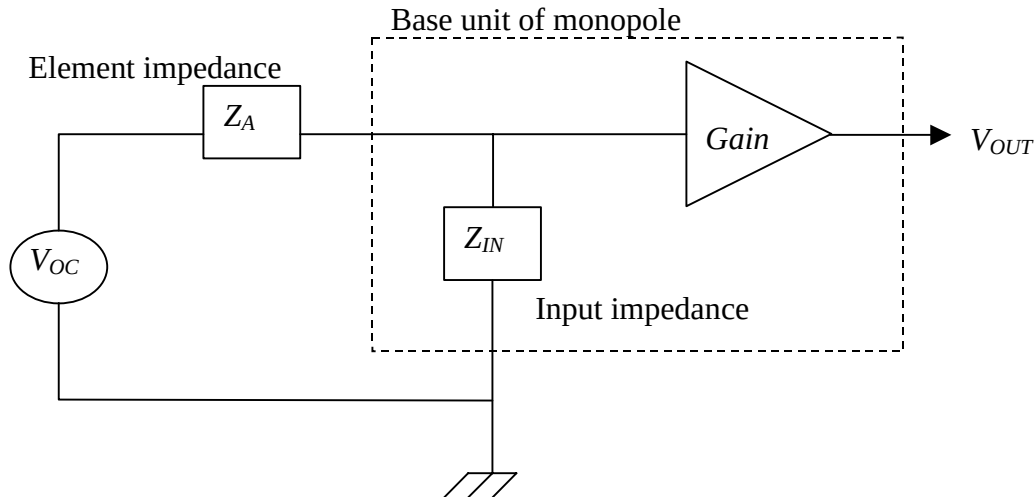


Figure 1-2: Equivalent circuit for receiving monopole.

During an ECSM calibration the V_{OC} is supplied by the source (V_D), and V_{OUT} is measured (V_R), so we get the following simple relationship which immediately gives us antenna factor:

$$\frac{V_D}{V_R} = \frac{Z_A + Z_{IN}}{Z_{IN} \cdot \text{Gain}} = \frac{V_{OC}}{V_{OUT}} = \frac{E_{field}}{V_{OUT}} \cdot h_E = AF \cdot h_E$$

The aim of this report is to study how sensitive the ECSM calibration is to variations in each parameter, and to compare antenna factor obtained from ECSM and a free-field type measurement.

1.3 Free-field calibration techniques

The major problem with free-field measurement of monopoles at low frequency is that the distance required for normal free-field propagation is enormous, which means that screened rooms are much too small. The two methods described here use the guided wave approach to generate low frequency (far-field) TEM waves in a very similar way to standard TEM cells. Once generated, the E-field may be calibrated by measurement with an independent reference antenna, and then the unknown antenna may be placed in the same E-field. In this way the ratio of calibrated field strength to output voltage gives a direct measure of antenna factor.

The first example of a calibration technique is called the long wire antenna. This is a wire suspended above a conducting floor (normally in a screened room), which is supplied at one end with a known voltage and matched at the other end to prevent excess standing waves developing. In this way the screened room is converted into a large wave guide. There are various documents (Ref [2]) which give guidance on how to match this long wire to a 50 Ω system, and with good design the E-field strength is approximately given by source voltage divided by the height of the wire above the floor. If the E-field is fairly uniform, a calibrated antenna may be used to determine the exact field strength at the test position, and then the antenna factor of the AUT may be measured. However, because this is a relatively complex system the accuracy of calibration can be very variable from one test laboratory to another.

The second approach uses a GTEM cell to generate the E-field and the AUT is placed in this field, with the base in contact with the floor of the cell. The E-field strength is calculated from the input power and the height of the septum. Below 5 MHz the GTEM environment is a fair approximation to free-space for a monopole because the impedance of the element is so large that there is minimal coupling with the cell walls. Above 5 MHz, where the coupling may be more significant, free-space antenna factor can still be measured by the standard antenna method (SAM) using a reference monopole. This works because, even with moderate coupling to the side walls, the relative change in received voltage is approximately the same for both the AUT and the reference standard. Above 30 MHz the measurement needs to be done on a ground plane because this is close to the resonant frequency of a 1 m element (75 MHz).

The GTEM technique has been developed at NPL (Ref [3]) so that the antenna factor of the AUT may be measured from 100 Hz to 30 MHz with an uncertainty of better than ± 1.5 dB ($k=2$). Above this frequency the antenna factor of the AUT is measured on a 30 m by 60 m ground plane by substitution with a passive standard monopole. The antenna factor of the passive standard monopole (reference antenna) has been calculated by numerical simulation using NEC (Ref [4] & [5]), and the results have been verified directly by measurement on the ground plane.

Whereas the ECSM simulates the performance of the monopole element, a free-field calibration will measure the performance of the whole monopole antenna, including the pre-amp casing at the base of the element, which provides added confidence that there is good contact between the element and input terminal of the base, and also that the element itself is functioning correctly. However, free-field techniques require much more careful setup to achieve accurate results (± 1.5 dB), whereas a fairly approximate ECSM measurement should give antenna factor within ± 3 dB of the correct value with minimum effort. The GTEM cell at NPL is an MEB1750 which has a maximum height of 1.75 m at the test volume position.

2 DERIVATION OF THE CORRECT VALUE OF CAPACITANCE

2.1 Theory of equivalent capacitance

The purpose of this section is to trace the origin of the equation which gives the self-capacitance of a monopole antenna. After the equation has been justified, the relative impact of each factor within the equation will be investigated. In CISPR 16-1 the self-capacitance of a rod antenna is given as follows:

$$C_a = \frac{55.6 \times h}{\left(\ln\left(\frac{h}{a}\right) - 1 \right)} \cdot \left[\frac{\tan\left(\frac{2\pi h}{\lambda}\right)}{\frac{2\pi h}{\lambda}} \right]$$

Equation 1

Where:

- C_a = self-capacitance of the rod antenna, in Pico farads
- h = actual height of the rod element, in metres
- a = radius of the rod element, in metres
- λ = wavelength, in metres

The sources given for this equation are Ref [6] and [7]. We can show how this equation is derived from the formulas given in the book by Schelkunoff and Friis. In Chapter 10 of Ref [6] the capacitance of thin dipole antennas with conical input tips, but of arbitrary shape beyond the input tips, is deduced and the equation is reproduced below. For logarithms, Schelkunoff uses 'log' in all his equations, and it has been confirmed from independent sources that this denotes the natural logarithm. Here we use 'ln' to avoid confusion with log to base 10.

$$C_{an} = \frac{\pi\epsilon (l - \frac{1}{2}s)}{\ln\left(\frac{2l}{a_{lm}}\right) - 1 - \ln 2 + f(s, l)} + \frac{\pi\epsilon s}{2\ln\left[\frac{2s}{\rho(\frac{1}{2}s)}\right]} + 2\epsilon\rho(l)$$

Equation 2

Where:

- C_{an} = antenna capacitance
- l = length of one arm of the dipole
- s = total length of both conical input tips
- $\rho(z)$ = radius of the antenna at any point z , according to Figure 2-1

The function $f(s,l)$ is a small correction term, which we will deal with later, and a_{lm} is called the *logarithmic mean radius* of the antenna. For many monopole antennas the radius of the element does not change over the antenna length, so a_{lm} becomes just antenna radius a , and $\rho(z)$ is a constant for all $z > s$. These parameters are illustrated in Figure 2-1.

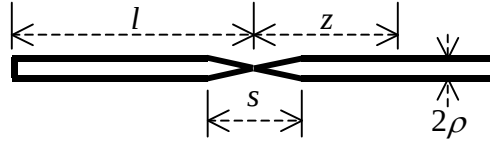


Figure 2-1: A dipole antenna

Explanation of logarithmic mean radius

The expression for *logarithmic mean radius* is given in Ref [6], and for a telescopic element it may be written in a discrete form, with the element divided into i segments:

$$\ln(a_{lm}) = \frac{1}{\sum l_i} \cdot \sum l_i \times \ln(r_i)$$

Where:

l_i = length of i th segment

r_i = radius of i th segment

For a fairly typical telescopic 41 inch element, when this expression was used to calculate capacitance from Equation 2, this produced a value of 12.09 pF. However, by using just a simple average of all the segments, the calculated figure was 12.13 pF, which suggests that for typical types of monopole an accurate value of capacitance may be derived using the average radius of the element.

Explanation of the function $f(s,l)$

There is an approximation given for the correction term $f(s,l)$ which is reproduced below. For some typical dimensions of a monopole element ($l = 1$ m, $s = 4$ cm, $a = 3$ mm) this function returns a value of 0.08 which may be compared to the sum of the other terms in the denominator of the first part of Equation 2. For these typical dimensions the sum of the other terms is 5.8, so it can be seen that this function is negligible.

$$f(s,l) \approx \frac{s}{2l-s} \ln \frac{2l}{s}$$

Removing 2nd and 3rd terms from Equation 2

The last term in Equation 2 represents the capacitance between the flat outer ends of the dipole. For a radius of 15 mm (relatively large for a wire antenna) this term is only 0.3 pF, which is small relative to the total capacitance value and therefore this term is usually ignored. However, in the case of top loaded elements, which usually have some kind of plate or radial wires at 90 degrees to the main element, the capacitance at the ends cannot be ignored. Top loaded monopoles are discussed in Section 2.3.

The second term of Equation 2 represents the approximate capacitance of the conical input tips. Again, this term is usually assumed to be negligible because, as the *logarithmic mean radius* approaches zero and the ratio (s/a_{lm}) is held constant, the capacitance approaches the following asymptotic limit:

$$C_{an} = \frac{\pi \epsilon l}{\ln\left(\frac{2l}{a_{lm}}\right) - 1 - \ln 2}$$

The length of the input region does not appear in this expression, so for practical dimensions the effect is considered to be small. It is worth noting that the second and third terms in Equation 2, which are relatively small, both act to increase the calculated value of capacitance. As we will show later, if these terms are significant, they will effectively reduce the overall antenna factor (i.e. make the antenna more sensitive).

Capacitance of a monopole with constant radius on a ground plane

When all the second order terms are removed we end up with something which looks more like Equation 1. However, Equation 2 was derived for a dipole, and we want to know what the capacitance value is for a monopole. A monopole (length l) on a perfect ground plane is equivalent to a dipole (length $2l$) in free-space, except that the impedance of the monopole is half that of the dipole. At low frequency this effectively means that the capacitance of the monopole is twice that of the dipole, so for a monopole:

$$C_{an} = \frac{55.6 \times l}{\ln\left(\frac{l}{a}\right) - 1} (pF)$$

The Tan() factor in Equation 1

The second term in Equation 1, which is referred to here as the Tan() factor, is an approximate correction for the change in electrical height of the antenna at frequencies approaching the natural resonance of the monopole. The approximation breaks down in the region of the resonant frequency where the reactive part of the impedance is close to zero. We can easily plot out the ratio expressed by this factor, and this shows that it is very nearly unity for very large wavelengths, but it increases rapidly as the height becomes close to $\lambda/4$ (i.e. resonance).

$$Ratio = \frac{\tan\left(\frac{2\pi h}{\lambda}\right)}{\left(\frac{2\pi h}{\lambda}\right)}$$

The ECSM should give best results when the height is a fraction of wavelength: in fact the limit of the method is often quoted as $h < \lambda/8$, which is 36 MHz for a 41 inch element.

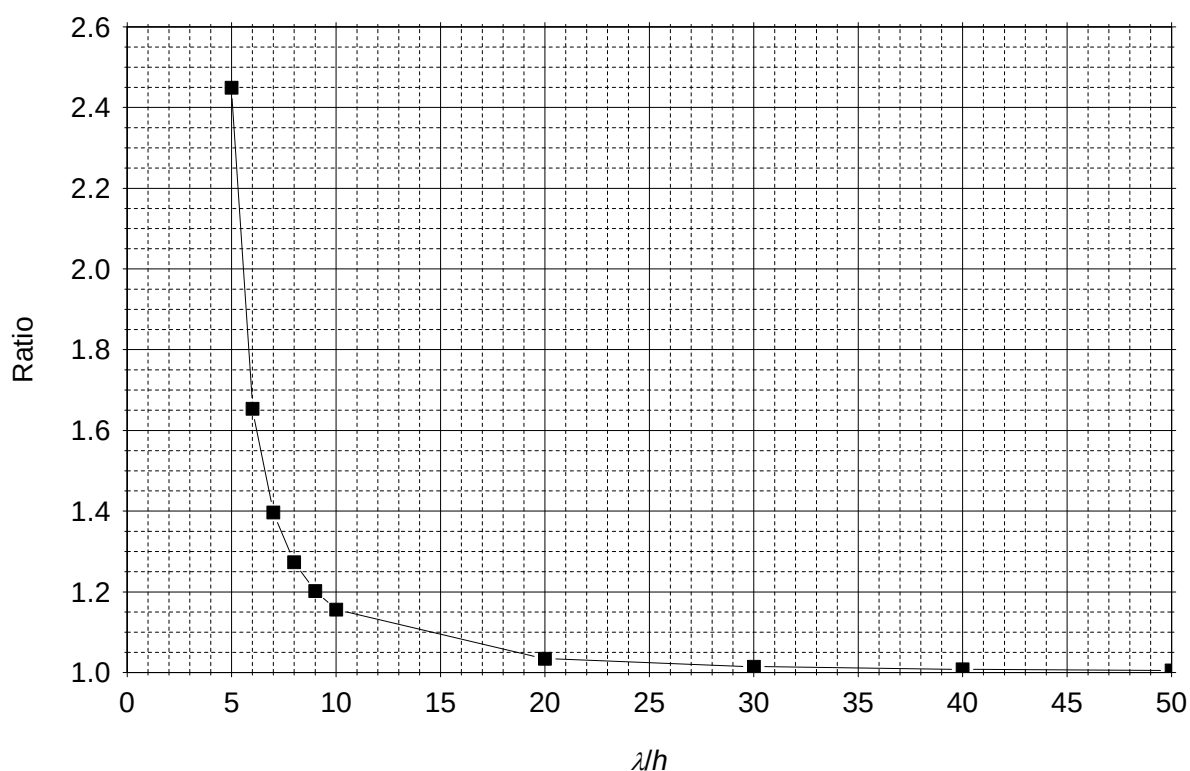


Figure 2-2: Graph showing the magnitude of the Tan() factor in the CISPR expression.

2.2 Comparison of calculated capacitance to numerical simulation

As a check on the theoretical expression for capacitance we modelled some monopoles in NEC, using different dimensions, and compared the value of capacitance from Equation 1 to the value obtained from NEC. We could derive the capacitance value from the input impedance which is given by the NEC output, using the fact that the reactive part of the impedance (X_A) is given by:

$$X_A = \frac{-1}{2 \cdot \pi \cdot f(\text{Hz}) \cdot C_A(\text{farad})} = \frac{-1.592 \times 10^5}{f(\text{MHz}) \cdot C_A(\text{pF})}$$

The following graphs show the comparison of results for three different dimensions, which give capacitance values ranging from 12 pF to 21 pF. This range should cover the vast majority of monopole elements used in practice. Considering the possible uncertainty in the impedance from the NEC model, particularly for electrically short antennas, the level of agreement shown here is good. The graphs illustrate how capacitance is constant at low frequencies, but it changes quite rapidly when the $\tan()$ factor becomes significant, particularly for the 2 m element which has a relatively low resonant frequency. The ECSM uses a fixed value of capacitance so, when the real value begins to change near resonance, inevitably there will be errors in the calculated antenna factor.

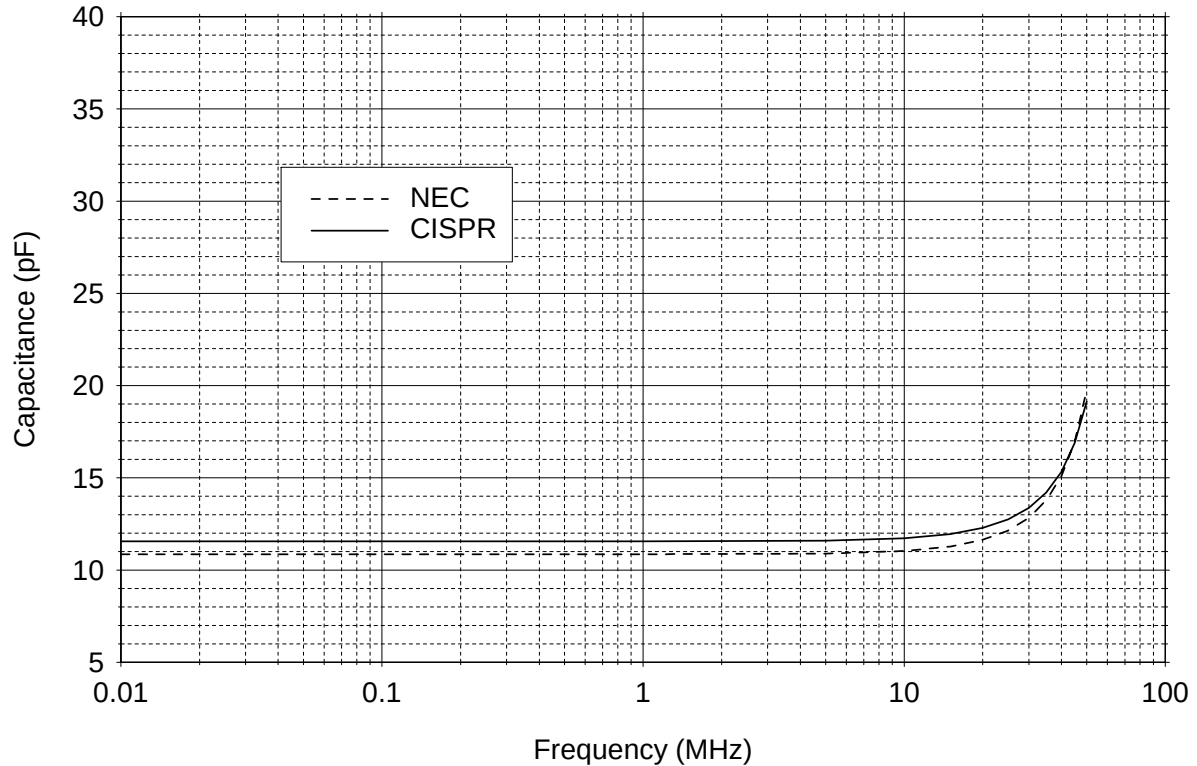


Figure 2-3: Comparison of capacitance for: $h = 1$ m, $a = 0.003$ m

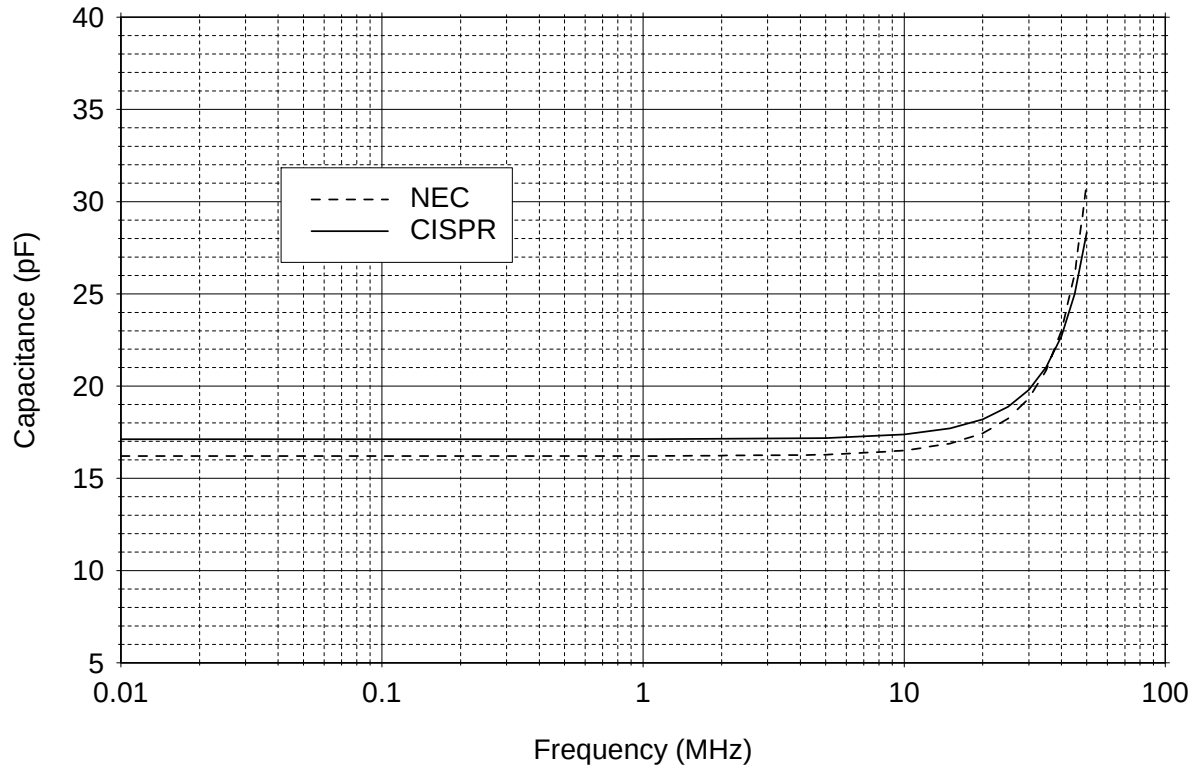


Figure 2-4: Comparison of capacitance for: $h = 1$ m, $a = 0.0143$ m (or $2a = 9/8$ inches)

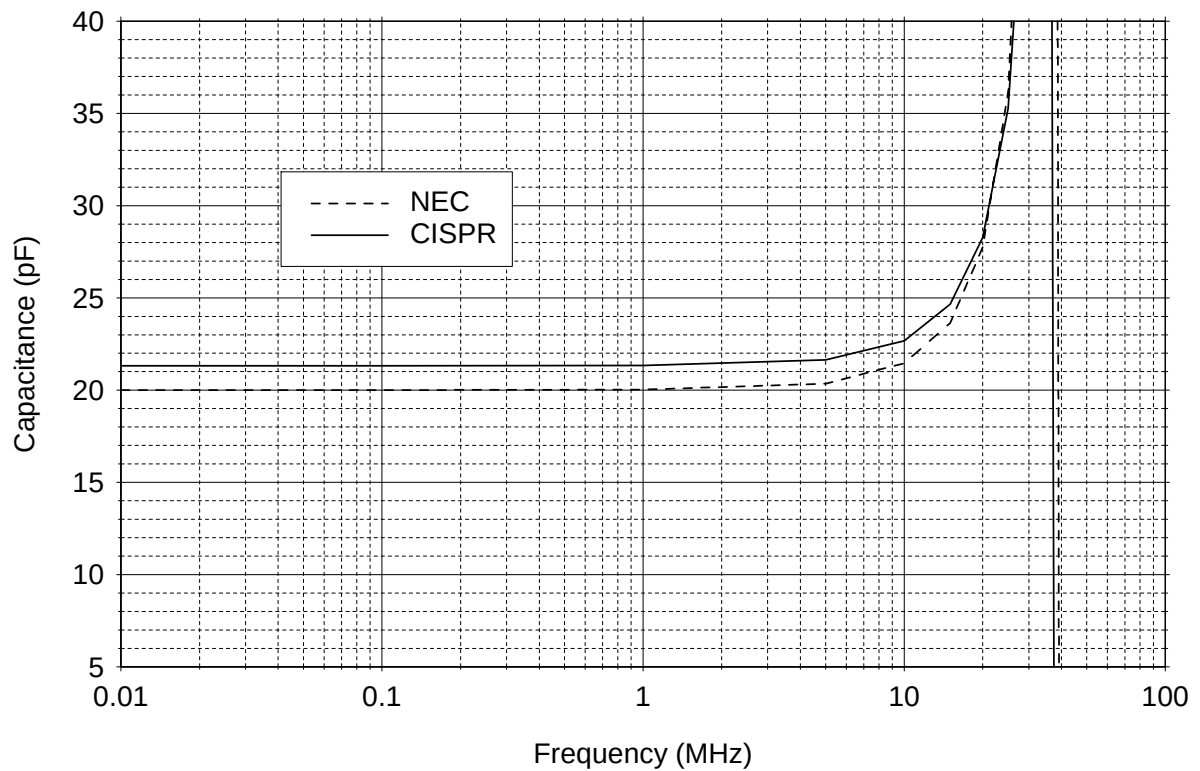


Figure 2-5: Comparison of capacitance for: $h = 2$ m, $a = 0.004$ m

2.3 Treatment of top loaded monopoles

Top loaded monopoles have either a plate or a configuration of wires attached to the top of the element. The attachment is usually orthogonal to the main element, and it is designed to increase the current distribution near the top of the element (Ref [8]), which increases the effective electrical height of the element, thus making the monopole more sensitive. This can be a useful technique to maximise signal strength when the antenna has restricted space.

The addition of a top plate makes the calibration a little more awkward. During the GTEM free-space calibration method, the standard antenna method approach requires a like-for-like substitution. To achieve this a standard has to be constructed which is physically similar to the AUT, and the antenna factor of the new standard is then derived from NEC simulation as before. This process needs to be done for each design of top loaded monopole.

During the ECSM we need to know what value of capacitance to use and what effective height to put into the expression for antenna factor. The usual expressions only apply for simple wire elements, so they cannot be applied to the case of top loaded elements. In some cases it may be possible to derive a theoretical capacitance, however the safest option is probably to model the element in NEC (or similar code). This section will describe a pragmatic approach to obtain capacitance for top loaded elements.

We already have antenna factor for one NPL standard, both with and without a circular 12 inch top plate. Therefore we study this case first, and we can then compare the result to the NEC antenna factor. During an ECSM measurement the coax T-piece is connected to the capacitance, and for straightforward passive antennas the output voltage from the capacitor will be V_R (see Figure 1-1) because there is no pre-amp stage. For such a simple construction it can be shown that the equivalent circuit for the ratio (V_D/V_R) is just a potential divider consisting of the capacitor and a $50\ \Omega$ characteristic impedance. At low frequency the capacitor is approximately an open circuit so the loss from source (through the T-piece) to V_D should be small, and the same voltage (V_D) appears at the input to the capacitor.

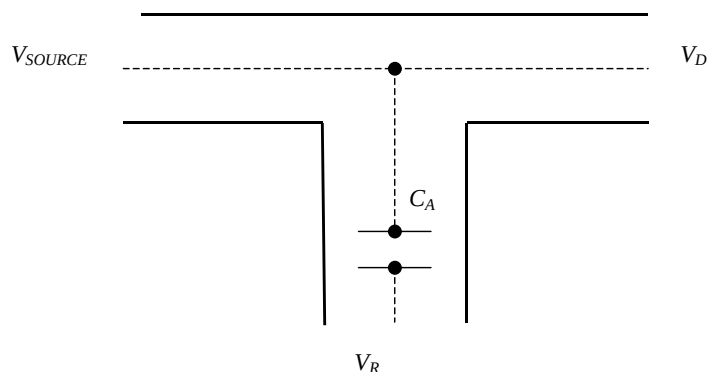


Figure 2-6: Schematic of coax T-piece and capacitor for passive element

From Figure 2-6 it is possible to calculate the ratio (V_D/V_R), and from here antenna factor is given by the standard expression for ECSM.

$$\left| \frac{V_D}{V_R} \right| = \sqrt{1 + \left(\frac{10000}{\pi \cdot f_{MHz} \cdot C_{pF}} \right)^2}$$

$$AF(dB/m) = 10 \cdot \log_{10} \left[1 + \left(\frac{10000}{\pi \cdot f_{MHz} \cdot C_{pF}} \right)^2 \right] - 20 \cdot \log_{10}(h_E)$$

In the standard CISPR 16-1-4 equations the capacitance is just a function of h and a , and h_E is a function of h . When we input the dimensions of the NPL standard (without top plate) the above expression agrees to within 0.25 dB with the NEC antenna factor, which gives good confidence in the NEC result. The derived capacitance for the un-loaded element was 16.2 pF.

The next stage is to model the top loaded element in NEC, and calculate the antenna factor. Using this NEC data as reference we can increase h in the above expression, keeping a constant, until the analytical antenna factor agrees with the NEC antenna factor. Thus we have found the equivalent h for the top loaded element, and the correct capacitance to use in the ECSM. In the case of the NPL standard, the equivalent height was 1.9 m and the capacitance value was 21.2 pF.

A common top loaded design is the Singer 95010-1. When processed as above, the equivalent height for the 50 inch element (plus plate) was 1.79 m, and the capacitance value was found to be 18.5 pF (using the average radius of the 50 inch telescopic element).

Top loading the element increases the effective height of the antenna, which therefore reduces the resonant frequency. Because the real capacitance value begins to deviate from the fixed ECSM value when near resonance, this means that the ECSM for top loaded antennas may be less accurate at the high end of the operational frequency band.

3 DESIGN OF ECSM CALIBRATION BLOCK

3.1 Critical aspects of design

ECSM requires an adaptor stage, or calibration block, which is connected to the input terminal of the AUT (where the element is normally attached). This adaptor allows the input RF signal to be passed through the capacitor and into the input terminal of the AUT. To achieve the most accurate measurements some precaution is required in the construction of the adaptor.

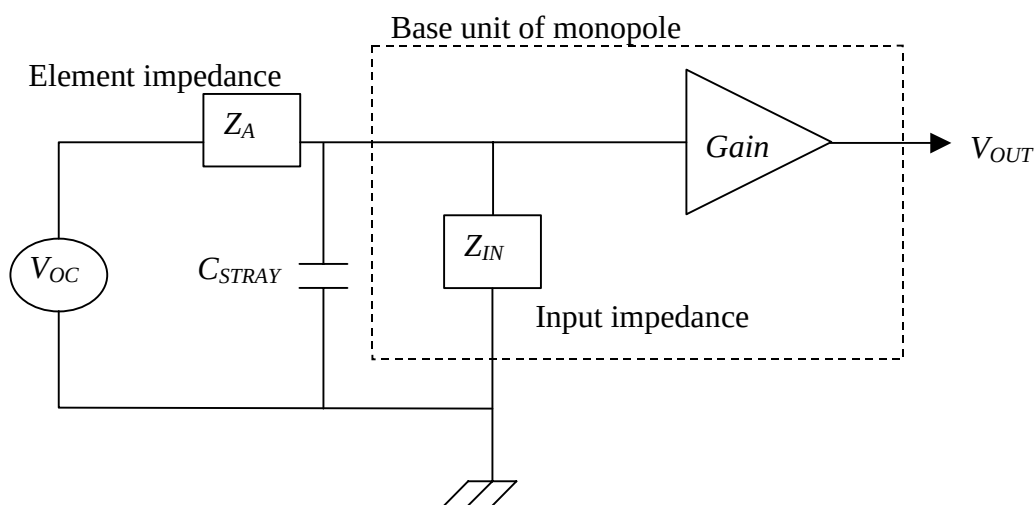


Figure 3-1: Circuit diagram of receiving monopole with stray capacitance.

Figure 3-1 shows the circuit of a monopole which illustrates where a stray capacitance (C_{STRAY}) may occur when the ECSM adaptor is attached to the monopole input. In practice C_{STRAY} represents the capacitance between the wires in the adaptor and the metal outer casing (see Figure 3-2).

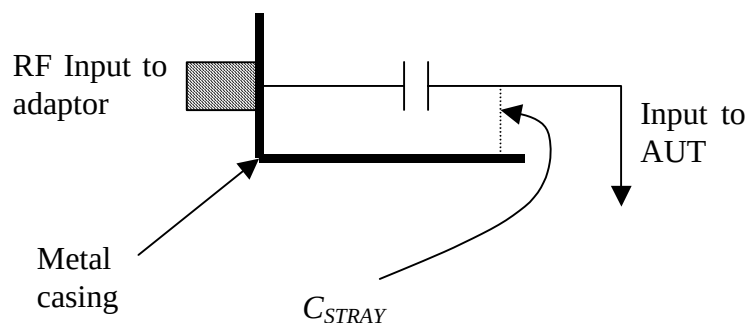
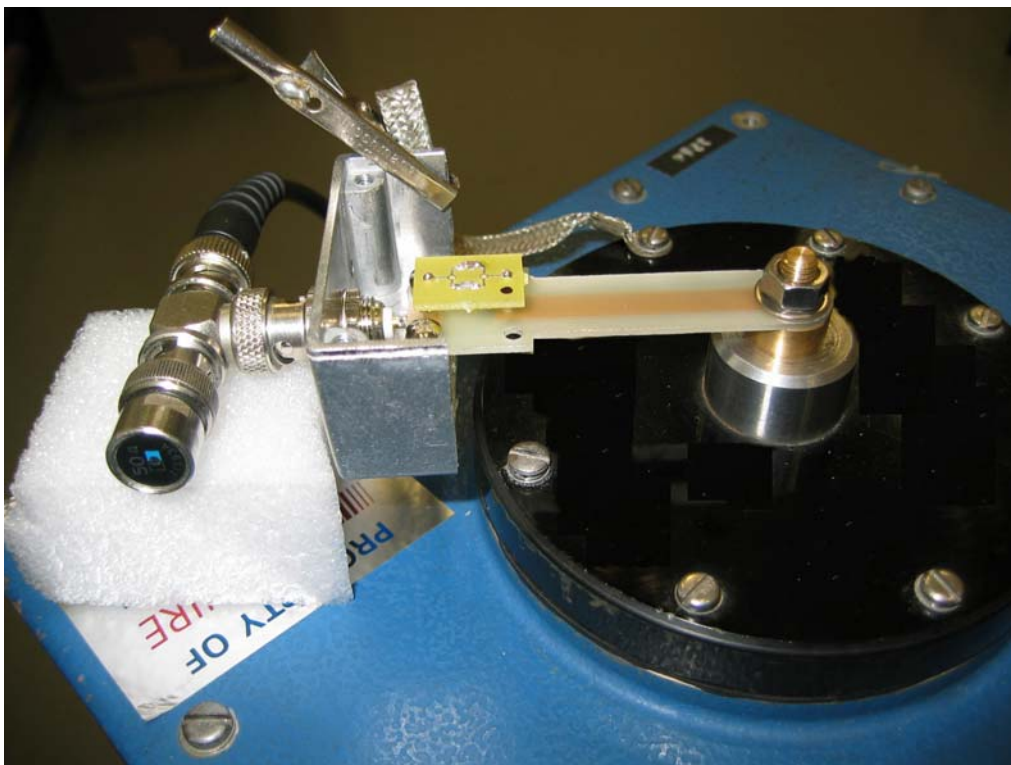


Figure 3-2: Schematic illustrating possible stray capacitance in ECSM adaptor.

A simple calculation shows that the effect of this stray capacitance is to reduce the voltage input to the AUT which effectively causes the ECSM measured AF to be larger. The size of the increase depends upon the value of C_{STRAY} and the input impedance of the monopole. For example, a 1 pF stray capacitance has an impedance of 1.6 M Ω at 100 kHz and a typical value for the capacitor used for ECSM (12 pF) is 130 k Ω at the same frequency. If we assume the input impedance to the AUT is much larger (perhaps 100 M Ω) then we can view the stray capacitance as a voltage divider acting with the nominal 12 pF in the adaptor. In this case, a 1 pF stray causes a 0.7 dB increase in measured AF.

During experimental work a real value of 1 pF was added as a stray capacitance in a typical ECSM setup, and the measured AF increased by 0.4 dB at 100 kHz which is in line with expectation. Obviously, to prevent these kinds of errors it is important to construct the adaptor such that the conductor connecting the capacitor to the AUT input does not pass close to other conductors at zero (ground) potential. In the published standards there are some guidance notes on the construction of the ECSM adaptor, and the illustrations seem to show the capacitor fully encased in a metal box which, as discussed above, may couple with the wires in the box (particularly where the wire exits the box and connects to the AUT). In Ref [1] the recommended spacing from the internal wires to the casing is 5 mm to 10 mm, but a better solution would be to support the wires with a non-conducting material. After some experimentation the adaptor we finally used at NPL was constructed and it is shown in the picture below.



The function of the small brass section, which is visible on the AUT input terminal, is to adapt from the particular screw size of the AUT input to a size suitable for the contact hole in the ECSM calibration attachment. Thus, with a range of different adaptors, the ECSM calibration block may be attached to different monopole designs (see below).



The capacitor is contained on the yellow PCB (Printed Circuit Board) which is supported on a dielectric strip. The RF is carried on a copper strip which is just visible on the underside of the dielectric. This design is very basic, and there is a minimum of conducting material used after the capacitor PCB, with the aim of reducing stray coupling after the capacitor stage. The adaptor is grounded via a braid attached to the small metallic half-box which holds the BNC input connector. This grounding strap is connected to the ground point on the AUT during calibration.

Good contact between the calibration block and the AUT input is vital for accurate measurement. A copper strip is ideal because it offers a low impedance path to the input and it has a large surface area with which to contact the AUT input.

Ideally the capacitor used in the calibration block should be chosen from any standard 5% tolerance types, and it should be physically small. In this work we used surface mount capacitors on a small section of PCB. The PCB could be removed and replaced with another, thus allowing us to run tests with many different values of capacitance. A very easy way to test the performance of a calibration block is to measure the RF loss through it and compare

the result to the theoretical loss which is given in Section 2.3. During this work the measured loss of the NPL calibration block was typically within 0.2 dB of the theoretical loss.

Some designs of calibration block include a T-network of matching resistors which are designed to minimise mismatch errors during calibration. When using a 50 Ω source and receiver the potential mismatch during ECSM can be large because the input to the AUT may present a large impedance. Mismatch errors are also a function of source/receiver match and we have found that, with modern well matched analysers, there is no significant advantage in having a matching network. If a matching network is included then an additional correction is required in the final ECSM calculation to adjust for whatever voltage divider loss is caused by the T-network. When constructing the T-network it is important to keep all the wires orthogonal to each other as much as possible in order to minimise mutual coupling.

3.2 Sensitivity of ECSM to variation in its main parameters

The first parameter to study is the value of capacitance used in the calibration block. With the NPL block it was possible to measure one AUT with a range of capacitance values, and the unprocessed data is shown in Figure 3-3. From this we can see that a (large) change of 9 pF produces a difference of only 3 dB in output, and thus we can say that the measurement is relatively insensitive to variation in capacitance. This does not mean that choosing the right value of capacitance is unimportant, only that the 5% tolerance in value will cause minimal measurement error. As a rough guide, this graph suggests that there is about 0.3 dB of error in the final antenna factor per 1 pF variation in capacitance.

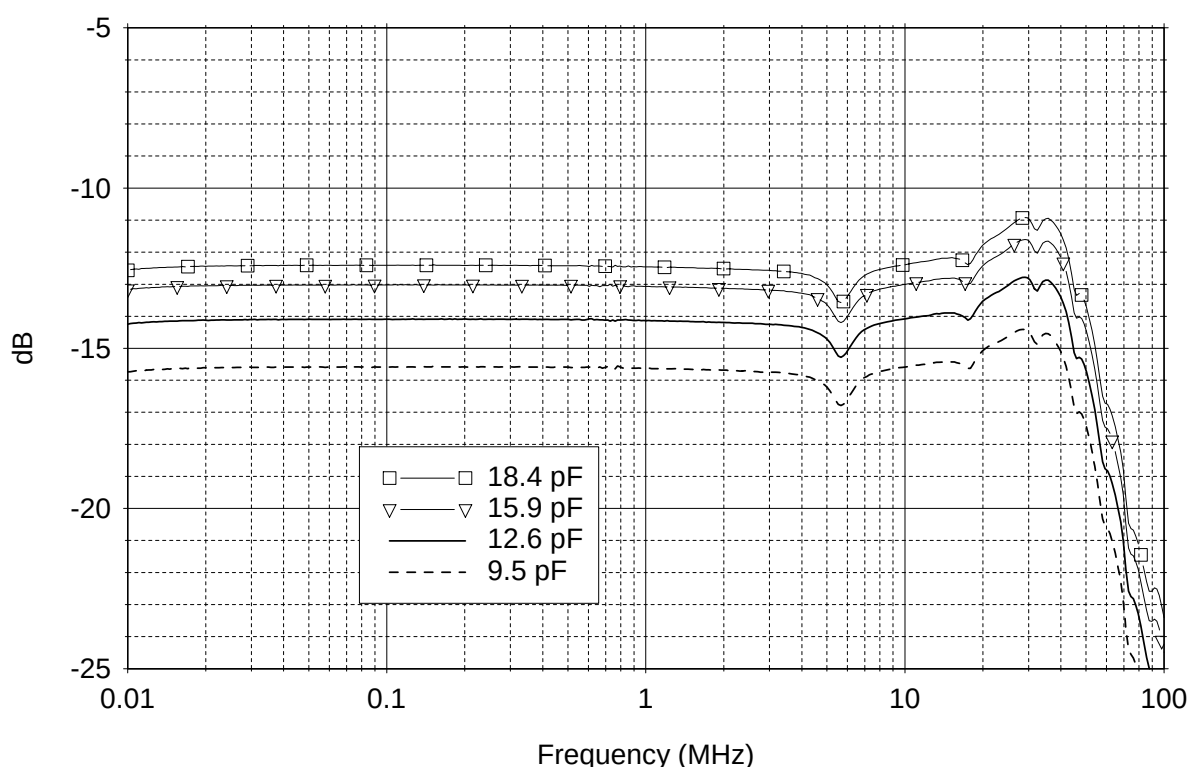


Figure 3-3: Unprocessed ECSM data taken with range of capacitance (Eaton 94607-1).

When calculating AF from an ECSM calibration there is an adjustment of $20 \cdot \log_{10}(h_e)$ to be made, where h_e is the effective height of the element which is given as a function of the actual element height. Because this is a logarithmic term it is relatively sensitive to which exact value of h_e is chosen. For example, many standards assume the effective height of a 41 inch element is 0.5 m, whereas 41 inches is 1.04 m so the effective height is 0.52 m. The 2 cm difference might not seem large, but when the logarithm is taken the value changes from 6.0 dB to 5.7 dB, which represents a 0.3 dB error due to a fairly small change in effective height.

In general, when the element is screwed into an AUT base, it might be difficult for the user to determine what value of actual height to use. One reason for this is that some pre-amps have raised platforms of dielectric supporting the base of the element, and within this material the antenna may have an extra bit of active element length which is not part of the removable element. In these situations a best estimate of height is normally sufficient if the target uncertainty for the complete ECSM calibration is ± 2 dB or greater ($k=2$), but an estimate may not be sufficient if lower uncertainties are required.

Confusion can arise when a figure of 6 dB adjustment is quoted in literature with no explanation of origin, which is a potential cause of error if the reader were unaware that this figure is related to element height and the AUT in question did not have a 1 m element.

Related to this issue is the difference in monopole performance when the grounding configuration is changed. Proprietary monopole designs usually have a casing at the base of the element which houses the associated circuitry, or pre-amp, and the height of this base can vary between 50 mm and 150 mm. Assuming we have an ideal vertical plane wave over a ground plane, we can simply place the monopole on the ground with its pre-amp casing earthed to the ground plane. Because of the presence of the pre-amp box, the feed point of the active element is raised above the ground. As a comparison we could cut a hole in the ground plane so that the base may be placed below ground level, and the pre-amp casing at the foot of the element could then be directly bonded to the surrounding ground plane. In the second case the E-field sees an effective height which is exactly calculated from the height of the element above the ground plane, but in the first case the feed point is raised so we cannot be so sure of the calculation.

We carried out some numerical modelling in NEC to see the effect of raising the feed point. We found that raising a simple passive element 75 mm above ground, while still maintaining a wire to ground, decreased its AF by around 1 dB (see Figure 3-4), which is equivalent to an increase in output signal. In the GTEM we could raise an actual passive antenna on conducting blocks, which simulate a pre-amp base, and the result is shown in Figure 3-5. The measurements show that a raise of 75 mm increases the output of the antenna by about 1.2 dB, which corresponds to an equal drop in antenna factor. These two results show good agreement, and suggest that raising the feed point will lower the antenna factor.

Based on this analysis and experimental work we can see that the GTEM free-field calibration technique (Ref [3]) effectively measures the AUT antenna factor when the base is placed on a ground plane and the element feed point is raised. This also suggests that an accurate ECSM will measure the other case, in which the ground plane reference is at the base of the element. Both measurements are valid, it is just a case of choosing which is suitable for any particular application. If an ECSM was performed, and one needed data equivalent to the GTEM method, then a reasonable approximation to the change in antenna factor may be found by re-calculating the effective height correction, but adding half the height of the base to the previously calculated effective height.

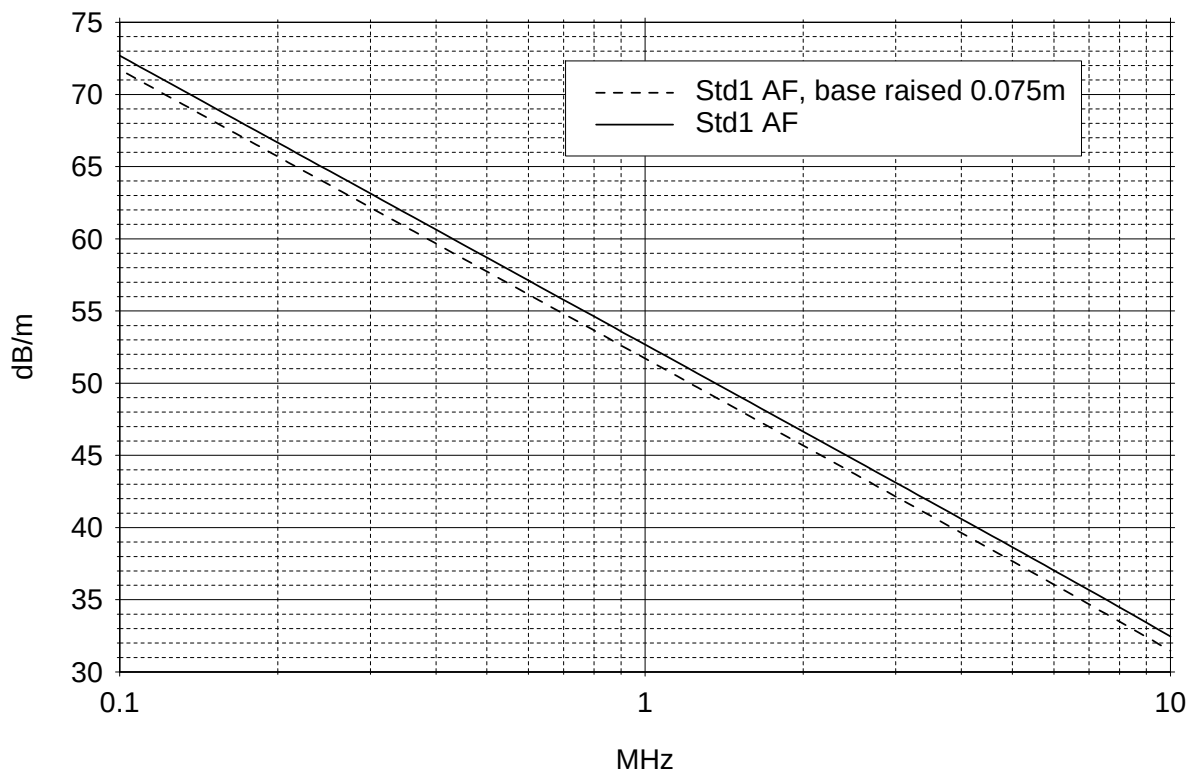


Figure 3-4: Change in AF caused by raising feed point of passive standard.

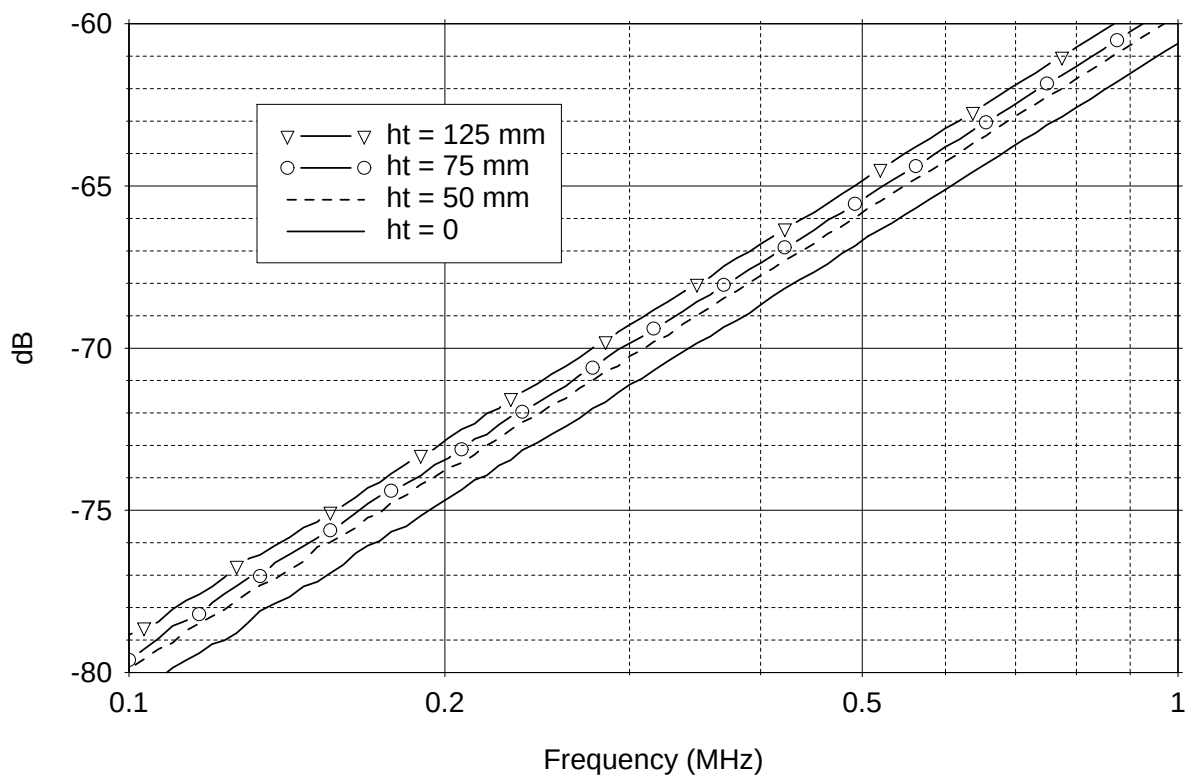


Figure 3-5: Change in measured output signal (in GTEM) when passive standard raised on conducting blocks.

4 SUMMARY OF MEASUREMENTS ON DIFFERENT TYPES OF MONOPOLE

4.1 Passive antenna (1.35 m)

For a simple passive element, it is possible to derive antenna factor in three ways. First, we can measure it in the GTEM as a normal free-space calibration; second, we can measure the equivalent capacitor value directly on the bench (while it is mounted in the calibration block) and convert this to antenna factor just as in a ECSM calibration; and lastly we can calculate a purely theoretical antenna factor from the derived response of the capacitor value (see Section 2.3). All three methods were applied to a 1.35 m passive antenna and the results are presented in Figure 4-1.

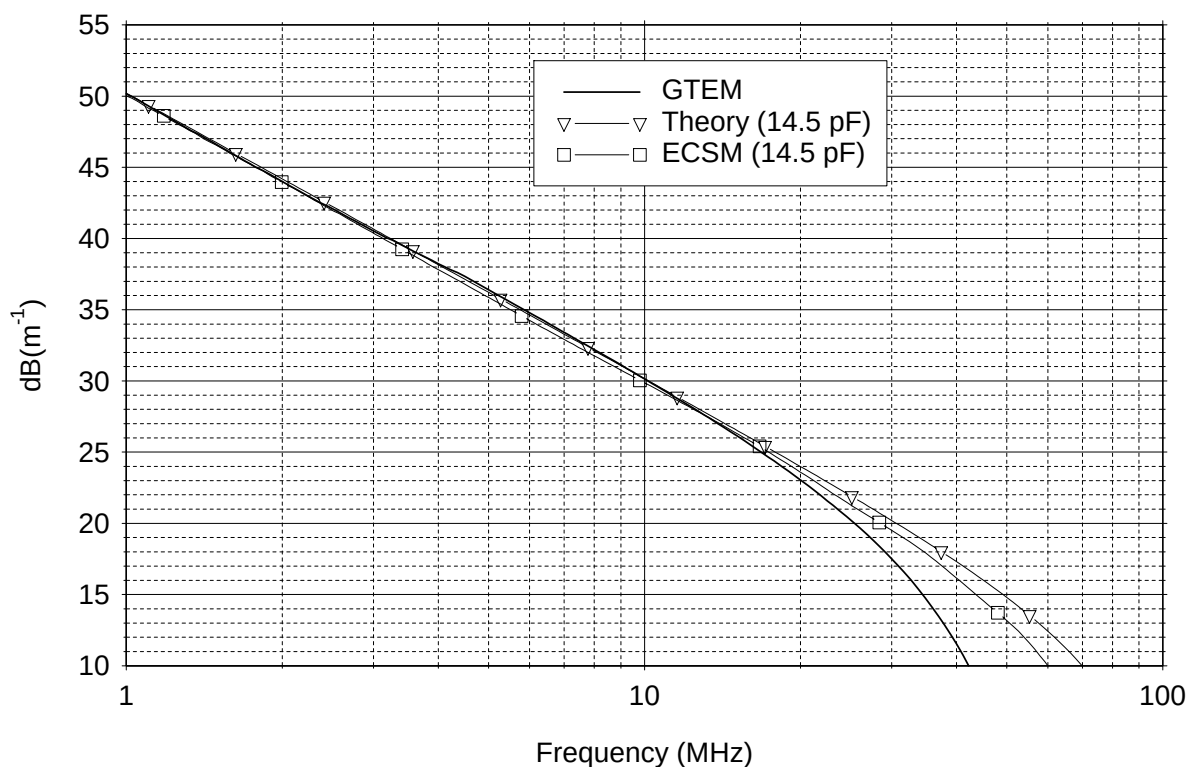


Figure 4-1: Comparison of AF for 1.35 m passive standard.

Since there is no base with this antenna, the GTEM method measures the same AF as with the ECSM measurement. Up to 15 MHz the theoretical value also agrees to within about 0.3 dB which provides some assurance that both practical measurement methods are sound.

The resonant frequency of a 1.35 m element is around 55 MHz, and in this region the element impedance will no longer be a mostly reactive value which is the assumption made for the ECSM and also for the theory AF. The graph shows that both the ECSM and theory

deviate from the GTEM AF as the frequency approaches resonance. Note that the GTEM calibration method in an MEB1750 cell has higher uncertainties above 30 MHz, but it does show the correct trend in AF (i.e. a fall off in value near resonance), which is confirmed by plane wave calibration on a large open site ground plane. The ECSM and Theory do not predict this because they are based on an assumption that the antenna length is less than $\lambda/8$. This assumption results in a 2 dB difference between ECSM and the GTEM method at 30 MHz. Ref [1] states that the formulae for ECSM are valid for monopoles shorter than $\lambda/8$, but perhaps this should be $\lambda/10$ if uncertainties of less than ± 1 dB are required.

4.2 Eaton 94607-1

This antenna has a 41 inch element which was substituted with a 12.6 pF capacitor. Figure 4-2 shows that the ECSM antenna factor is about 1.3 dB higher than the GTEM value, which we suspect is due to the raised feed point of the antenna when it is measured in the GTEM.

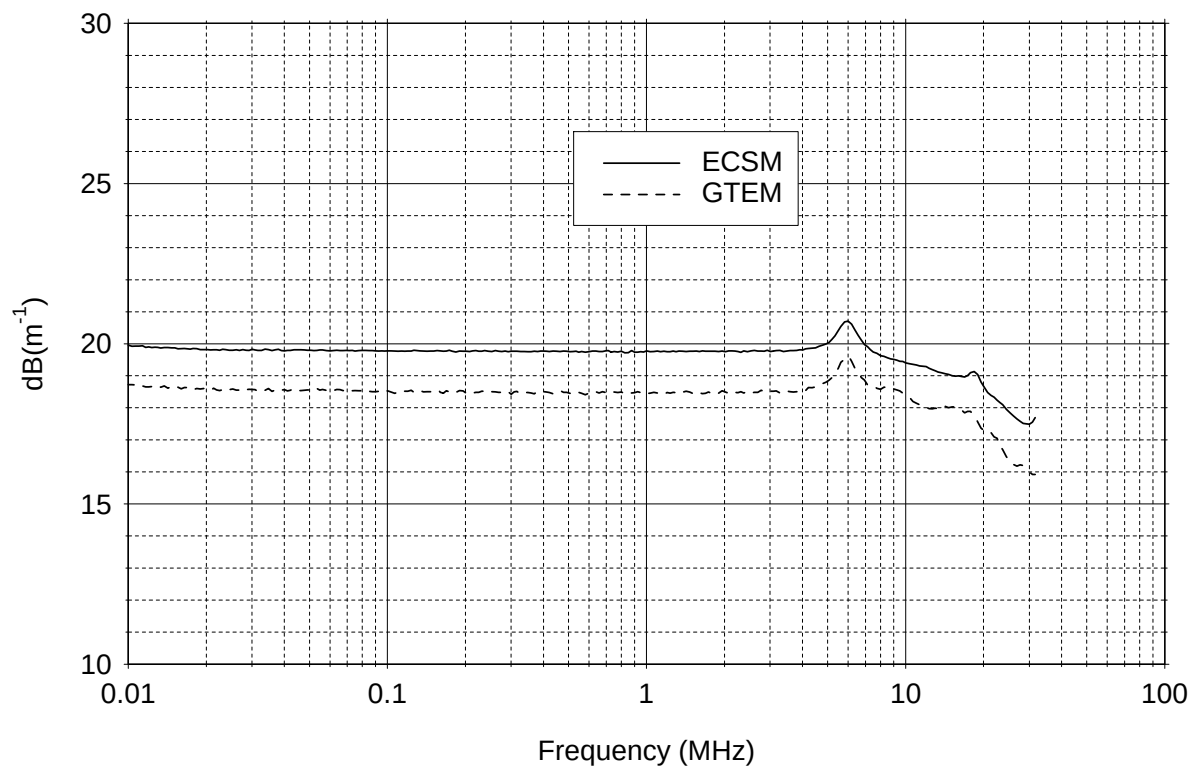


Figure 4-2: Measured AF for Eaton 94607-1

4.3 EMCO 3301B

Figure 4-3 shows the results for an EMCO 3301B antenna. Above 100 kHz the ECSM is higher than the GTEM value, as we expected from previous data, however the two methods tend to agree at low frequencies (below 10 kHz). At this moment there is no explanation for this result.

For comparison we also measured the antenna with 10 pF, which is the 'incorrect' value predicted by the ANSI expression for capacitance. The result is about 2 dB higher than the GTEM value right across the frequency range.

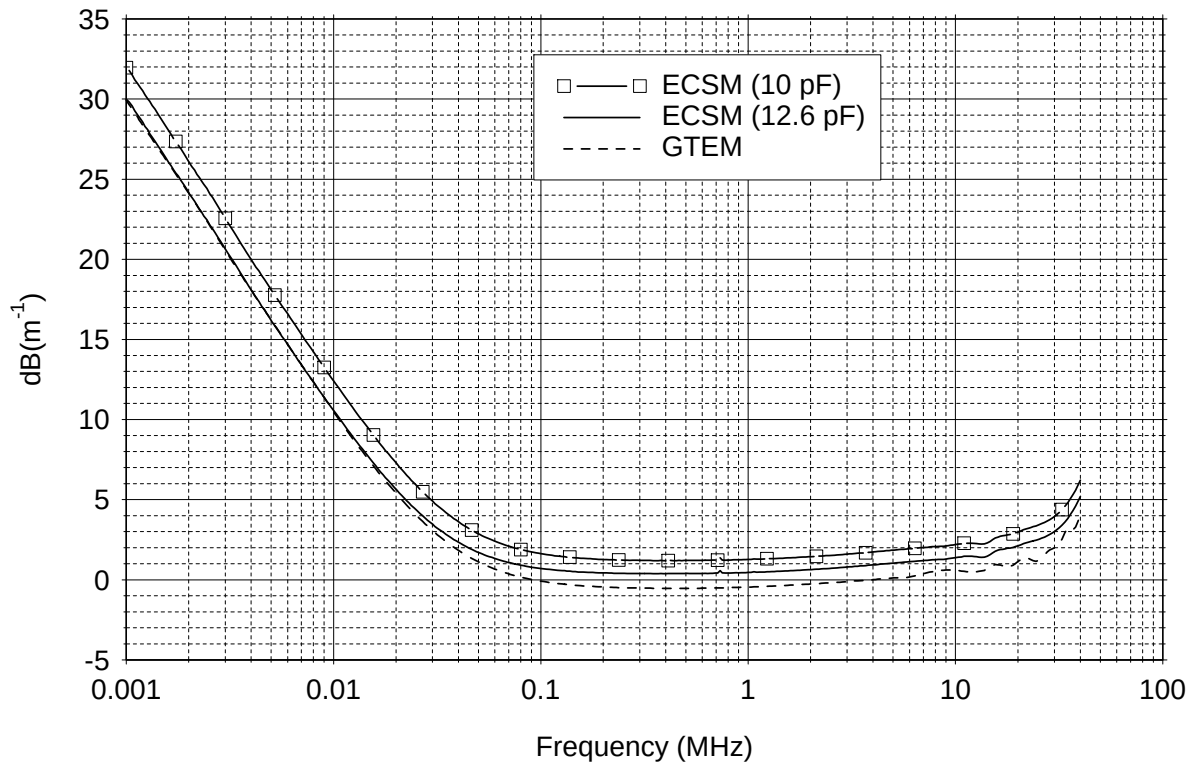


Figure 4-3: Measured AF for EMCO 3301B

4.4 EMCO 3303

The data in Figure 4-4 illustrates the same difference between calibration methods as we have seen before, except this time the difference is larger, at between 1.5 dB and 2 dB. This model of antenna has one of the tallest base units (at about 150 mm), so it is not surprising that it produces the biggest difference between the two methods.

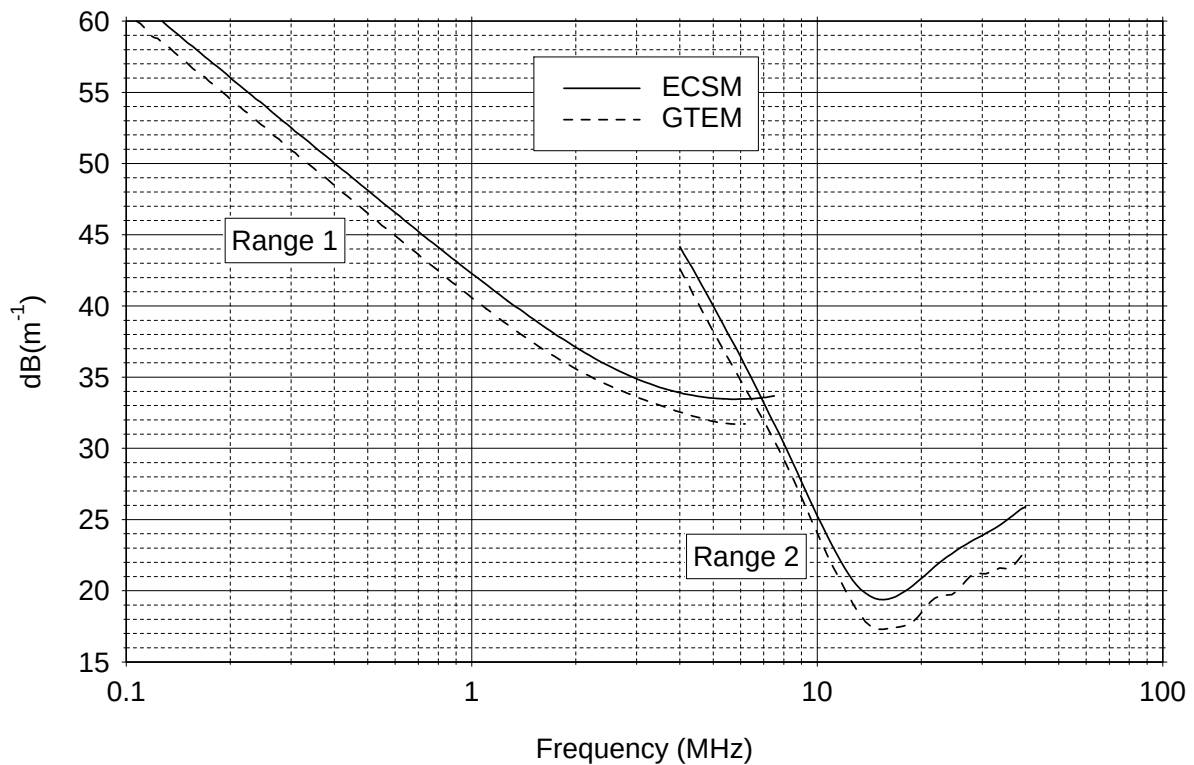


Figure 4-4: Measured AF for EMCO 3303

4.5 Rohde & Schwarz HFH2-Z1

This antenna was measured with several different element sizes; with dimensions ranging from a fatter 1 m element up to a thin 1.35 m one. The idea here was to test the accuracy of the calibration methods when they are used for extreme cases. All three graphs illustrate very similar differences between the two methods, which reinforces the idea that the difference in antenna factor is due to the height of the base, which was about 75 mm for this antenna. In these graphs the parameter a is element radius.

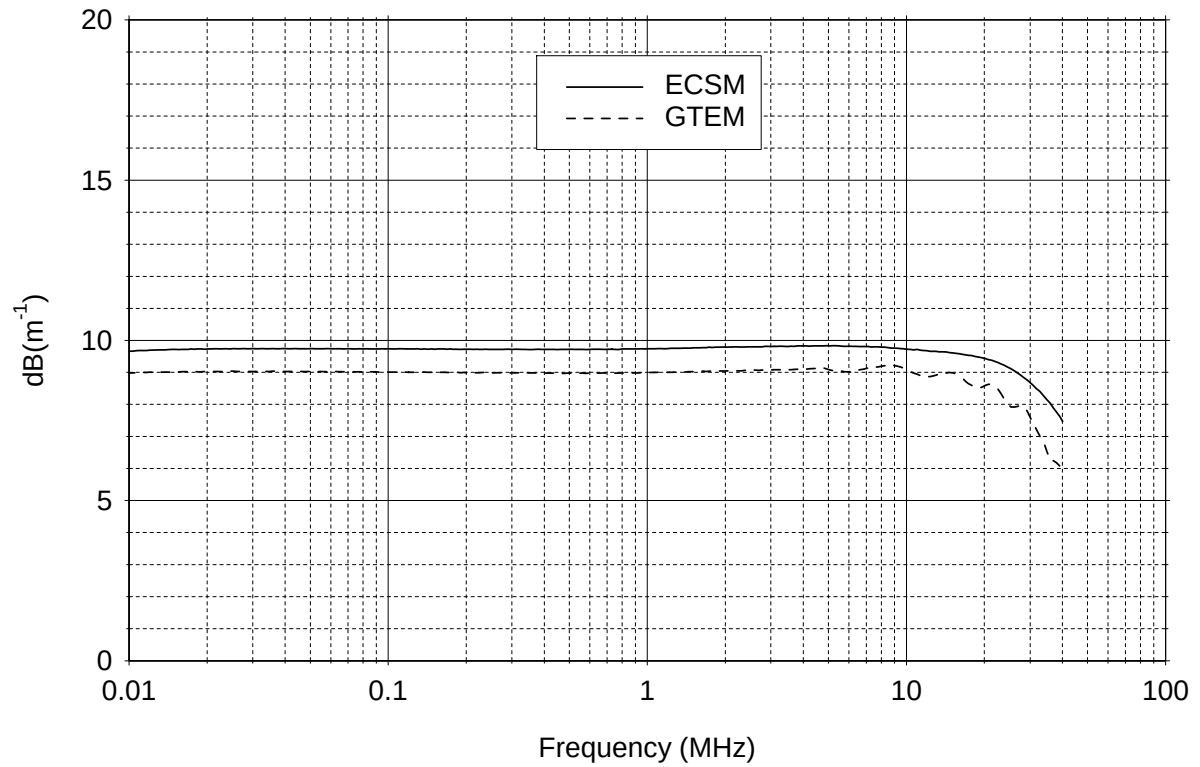


Figure 4-5: AF for element: $h = 0.96$ m, $2a = 0.0254$ m (1 inch)

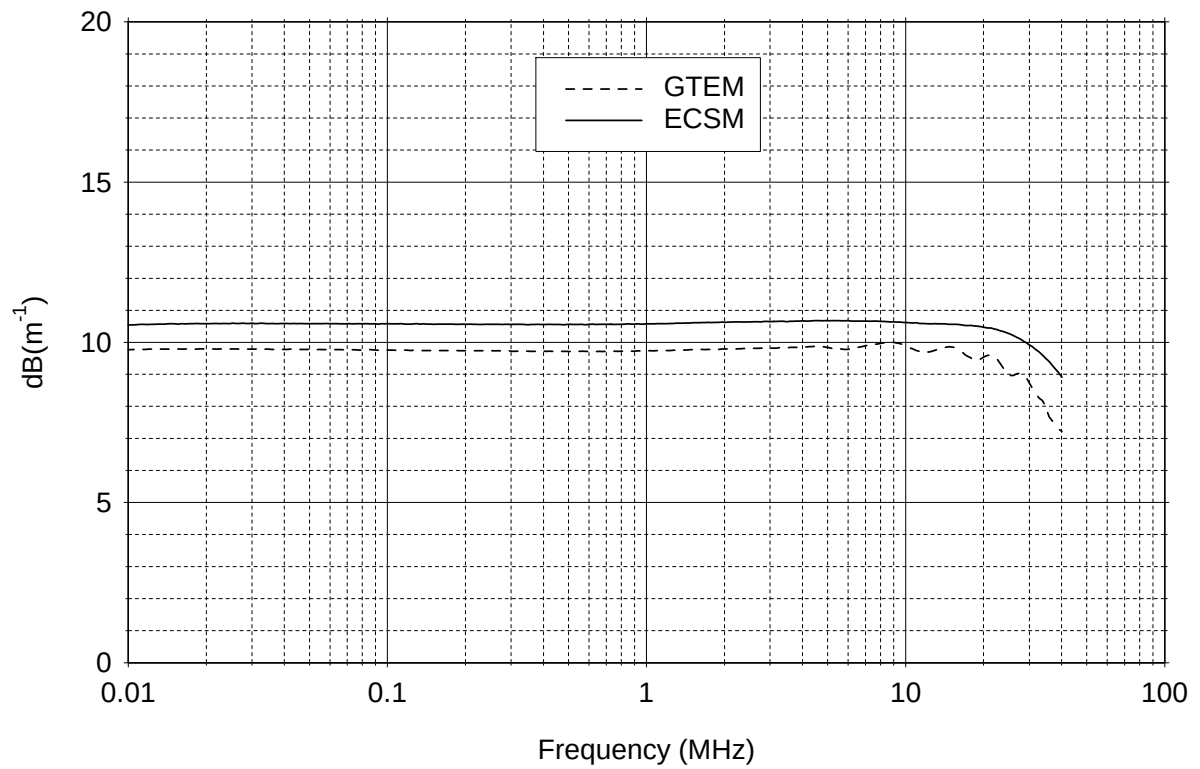


Figure 4-6: AF for element: $h = 1.0$ m, $2a = 0.005$ m

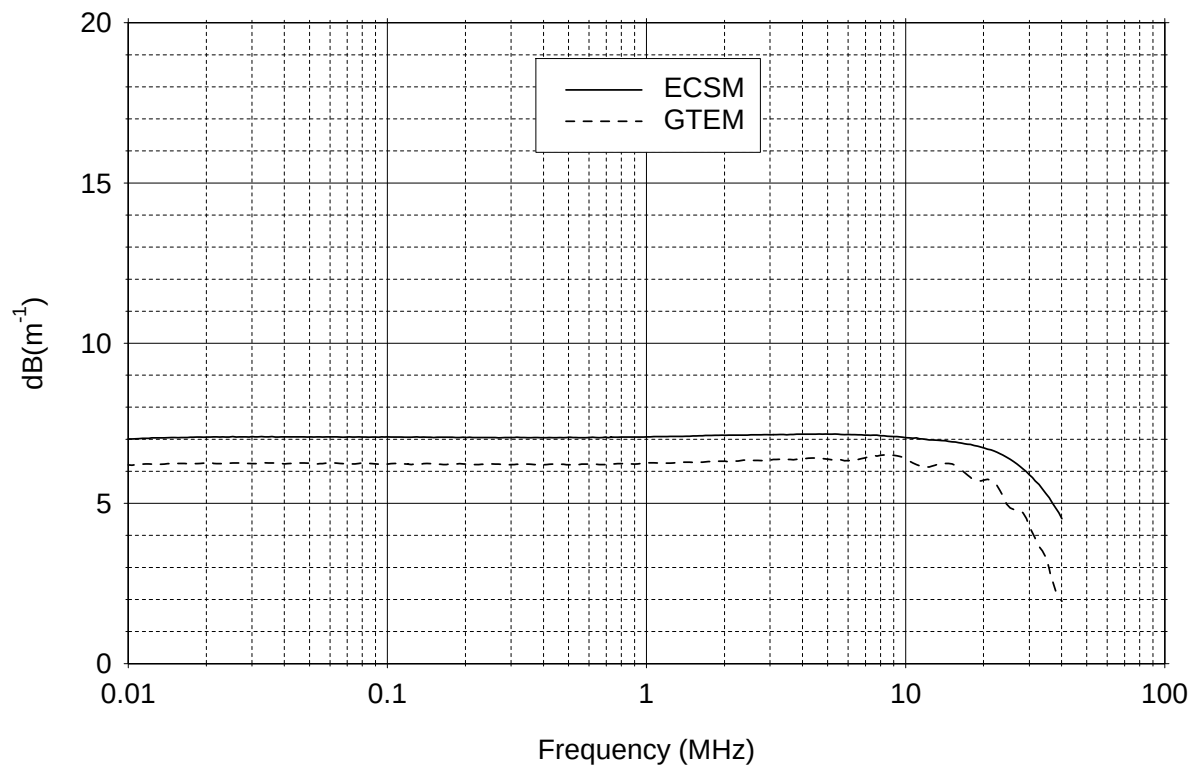


Figure 4-7: AF for element: $h = 1.35$ m, $2a = 0.005$ m

4.6 Singer 95010-1

This model of antenna has two configurations: one is a standard 41 inch element, and the other is a top loaded 50 inch element. For the ECSM calibration the top loaded element was treated as described in Section 2.3. The differences illustrated in Figure 4-8 are similar to those observed for other antennas but a little bigger because this antenna has a slightly larger base unit than typical.

The graphs show that, for the 50 inch configuration, there is an increasing level of difference between the two calibration methods above 15 MHz. The reason for this deviation is that the 50 inch element boosts its effective height with top-loading and thus lowers the resonant frequency. We know already that near the resonant frequency we expect the two calibration methods to disagree. Note that the GTEM data here has been joined with some Open Field Site data which gives accurate antenna factor right up to 100 MHz, and this allows us to measure antenna factor around the resonant frequency.

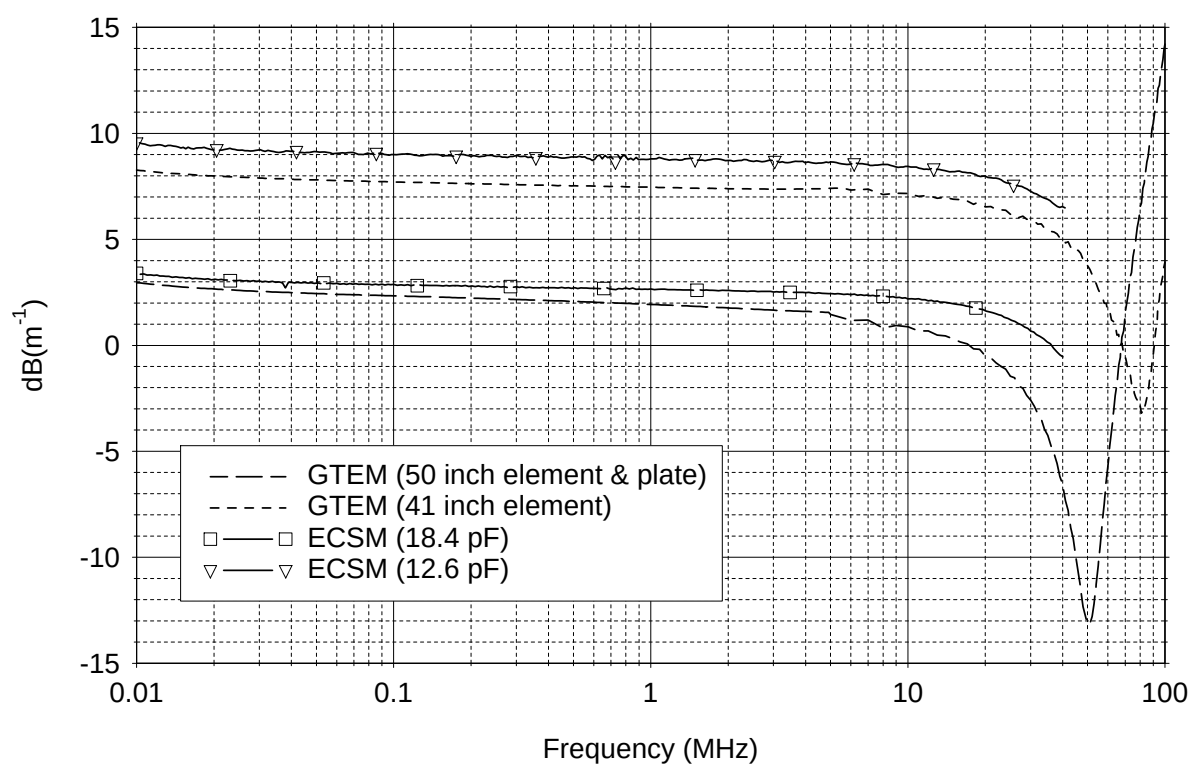


Figure 4-8: Measured AF for Singer 95010-1

5 CONCLUSIONS

We have investigated the origins and application of the Equivalent Capacitance Substitution Method (ECSM). As part of this we have derived the correct expression to calculate the equivalent capacitance for an element of given dimensions. In doing so we have resolved a contradiction between two different forms of the expression which one finds in the international standards. Also, some advice is given on how to construct an adaptor block for the ECSM, in particular minimising stray capacitances, with the aim of reducing uncertainty in the measurement.

In order to verify the calibration methods we have demonstrated that the calibration of a simple 1.35 m passive monopole, by both ECSM and NPL's GTEM method, agrees to within 0.3 dB with a purely theoretical antenna factor which is derived using the value of the equivalent capacitance. We have explored the sensitivity of the ECSM to variations in two of the most important parameters: namely, effective height and value of capacitance. We have illustrated the limitations of the ECSM when the monopole element height becomes a significant fraction of wavelength ($>\lambda/8$), and the assumption of purely reactive impedance is no longer valid.

Finally, we have measured a range of different monopoles and demonstrated that there seems to be a consistent offset between the ECSM and GTEM method when applied to a commercial antenna which has a pre-amp base unit. The GTEM calibration method produces a lower AF than the ECSM. Our work suggests that the cause of this offset is the apparent increase in effective height of the element when its feed point is raised up (i.e. on top of the pre-amp unit).

The GTEM method measures the whole antenna as it rests on the floor of the GTEM cell, which corresponds closely with the principal method of calibrating a monopole in plane wave conditions on a large open site ground plane. Thus, the GTEM method will generate antenna factor which is appropriate for those test scenarios in which the monopole pre-amp base unit is placed on top of the ground plane. In contrast, the ECSM simulates the idealised situation in which the zero potential ground reference is at the level of the feed of the element (however large the pre-amp unit actually is). For the case of commercial monopoles the observed difference between the ECSM and GTEM method ranged from 0.8 dB to 2 dB, depending on the size of pre-amp unit. There are some designs of monopole with a ground plane which may be attached to the top of the pre-amp base, and this is then bonded to the ground plane of the test configuration, so in these situations an accurate ECSM calibration will generate a more appropriate antenna factor. However, in the context of typical EMC measurement uncertainty, the differences described here are usually not significant. If lower uncertainty were required, it is possible to derive an approximate conversion factor

(Section 3.2) which allows a calibration by either the ECSM or the GTEM method to be adjusted to suit the grounding configuration of the particular test environment.

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