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A High Performance Active Antenna
for the High Frequency Band

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ABSTRACT

The design of an active antenna with an operational frequency range from low- to high-frequency is presented. The main cause of inter-modulation distortion is identified and recommendations are given to minimize their generation. A detailed analysis is carried out into the various sources of noise generated within the active antenna and how their summation affects the total noise seen at the output. Lightning protection is also discussed.

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A High Performance Active Antenna for the High Frequency Band

Executive Summary

This report describes the design of a high performance high frequency (HF) active antenna starting from the antenna rod and finishing with the output buffer stage which drives the co-axial transmission line feeding a receiver. The main source of inter-modulation products is identified and steps that can be taken to reduce their magnitudes are described and implemented. Internal noise generated within the design, which can limit the antenna's sensitivity, is also examined in detail and the main source of this noise identified across the antenna's usable frequency range. Recommendations are given on how to install these types of antennas in an antenna site with the resonance formed by the mast together with the active antenna readily calculated. Operating an active antenna at frequencies higher than this resonance will result in a monotonic decrease in sensitivity.

The completed design is a compact receive only active antenna covering the frequency range from LF to HF. It is broadband with a constant antenna factor over this frequency range. It can be used to measure the vertical E field strength of a signal, perform general surveillance work (either fixed or vehicle mounted) or used to create a quickly deployable antenna array for the geo-location of signals in the HF band.

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Wayne Martinsen was indentured as an apprentice communications technician in 1972 with Transport Communications Pty. Ltd. (Buranda, Qld.). He attended Yeronga Technical College (Brisbane, Qld.) where he received honours in the majority of his subjects. In 1975 his apprenticeship indentures were transferred to The Radio Centre (Archerfield Aerodrome, Qld.) where he was employed as an Air Maintenance Engineer, Category Radio. He was involved in all aspects of repair and maintenance of avionics systems associated with the light aircraft industry. Upon completing his training as an apprentice he sat for and passed various Dept. of Transport Licensed Air Maintenance Engineer (L.A.M.E.) exams. In 1977 he joined the R.A.A.F. and after graduating from their School of Radio, where he received training in military systems, he worked on a wide variety of airborne communications, navigation and anti-submarine warfare systems. In 1987 he was discharged from the R.A.A.F. and took up the position of Technical Officer (Engineering) at Defence Science and Technology Group. Here he is tasked with the design and development of R.F. circuits required for tropospheric and ionospheric research.

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Contents

1.   Introduction

Broadband vertical monopole type antennas used for omni-directional surveillance, high frequency (HF) site noise measurement and direction finding (DF) arrays are generally made of long metallic rods or tubes. Such antennas are intrinsically quiet. It is only their distributed DC resistance, usually in the order of a few ohms, which generates noise. This noise is small when compared to the noise generated by the receiver's internal resistive termination, typically 50 ohms. The majority of HF noise seen at the terminals of these types of receiving antenna is that which has been induced into the antenna from noise energy propagating in free space.[1] The 377 ohms characteristic impedance of free space, which is usually thought of as being made up of distributed inductance and capacitance, cannot by itself generate noise. Pure reactance does not generate noise. Noise propagating in free space is due to a noise source coupling its energy into free space via some rudimentary antenna.

The noise voltage induced into a vertical monopole HF antenna suitably matched to a receiver by the noise propagating in free space will generally far exceed the noise generated by the DC ohmic resistance of the metal of the antenna itself and is usually also greater than the internal noise generated within most modern HF receivers. As a consequence this external HF noise is usually the factor which limits the sensitivity of a HF receiving system.

A problem with these antennas when used in a DF array is the mutual coupling between antennas which induces bearing errors in the direction of arrival of signals. It is their physical height, some six to ten metres high, with typical antenna spacing varying from seven to seventeen metres which makes these antennas so easy to couple energy into one another. A solution to the mutual coupling problem is to use an antenna that has a short, thin metallic rod terminated into a high impedance. A short thin rod minimises the antenna profile thus reducing the mutual coupling between antennas. The high terminating impedance reduces the induced current flowing between the rod and ground system minimizing antenna re-radiation which again reduces the mutual coupling between antennas. All that is needed is a transconductance amplifier to convert the induced voltage appearing across the high impedance load into a signal capable of driving a 50 ohm transmission line. What has just been described is more commonly known as an active antenna.

Active antennas are of interest to anyone who wants a compact receiving antenna for the broadband reception of signals from long waves to approximately 50 MHz. They are ideal for general omni-directional surveillance work or can be used to measure the *E* field strength of a signal. Their lack of pre-selection filtering makes these antennas susceptible to the generation of intermodulation products by strong signals appearing anywhere

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[1] The exception to this are antennas which have a very low radiation resistance where this resistance is close to or less than the DC resistance of its metallic structure and the receiver's input impedance has been matched to this low radiation resistance. These antennas generally have physical dimensions that are very small when compared to the operating wavelength.

within their operational frequency range. The causes of the intermodulation products and steps that can be taken to minimize their generation will be addressed in the design.

A problem with a short rod as used with active antennas is the resultant induced voltage from a site's background noise may not be sufficient to overcome the internal noise generated by the active antenna's transconductance amplifier circuitry. This would make a receive system using this type of antenna internally noise limited. It is for this reason that active antennas are not recommended for site noise measurements unless it can be proven the site to be measured is externally noise limited throughout the range of frequencies to be measured under all seasonal and sunspot cycle conditions.

2.   Active Antenna Design

There are many designs for active antennas that can be found in literature and chief amongst the designers concerns is the spurious products generated by strong RF signals; high power AM broadcast stations in the AM band (530 KHz to 1600 KHz) are the main causes of intermodulation products. Most of these designs are based on the U310 JFET. This JFET is well known for its performance in the design of active antennas and is the active component of choice for this design. Due to the many variations in which the U310 has been used in this role and the lack of adequate documentation clearly defining performance, it is not clear which circuit configuration produces the best result. A more detailed analysis than can be found in literature is therefore required.

2.1   Design Criteria

The aim is to design an active antenna that has:

Minimum frequency range of 2 – 30 MHz: This frequency range should easily be met. The only expected frequency limitation is the -6dB/order/octave of the reactive components (inductors and capacitors) used and the gain-bandwidth product of the active devices.

Circuitry with the lowest practicable internally generated noise: Every active semiconductor device used in a circuit is a source of excess noise. This excess noise is one of the factors which limit the sensitivity of a system. Their use must therefore be kept to a minimum. Carbon based resistors for example are known for their large excess noise and will not be used. Metal film resistors with their very low noise levels are the preferred type. Some ceramic low K SMD and lead type capacitors have been found to exhibit popcorn type noise. These types of capacitors will not be used in the through path of the signal. Silver mica capacitors have extremely low effective series resistance and extremely high shunt resistance. As a consequence, they have very low levels of internally generated noise. These capacitors will be used in the noise critical area of the U310 gate circuitry.

Highest possible second and third order intercept points: Active devices are the main limitation in this area. They must be selected for their linearity as well as being used appropriately. Some form of push-pull output stage will be used to help suppress the even order distortion products.

Broadband 50 ohm output impedance: The output impedance of some active antenna designs found in literature is of the order of a few ohms. This gives an approximate 6dB increase in output signal level when used in a 50 ohm system, which is probably its main attraction. The input impedance of some receivers is not necessarily a broadband 50 ohms. Some receivers have tracking pre-selection filters which only present a nominal 50 ohm termination to the passband signals. Out-of-band signals are reflected back to the active antenna via the 50 ohm co-ax where they are reflected

back to the receiver by the few ohms output impedance of the active antenna. Multiple reflections of unwanted signals can cause degradation of the intercept points of both the active antenna and the receiver in such a scenario. This can be minimised by making the output impedance of the active antenna a broadband 50 ohms thereby absorbing and dissipating as heat any reflected signals from the input of the receiver. A more detailed discussion is given in section 2.3.3.

Smallest antenna factor (k) without resorting to the use of a voltage gain amplifier.
The antenna factor k is the E field strength in V/m surrounding the antenna divided by the antenna's terminated output voltage.

$$k = \frac{E}{V_{out}} \qquad \text{and expressed in decibels} \qquad K = 20\log k \qquad (2.1)$$

where: k = antenna factor

K = antenna factor expressed in dB

E = strength of E field (V/m)

V_{out} = terminated antenna output voltage (V)

HF active antennas are generally very broadband devices covering a frequency range from LF to low band VHF frequencies (approx. 30 KHz to 50 MHz). Any electronic amplifier used in an active antenna must be able to handle the vector sum of all the signals residing in this frequency range without creating distortion products. The two factors that affect the magnitude of V_{out} in equation (2.1) for a given E field strength are:

1. the length of rod used to sample the E field and

2. the electronic gain applied to the magnitude of voltage induced into the length of rod.

Both 1 and 2 above have a directly proportional relationship with V_{out}. Either doubling the rod length or the electronic gain will double V_{out}. It was highlighted in the introduction that passive antennas are generally intrinsically low noise devices. This implies that antenna gain is preferred over electronic gain in communication systems, especially in antenna sites with very low external noise. For example, a passive antenna's gain is not characterised by the 1dB compression point, intercept points or spurious free dynamic range. The magnitude of noise appearing at the output of an electronic amplifier is dependent upon its active devices internally generated noise (noise factor), bandwidth, gain and its distortion products described via the amplifier's 1dB compression point, intercept points and spurious free dynamic range. In the case of an active antenna, a balance is needed between antenna gain (rod length) and electronic gain of the transconductance amplifier used to convert the voltage at its high impedance input into a current at its low impedance output. Too much rod length in the presence of strong E fields can overdrive the electronic amplifier and generate unwanted distortion products. Too much electronic gain has a

similar effect. Preference will be given to the intrinsically noise free and distortion free gain obtainable by increasing the rod length with the amplifier being designed to maximise its spurious free dynamic range at the expense of electronic gain.

2.2   The Input stage

There is a problem with the local bird wildlife on the antenna test site. Sulphur Crested Cockatoos and Galahs are quite at home hanging onto very thin structures while destructively pecking at anything that looks interesting to them. Long thin vertical rods used in the past have been bent and in some cases snapped off by these birds either skylarking in the test site or through crashing into them whilst in full flight. A 1.2m long tapered stainless steel antenna rod designed for VHF mobile communications was purchased to combat this menace. This tapered rod (2.45 mm at the bottom to 1.25 mm at its tip) can have its ends brought together to form a circle and, when one end is let go, springs back to its original straight shape. This rod will be used for the antenna.

In order to understand the impact the rod will have on the design of the active antenna it is first necessary to define the rod's electrical equivalence. At frequencies below the rod's natural resonant frequency the rod's input impedance will appear as a small capacitor with its associated distributed inductance in series with a resistance, see figure 2.1(a).
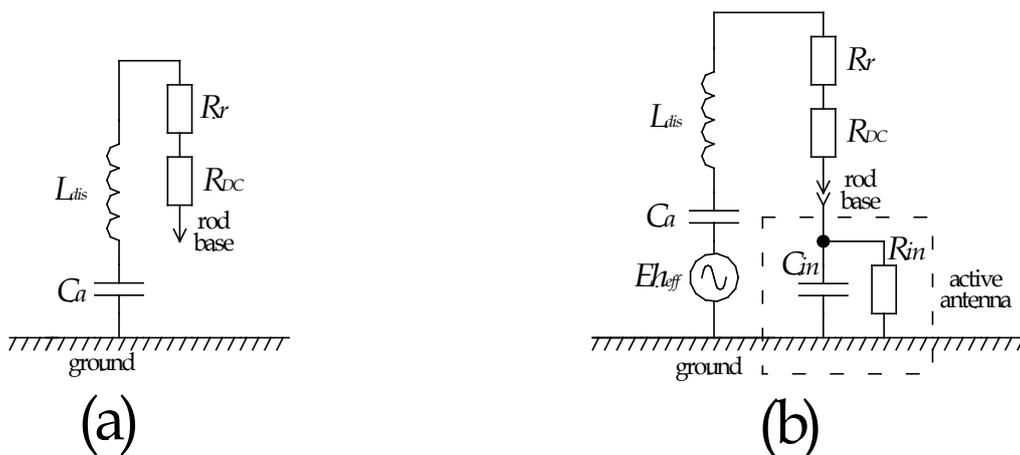
Figure 2.1   (a) schematic representation of a short vertical rod antenna above a ground plane and in (b) voltage generator created by a time varying E field and the effective height of the rod connected to the input circuitry of the active antenna shown in the dashed box

The capacitance will be the true capacitance measured at some low frequency which will increase in value as the frequency is raised due to the distributed inductance associated with the length of the rod. This increase in value of capacitance with frequency is more commonly known as *apparent capacitance*. The apparent capacitance to ground of a short vertical rod over a perfectly conducting ground plane is given as [ref. 4, ch. 16 pg. 3 to 6]:

$$C_a = \frac{55.78h}{\left[\left(\ln \frac{h}{a}\right) - 1\right]\left[1 - \left(\frac{fh}{71.32}\right)^2\right]}$$

(2.2)

The resistance is the series combination of the radiation resistance R_r plus the DC ohmic resistance R_{DC} of the metal. The radiation resistance can be calculated by:

$$R_r = \frac{1.44\, h^2 f^2}{312}$$

(2.3)

where: R_r = radiation resistance (Ω)
h = height of rod (m)
f = frequency (MHz)
C_a = rod apparent capacitance to ground (pF)
a = rod diameter (m).

Equations (2.2) and (2.3) were used to calculate the expected values for the 1.2 m rod over the HF band, see figure 2.2. Calculating the impedance using $Z = \sqrt{(R_r + R_{DC})^2 + X_{Ca}^2}$ will show the impedance very closely follows the calculated reactance of the apparent capacitance to ground. The radiation resistance and DC resistance (R_{DC}) have very little impact over the frequency range of interest can therefore be ignored.

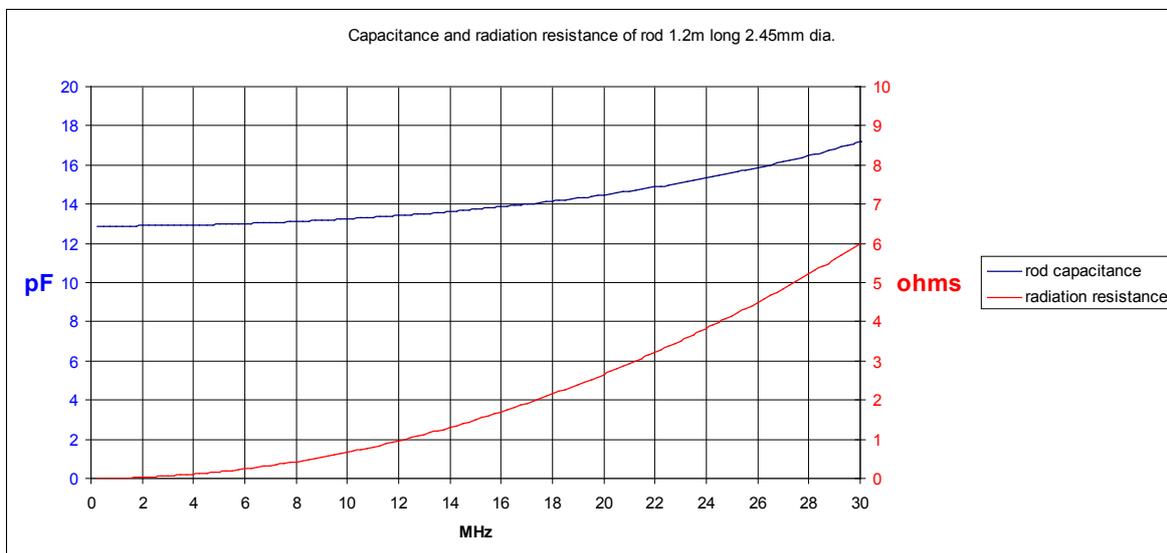*Figure 2.2   Capacitance (upper trace) and radiation resistance (lower trace) of a 1.2 m vertical rod 2.45 mm diameter above a perfectly conducting ground plane*

Figure 2.1(b) shows the voltage generator created by a passing E field and the rod's connection to the input of the active antenna. The total input circuit capacitance of the

active antenna C_{in}, and C_a form a capacitive divider sharing the output voltage from the generator in proportion to their reactance ratio. C_{in} must therefore be made as small as possible to maximize its reactance and hence the voltage appearing across its terminals. Some active antenna designs increase the value of C_a by fitting a capacitive top hat to the vertical rod thus increasing the voltage drop across C_{in}. The total input circuit resistance of the active antenna R_{in},(which is in shunt with C_{in}) is to be made as large as possible relative to the maximum reactance value of C_{in} for the same reason. C_{in} and R_{in} also have an effect on the LF (Low Frequency) RF noise generated by the active antenna's input circuit. This is covered in detail in section 3.

2.2.1   PCB Layout Capacitance

Care needs to be taken to arrange the PCB layout so as to minimise undesirable parasitic capacitance. FR4 for instance has an ε_r (relative dielectric constant) of approximately 4.7. Any PCB layout using this type of or simular substrate must use tracks that are short and very thin with a substrate thickness as thick as possible in an effort to reduce layout capacitance to ground. The preferred layout method, and that used by this author, is the "in air three dimensional point to point wiring on a copper ground plane", also known as the "dead bug" or "ugly bug" technique in some texts. It produces the lowest distributed capacitance to ground and hence minimises C_{in}. Only the active antenna's input circuitry between the antenna rod base and the input lead to the first active device need be wired in this fashion. The output of the first active device and subsequent circuitry can tolerate the few extra pF added by the PCB tracks due to their much lower circuit impedance.

2.2.2   Choosing the FET configuration

FETs with their high gate input impedance are the logical choice for use as the first active device. There are two possible design configurations that suit the requirement.

1. *common source*: The common source configuration is generally associated with voltage gain (A_v) which is approximated by:

$$A_v = \frac{R_L}{\dfrac{1}{g_{fs}} + R_S} \tag{2.4}$$

   where: R_L = total load resistance seen by the drain (Ω)
   R_S = un-bypassed FET source resistance (Ω)
   g_{fs} = common source forward transconductance (S)

   Any voltage gain in an RF circuit requires higher intercept points to keep unwanted distortion products of strong signals below the circuit noise floor. Higher intercept points generally indicate higher 1dB compression points and correspondently higher quiescent current. Higher current capability means larger FET channel dimensions and as a consequence larger internal distributed capacitances within the FET. C_{in} can now become large relative to C_a reducing

its voltage drop and sensitivity. Common source configuration will not be considered any further for this application.

2. *common drain*: The voltage gain (A_v) of a common drain (source follower) is given by [ref. 7, pg. 58]:

$$A_v = \frac{V_{out}}{V_{in}} = \frac{g_{fd}R_L}{1 + g_{fd}R_L}$$

(2.5)

where: g_{fd} = common drain forward transconductance (S)
                 R_L = total load resistance seen by the JFET source (Ω)

The voltage gain being less than unity eases the need for large channel dimensions in the FET giving rise to a smaller gate capacitance. Hence, a common drain configuration with its high input impedance and low output impedance will be used for the current design. Other equations that are useful in designing common drain amplifiers are:

$$Z_{in} = \frac{R_{in}X_{in}}{\sqrt{R_{in}^2 + X_{in}^2}} \qquad \theta = arctan\left(\frac{X_{in}}{R_{in}}\right)$$

(2.6)

$$R_o \approx \frac{1}{g_{fd}} \parallel R_s$$

(2.7)

$$I_d = g_{fd}V_g$$

(2.8)

where: Z_{in} = input impedance seen at the FET's gate terminal (Ω)
                 C_{in} = total input shunt capacitance of FET input circuitry (F)
                 X_{in} = total input shunt capacitive reactance of FET input circuitry (Ω)
                 R_{in} = total input circuit shunt resistance of FET (Ω)
                 f = frequency (Hz)
                 θ = phase angle (degrees)
                 R_o = output resistance seen at the FET source lead (Ω)
                 R_s = un-bypassed source resistance (Ω)
                 I_d = FET's drain current (A)
                 V_g = input signal voltage drop across R_{in} (V)
                 $\parallel$ = in parallel with

The corner frequency for a common drain amplifier is [ref. 2, pg. 113]:

$$f_{out} = \frac{1}{2\pi r_{ds} C_{out}} \quad = \quad \frac{g_{fd}}{2\pi C_{out}}$$

(2.9)

$$r_{ds} = \frac{1}{g_{fd}}$$

(2.10)

$$C_{out} = C_{gs} + C_{ds} + C_{load} + C_{stray}$$

(2.11)

where: r_{ds} = FET's internal resistance between source and drain leads (Ω)
C_{out} = total output capacitance (F)
C_{gs} = capacitance between the FET's gate and source leads (F)
C_{ds} = capacitance between the FET's drain and source leads (F)
C_{load} = capacitance of the load connected to the FET (F)
C_{stray} = stray circuit capacitance (F)

2.2.3 Minimising Intermodulation Distortion

All discrete solid-state electronic amplifying devices generate distortion products at their outputs when a signal is applied to their inputs. In the case of the JFET, the main cause of distortion is the variation in the forward transconductance when the drain current is changed. There are many factors that can cause a change in drain current. The two that have the greatest impact of the generation of distortion products are:

1. the application of an input signal which causes the current to swing about the bias point

2. the bias point varying through the effects of a large input signal.

It should be noted that due to the reverse bias of the gate/source junction and the very low leakage current associated with this reverse bias, the same current flows in the drain lead as in the source; therefore the common drain transconductance (g_{fd}) is the same as the common source transconductance (g_{fs}).

Figure 2.3, obtained from [ref. 8], is a plot showing the dependence of g_{fs} on drain current for the U310 JFET. Before this figure can be examined in detail it is first necessary to derive an approximate formula for the prediction of distortion products in active semiconductor devices.
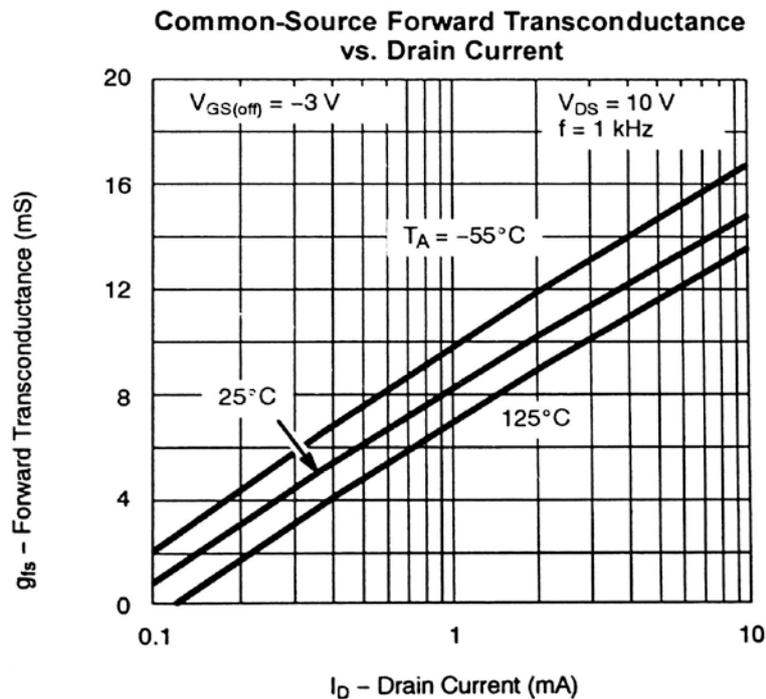
Common-Source Forward Transconductance vs. Drain Current

Figure 2.3   Variation in transconductance with I_D for the U310 JFET from [ref. 8]

2.2.4   Approximating the Output Third Order Intercept Point of a Class A Transistor Amplifier

The following derived formula can be used to approximate the output third order intercept point of a single ended class A transistor amplifier. It assumes a conjugate match between the transistor and load.

The product of V_{CE} and I_C gives the total power dissipated by an active device. Assuming 25% efficiency for a class A amplifier gives an expected output 1 dB compression of ($V_{CE} I_C$ /4). Typically an amplifier's output third order intercept point is located 10 to 15 dB above its 1 dB compression. Taking 13 dB as approximately midway in the range of 10 to 15 dB, the power multiplier for approximating the output third order intercept becomes

$anti \log_{10}\left(\dfrac{13}{10}\right) = 20$. Combining this with the expected 1 dB compression point gives

$10 \log_{10}\left(\dfrac{20 V_{CE} I_C}{4}\right)$ and finally, by specifying the current to be in mA and not amps

converts the result from dbW to dBm. The final equation is:

$$OPIp^3 \approx 10 \log_{10}\left(5 V_{CE} I_C\right)$$

(2.12)

where: $OPIp^3$ = output third order intercept point (dBm)
V_{CE} = voltage between collector (or drain) and emitter (or source) (V)
I_C = collector (or drain) current (mA)

Equation (2.12) shows the distortion in a transistor is minimised by allowing the transistor to draw as much current as possible with the highest V_{CE} (consistent with a safety margin from the manufacturer's absolute maximum ratings). The manufacturer's maximum power rating for the U310 is 500 mW. Operating the JFET at approximately 400 mW should give adequate head room making allowance for component tolerances and the U310 de-rating of 4mW/°C above 25°C. A drain voltage of +15V with a drain/source current of approximately 30 mA was chosen. The 1.17V bias required to draw 30 mA will be derived by the voltage drop across a 39 ohm resistor placed between the source lead and ground. Two series connected Radio Frequency Chokes (RFC) will be inserted in series with source resistor to prevent the 39 ohms from loading the wanted RF signals. Series connecting two chokes this way reduces the total associated distributed capacitance for a given inductance and reduces the ferrite losses. These losses are seen as a high value shunt resistor across each RFC and by connecting these two ferrite losses (high value shunt resistor) in series reduces the power lost through heating the ferrite thus increasing the overall combined inductor Q. This ensures the RFC's presents the highest impedance over the broadest frequency range. The expected power dissipated by the JFET is the drain/source current multiplied by the drain/source voltage, 0.03(15-1.17) = 415 mW. Measures will be taken to dissipate this heat to keep the JFET as cool as possible. The approximate $OPIp^3$ via (2.12) is $10Log_{10}(5*(15-1.17)*30)$ = +33 dBm.

The JFET was connected as seen in figure 2.4 (a) with a load resistor (R_L) of 50 ohms and the forward gain was measured to be -6.3dB at 8 MHz. The single RF generator was replaced with an RF two tone generator (7 MHz and 11 MHz) and the following output parameters measured, $OPIp^3$ = +30 dBm and $OPIp^2$ = +32 dBm.
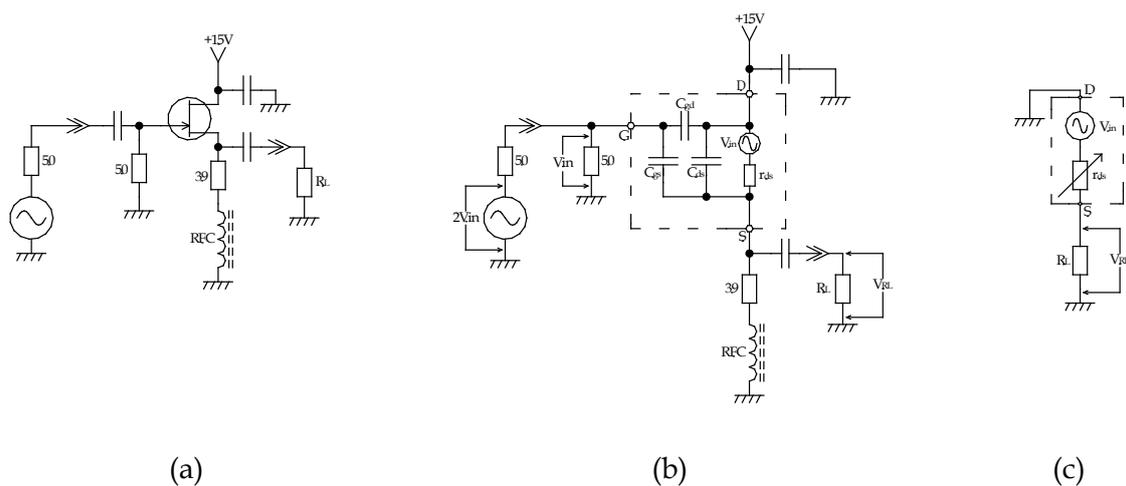

(a)   (b)   (c)

Figure 2.4   *(a) Test setup for initial measurements of the JFET, (b) Thevenin equivalent model of the FET, (c) Simplified circuit*

2.2.5 The Main Cause of Distortion Products

In order to take measures to improve these intercept points of the JFET, it is first necessary to understand the main causes of the JFET's internally generated distortion products. Figure 2.3, taken from [ref. 8], highlights the variation in the JFET's g_{fs} with drain current. The drain current will vary in a direct relationship to the magnitude of the input signal, assuming the bias is held constant.

The approximate output impedance for a U310 JFET with an I_D of 30 mA was calculated by extrapolating the manufacturer's data sheet to derive a value of 18 mS for the g_{fs} and via equation (2.10) returns a value of 55 ohms. The data in figure 2.3 was further extrapolated to create the equivalent drain current range of 2 to 60 mA. Equation (2.5) was used to plot figure 2.5 which shows the variation in voltage gain of a U310 common drain amplifier with two load resistors, 55 ohms and 5K7 ohms.

The choice of 5K7 ohms will become clear later but for the current purposes assume it is an arbitrary value many times larger than the JFET's drain/source internal resistance of 55 ohms for the intended operating bias. Another way of looking at figure 2.5 is, the steeper the gradient the larger the magnitude of distortion products for a given range of g_{fs}. The ideal is to have no variation in gain with variations in drain current i.e. a horizontal line. Of the two plots seen in figure 2.5 it is the 5700 ohm load that is the closest to a horizontal line and hence would produce the least amount of distortion.
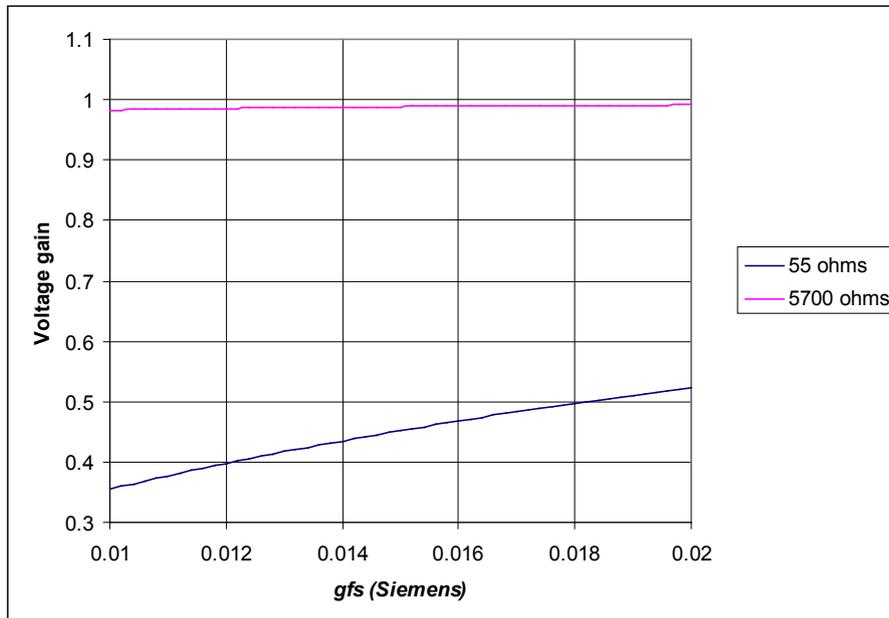

Figure 2.5   *Variation in voltage gain with two load resistances over a g_{fs} range 0.01 to 0.02 Siemens equivalent to a U310 drain current swing in the range between 2 to 60 mA*

In order to get a clearer picture of the mechanism which creates the gradient, figure 2.4(a) has been redrawn into its Thevenin equivalent model in (b) and then simplified in (c). The magnitude of the voltage seen at the source can be approximated as follows. If R_L is made very large relative to r_{ds} then very little current can flow around the circuit; equation (2.10) gives the relationship between r_{ds} and g_{fd}. The voltage appearing across R_L will therefore reflect the open circuit voltage of the internal voltage generator. The ratio of V_{RL} to V_{in} is given by equation (2.5) as

$$A_v = \frac{V_{out}}{V_{in}} = \frac{V_{RL}}{V_{in}} = \frac{g_{fd}R_L}{1 + g_{fd}R_L}$$

re-arranging to solve for V_{RL} gives,

$$V_{RL} = V_{in}\frac{g_{fd}R_L}{1 + g_{fd}R_L}$$

and by making R_L very large relative to r_{ds} simplifies to

$$V_{RL} \approx V_{in}$$

The voltage appearing at the source of the JFET is therefore approximately equal to V_{in}.

If the value of r_{ds} in figure 2.4(c) is held constant and $R_L = r_{ds}$ then the voltage seen across the load resistor R_L would be a smaller but faithful reproduction of V_{in} and readily calculated by,

$$V_{R_L} = V_{in}\frac{R_L}{R_L + r_{ds}}$$

(2.13)

In reality, r_{ds} is not constant. Equation (2.10) was used to generate figure 2.6. It shows the variation in r_{ds} with g_{fs}. Applying a sine wave to the gate of the JFET will show that r_{ds} will decrease in value in line with the magnitude of the positive excursions of the wave and increase with the magnitude of the negative. Inserting these variations in r_{ds} for the 360 degree conduction angle of a sine wave into equation (2.13) shows the positive half of the sine wave will be slightly larger in magnitude than the negative half. This asymmetrical distortion generates spurious products in proportion to the magnitude of the input signal and is a reason why large signals in the HF band generate spurs in a JFET input active HF antenna.
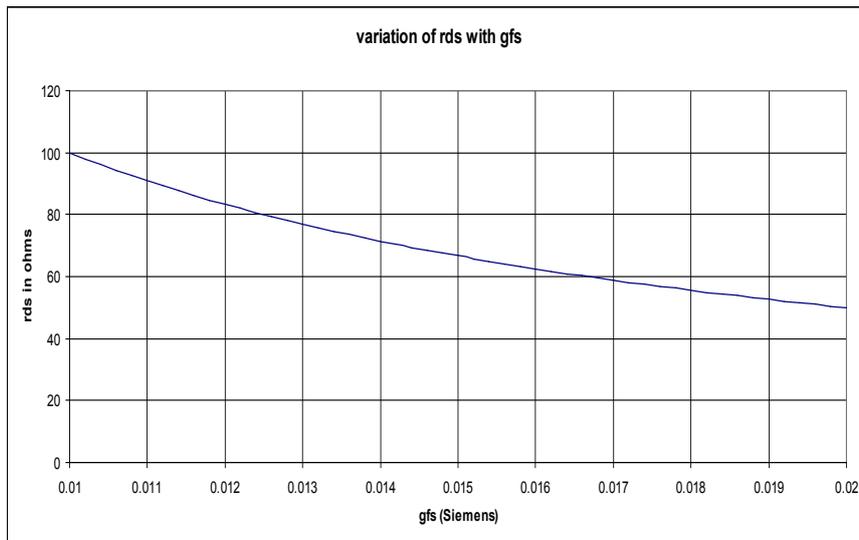
variation of rds with gfs

Figure 2.6 Variation in r_{ds} resistance over the g_{fs} range of 0.01 to 0.02 Siemens, equivalent to a U310 drain current swinging between 2 and 60 mA

This simplistic view of how distortion products are generated within the JFET channel is far from complete; however it is a sufficient analogy for the purpose of improving the design of an active antenna based on a JFET. There are two things that this analogy points to that can reduce the generation of distortion products. They are,

1. find a FET that has the smallest ohmic variation in r_{ds} with a given change in drain/source current and,
2. make R_L as large as possible in comparison to r_{ds} to reduce the effects that the variation in r_{ds} has on the voltage dropped across the load resistor R_L.

Small ohmic variations with given changes in drain current imply the use of power FETs. These devices however have large gate capacitances and would de-sensitise the active antenna by shunting the RF signal to ground via this capacitance. Small signal FETs have much smaller capacitances but have large values of r_{ds} with proportionally larger variations with changes in drain current. The U310 JFET is a good compromise between these two dilemmas.

2.2.6 JFET's Bias Altered by a Large Input Signal

The bias applied between the gate and source leads of the JFET can be altered if an input signal of sufficient magnitude is applied to the gate that overcomes the reverse bias of the diode between the gate and source leads inherent in the JFET's manufacture. Figure 2.7(*a*) shows the circuitry immediately attached to the JFET. The components in the dashed box are the circuit equivalent of the buffer's input impedance. In figure 2.7(*b*) the internal resistance between the drain and source leads (r_{ds}) is shown broken into two parts with the diode junction of the gate connected to where they are joined. The gate terminal is physically closer to the source connection on the JFET's channel than it is to the drain. This makes the resistance of the forward biased gate-source path measurably smaller than the

gate-drain; a useful point to remember when trying to identify the leads of a JFET in the absence of the manufacturer's data sheet. The reverse bias of the gate-source junction is approximately 1.17 volts and is set by the voltage drop across the 39 ohm source resistor by the 30 mA source current. The peak of the positive portion of an input signal must exceed this voltage for the gate-source diode to conduct. It is unlikely that the vector sum of all the signals induced into the 1.2 metre rod antenna will approach this value under normal operating conditions. The two most likely causes for this internal diode to conduct are:

1. close-in lightning strikes and,

2. the active antenna co-located with an RF transmitter.

The positive excursions of signals that are sufficient in magnitude to force the JFET's internal diode into conduction will deposit a charge across the 0.2 uF coupling capacitor to the buffer; the charge being deposited at a rate governed by the RC time constant where R = 5700 ohms and C = 0.2 uF. This charge and subsequent discharge, via approximately the same RC time constant, will increase the JFET's quiescent bias reducing the source current. The same positive excursion of signals causing gate current to flow will also charge the two series connected 220 pF between the antenna rod and the gate. The charge time constant of the equivalent 110 pF in series with the 5700 ohms input resistance of the buffer amplifier is very short. The discharge time constant however is considerably longer. When the forward conduction of the gate/source junction ceases the discharge time constant becomes 110 pF and the 4.7 Mohm gate bias resistor. The current flowing through the 4.7 Mohm resistor from the charge on the capacitor produces a voltage on the gate which is negative with respect to the chassis ground effectively reducing the current flow through the JFET. Equation (2.12) indicates the JFET's 1 dB compression point will be reduced dragging down with it the JFET's second and third order intercept points. Unwanted distortion products will therefore increase in magnitude until the bias is restored to its quiescent state.
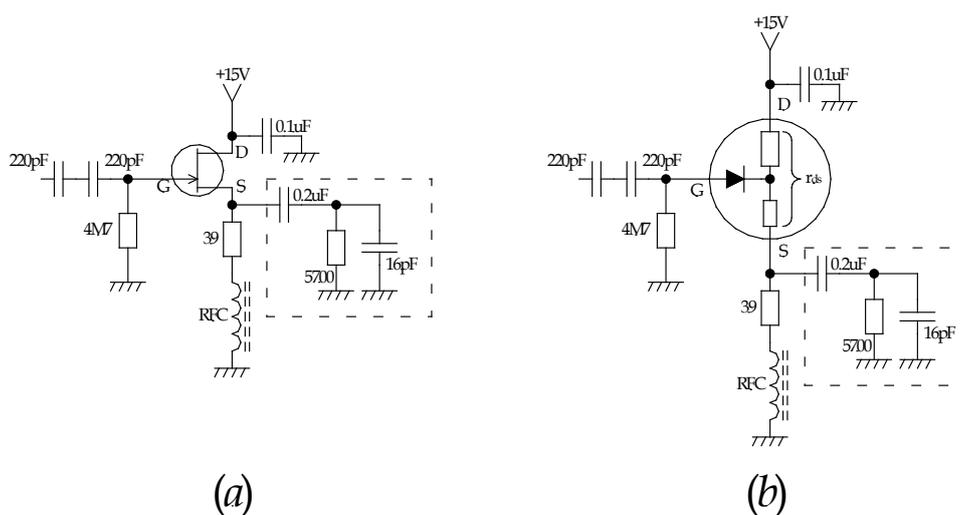

Figure 2.7   *(a) circuitry attached to the JFET, (b) the diode junction internal to a JFET*

Distant and close-in lightning strikes with their energy spread across a very wide bandwidth are easy to discern and with a duration of only a fraction of a second are an accepted interference to any receiver operating from LF to HF frequencies. The reverse biased protection diodes in the JFET's gate circuitry limit the magnitude of signal voltage from a close-in lightning strike that can be applied to the JFET. However, the gate-source internal diode will start to conduct before the limiting diodes start to work. A lightning flash is usually made up of many strokes and the JFET's bias is expected to return to normal in approximately five RC time constants 5*0.2E-6*5700 = approximately 6 mS after the last stroke. The other RC time constant affecting the gate bias is the discharge of the two series connected 220 pF capacitors in the JFET gate circuit. They will discharge in 5*110E-12*4.7E6 = approximately 2.6 mS.

There are no filters within the design of an active antenna beyond the 6 dB/octave of the distributed capacitance to ground of the active antenna's input circuitry and the frequency response of the active semiconductor devices. These types of antennas are wide open to RF energy from LF to the lower part of the VHF band. A high power transmitter co-located with an active antenna can be keyed on/off as required or be continually on. If the RF voltage induced into the rod is of sufficient magnitude to forward bias the JFET's internal diode, then as before the JFET's source current will decrease. This will cause intermodulation products to be generated. If the receiver connected to the active antenna is not tuned to where these distortion products will fall in frequency then a pronounced increased in noise floor will be noticed for the duration in which the transmitter is operating desensitising any receiver connected to it. Care needs to be taken when installing any active antenna in an antenna site to ensure there is adequate distance between an active antenna and any transmitters.

2.2.7 Static Electricity and Lightning Protection

The fine weather electric field vector at ground level is in the order of 100 V/m with the Earth negatively charged and the atmosphere above the Earth being positively charged [ref. 9, pg. 313]. Clear air can support a charge which increases the fine weather field sometimes to a magnitude where breakdown can occur giving rise to the "bolt from the blue" phenomenon [ref. 9, pg. 24]. FET input active antennas, with their very high DC input resistance between the antenna rod and ground, need protection from the static E fields that can develop a charge across the distributed input capacitance of the antenna exceeding its static voltage rating. A 4M7 resistor connected between the base of the antenna rod and ground is used to bleed away any normal charge build-up. A gas discharge tube with a spark-over voltage of approximately 90 V is placed in parallel with the 4M7 resistor for those infrequent occasions when a charged mass of air deposits a charge faster than the 4M7 resistor can bleed it away. The gas discharge tube also serves as the front line defence by shunting to ground the energy induced into the antenna rod from a nearby cloud to ground lightning strike. Silver mica DC blocking capacitors connected between the base of the antenna rod and the JFET's gate terminal prevent the voltage drop across the 4M7 resistor through the bleeding of the antenna charge from altering the JFET's bias point.

JFET input active antennas are generally broadband devices that are capable of responding to RF energy from LF to low band VHF frequencies. A cloud to ground lightning strike consists of multiple pulses of discharge current and has a large portion of its energy spread over this frequency range. Active antennas need protection from these large induced time varying voltages. Two 220 pF 350 volt silver mica capacitors are connected in series between the antenna rod and the gate of the JFET. Silver mica capacitors have very low internally generated noise and the combined 700 volt rating is many times the spark-over voltage of the gas discharge tube. The junction of the two capacitors has a pair of reverse biased microwave diodes biased at half the Vcc rail. These diodes are considerably faster than the JFET that they are protecting and protect the JFET's gate from large voltage excursions whose magnitude is insufficient to fire the gas discharge tube. The combined 110 pF capacitance of the two 220 pF silver mica capacitors connected in series together with the measured 13 pF of the JFET input circuitry form a voltage divider which reduces the signal level appearing at the base of the rod antenna by 1dB when it is coupled via this circuitry to the JFET's gate. This small insertion loss, which increases the active antenna's noise figure by 1 dB, is an acceptable price to pay for lightning protection.

2.2.8   Suppression of Parasitic Oscillation caused by the Antenna Rod

Figure 2.8 is the schematic of the input circuitry of the active antenna. The 22 ohm resistor in series with the JFET's gate is a VHF parasitic stopper and is required to prevent the JFET from breaking into lowband VHF oscillation when an antenna rod is connected to the JFET's input circuitry. Oscillations usually occur at a frequency just above the natural ¼ wave resonance of the antenna rod. This is where the rod appears as an inductance which resonates with the circuit's distributed capacitance forming a Colpitts oscillator. A more detailed explanation on how this oscillator is formed is given in section 3.1. The value of resistor is dependent upon the JFET, circuit layout, antenna rod length and the completed antenna's installed environment. This resistor may need to be increased to suit a different physical layout or a much longer antenna rod length. The total capacitance seen at the gate terminal of the JFET is in the order of 6 pF. The capacitive reactance of this capacitance at 30 MHz is 884 ohms. The 22 ohm resistor installed in the series path of this high reactance will have minimal effect on the circuit's overall performance.
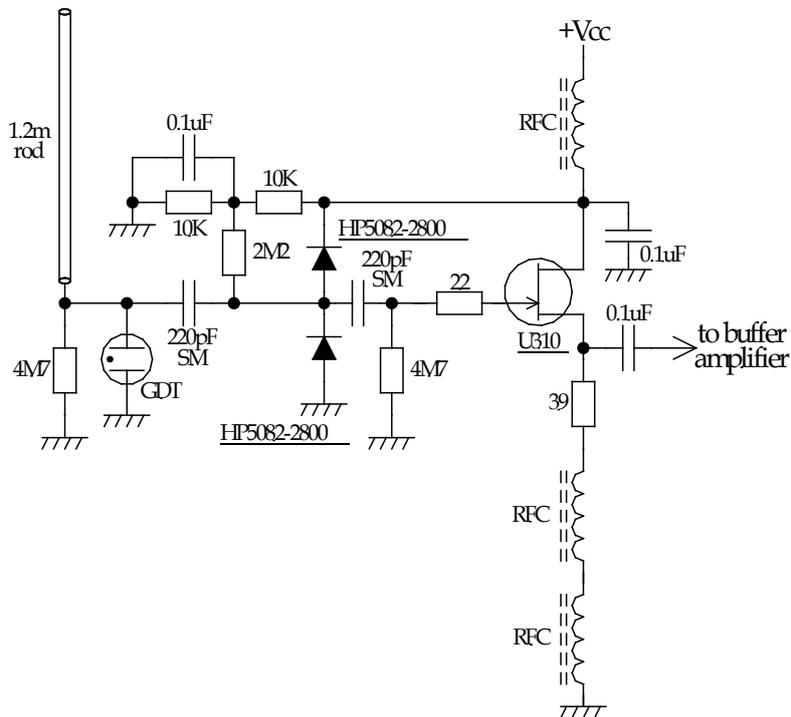
Figure 2.8   Input circuitry of the active antenna

2.3   The Output Buffer Stage

The output buffer stage must have:

- an input impedance as high as possible to minimise the distortion products generated within the JFET

- operate over a minimum frequency range of 2 to 30 MHz with constant gain

- have a broadband 50 ohms output impedance to control any reflected waves from a receiver and,

- some sort of push-pull configuration to help suppress the even order distortion products.

Several push-pull configurations were tried before settling on the design as seen in figure 2.9. This design produced the highest OPIp[2] and OPIp[3] points at the cost of a -6 dB amplifier gain. The complementary pair of transistors should have their current gains (β) matched in order to maximise the suppression of the even order distortion products. Because the transistors are used as emitter followers (voltage gain less than one), and the approximate 3.6 ohms output impedance seen at the transistor's emitter is loaded by a 96.4 ohm un-bypassed emitter resistor, then a fair spread in β can be tolerated when matching transistors. Had the transistors been connected to give a voltage gain (common emitter) then much more stringent requirements would be required when matching transistors.
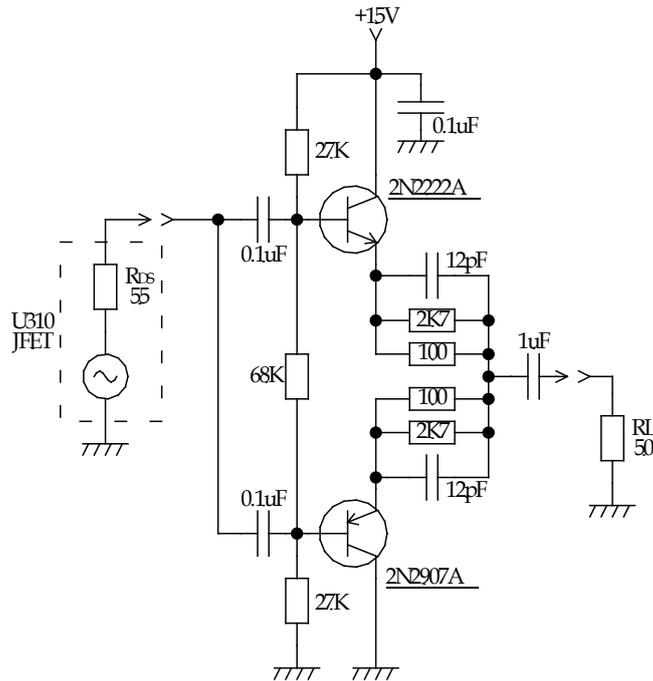
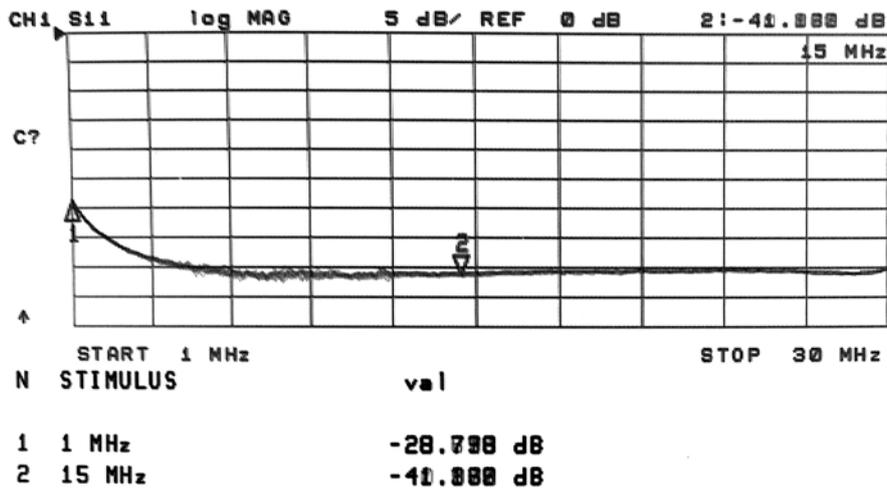*Figure 2.9   Output Buffer circuitry*

Figure 2.10   Measured return loss at the output terminal of the completed active antenna

The input impedance of the buffer was measured and can be represented as a 5K7 ohm resistor in parallel with a capacitance of 16 pF. This is the load that will be connected to the output of the JFET input stage as seen in figure 2.8. Figure 2.10 is the buffer amplifier's measured output return loss of -28 dB at 1 MHz improving to a minimum of -40 dB from 4.5 MHz to 30 MHz indicating the output impedance of the buffer is a broadband 50 ohms. The 12 pF capacitor across the paralleled emitter resistors is to compensate for the output

DST-Group-TR-3522

circuit's distributed inductance. This will need to be finely adjusted if the physical component layout is changed in any way.

2.3.1 Changes needed to suit a 75 ohm system

The output impedance can easily be changed from 50 ohms to suit a 75 ohm system if this is preferred. All that is required is for the un-bypassed emitter resistor parallel combination of 100 and 2K7 as seen in figure 2.9 to be replaced with the parallel combination of 150 and 6K2. The 12 pF frequency compensation capacitor will need to have its value reduced slightly and is best done with the aid of a network analyser in the high frequency part of the HF band on the finished design.

2.3.2 Testing the buffer amplifier

The buffer amplifier was connected as seen in figure 2.11 with a 50 ohm termination resistor installed at its input. The forward gain at 8 MHz was measured to be -6.3 dB. In figure 2.12 the red top trace is the input from a signal generator as applied to the buffer stage input and the bottom blue trace the output. Note the 2 dB/division scale and the output follows the input from 1 MHz to approximately 50 MHz. The gain slowly increases from -6.3 dB at 50 MHz to -5.3dB at 100 MHz.

The single RF generator was replaced with an RF two tone generator (7 MHz and 11 MHz) and the following output parameters measured, $OPIp^3$ = +56.5 dBm and $OPIp^2$ = +110dBm. These figures exceed the measured parameters of the JFET input stage and therefore will not be the limiting factor in the completed cascade design.

2.3.3 The need for a 50 ohm output impedance and the associated 6 dB loss

The input impedance of most modern receivers is usually specified as '*nominally 50 ohms*' implying the input impedance may wander with frequency in this vicinity. This is especially true if tracking pre-selection filters have been employed in an effort to control the generation of unwanted spurs. Figure 2.13 is the measured input impedance at 7 MHz of a R&S EK890 HF receiver when it was set to receive a 7 MHz signal. Notice the real component is only 21 ohms with a small amount of series inductance of approximately +j2 ohms; a second EK890 receiver measured 32+*j*3 ohms at the same frequency.

RF energy reflected back to the active antenna from the receiver can have an appreciable effect on the system gain, phase and generated distortion products; the system being the active antenna connected to a long length of co-axial cable whose other end is connected to the input of a receiver.
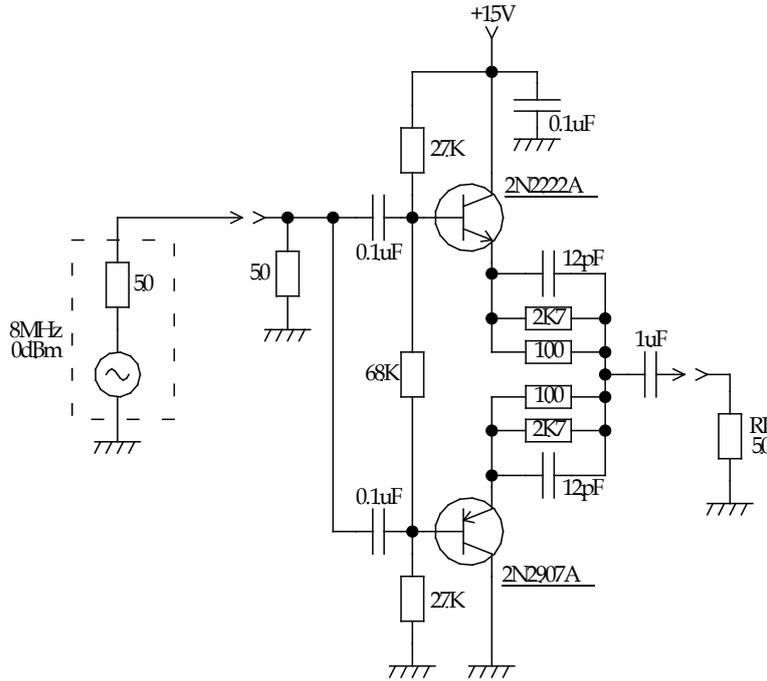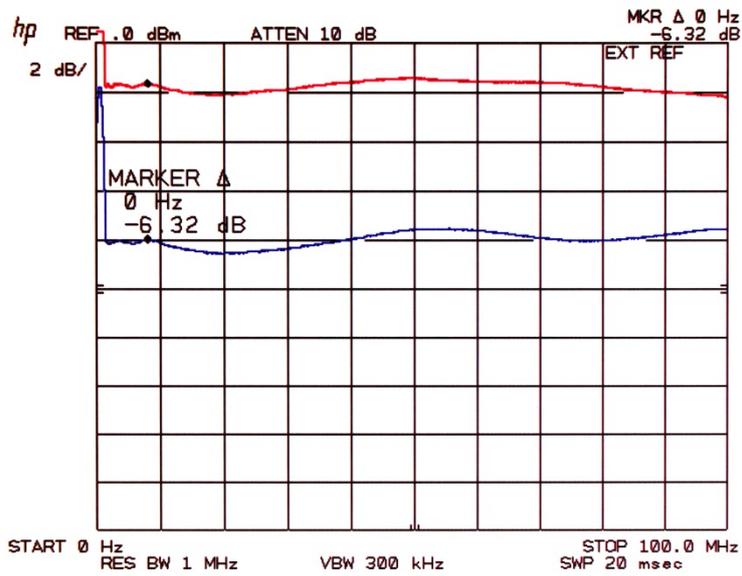
Figure 2.11   The output buffer circuitry test setup

Figure 2.12   Frequency response of the output buffer circuit, Δ marker at 8 MHz
        Red: input signal level into buffer input
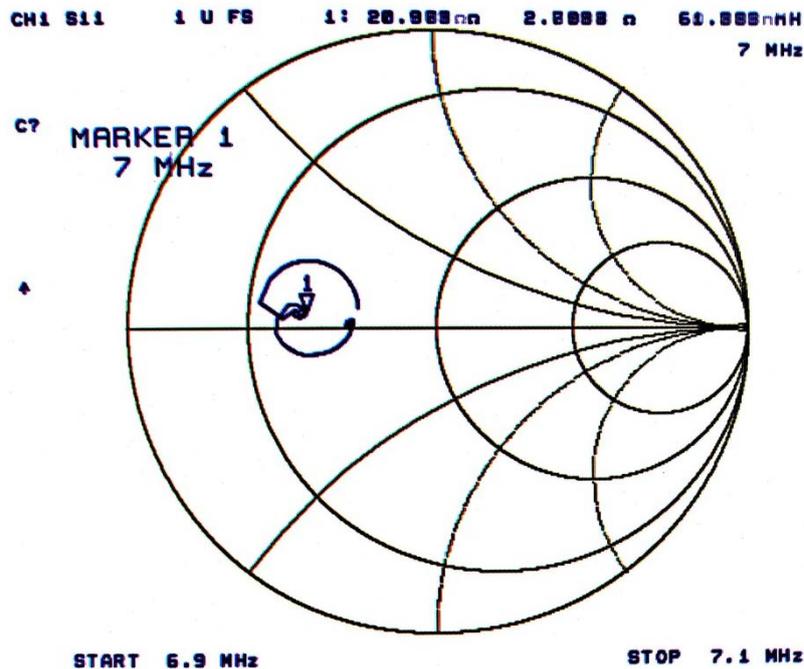   *Blue: buffer output*

Figure 2.13   Measured input impedance of 21+j2 ohms seen at the antenna input of a R&S EK890 receiver when tuned to 7 MHz

Some designs for active antennas freely available in the public domain e.g. on the internet, do not have an output impedance matching the transmission line connecting the antenna to the receiver, usually 50 ohms, and the output is taken directly from the emitters of the push-pull pair. Figure 2.9 has been reconfigured to reflect this and is seen in figure 2.14. This configuration's output impedance is about 1.8 ohms. The main attraction for connecting the output buffer this way is the voltage gain approaches unity when used in a 50 ohm system. This is equivalent to an approximate 6 dB increase in output signal level when compared with a stage that has been matched to the transmission line. If the assumption is made that the input to the receiver is a constant broadband 50 ohms then figure 2.15 is the simulated expected system frequency response with a 30 m length of ideal co-ax between the active antenna and the receiver; notice the flat gain and linear phase change with frequency and the voltage gain just slightly less than one.

This however is not reality. Figure 2.13 shows the EK890 receiver has an input resistance of 21 ohms at 7 MHz. While this resistance will vary as the receiver is tuned to other frequencies, a fixed input resistance of 21 ohms will be used as an example of how standing waves from the receiver can alter the system voltage gain and phase with frequency. The input impedance of the receiver in the simulation model was changed from 50 ohms to 21 ohms and figure 2.16 displays the result.
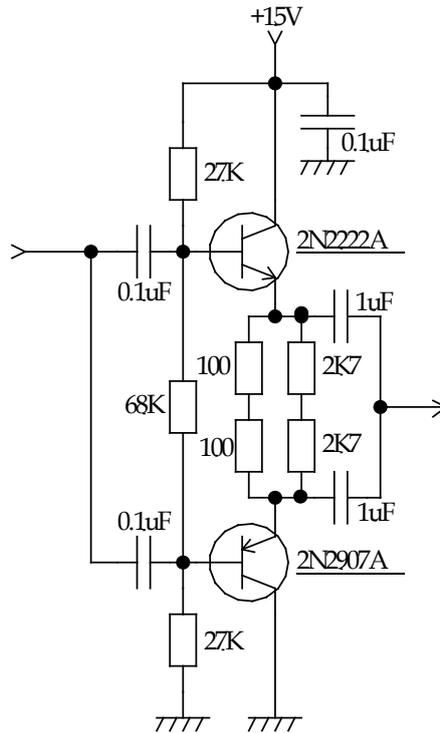
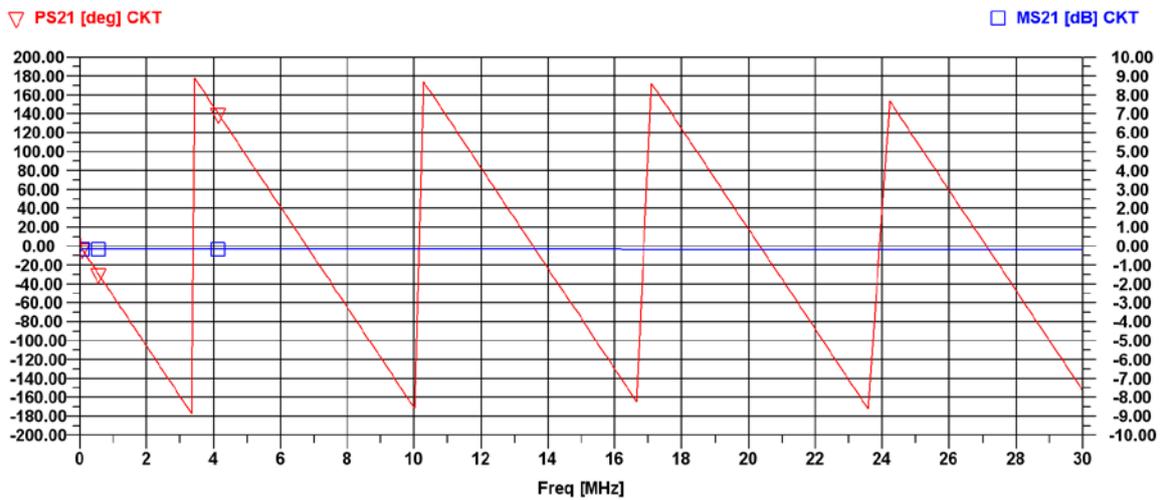*Figure 2.14   Output buffer circuit. The output impedance of this configuration is approximately 1.8 ohms*

Figure 2.15   Phase(red left scale) and magnitude (blue right scale) response with frequency of receiving system with output buffer output Z = 1.8 ohms, 30 m of ideal 50 ohm co-ax connecting to a receiver input Z = 50 ohms
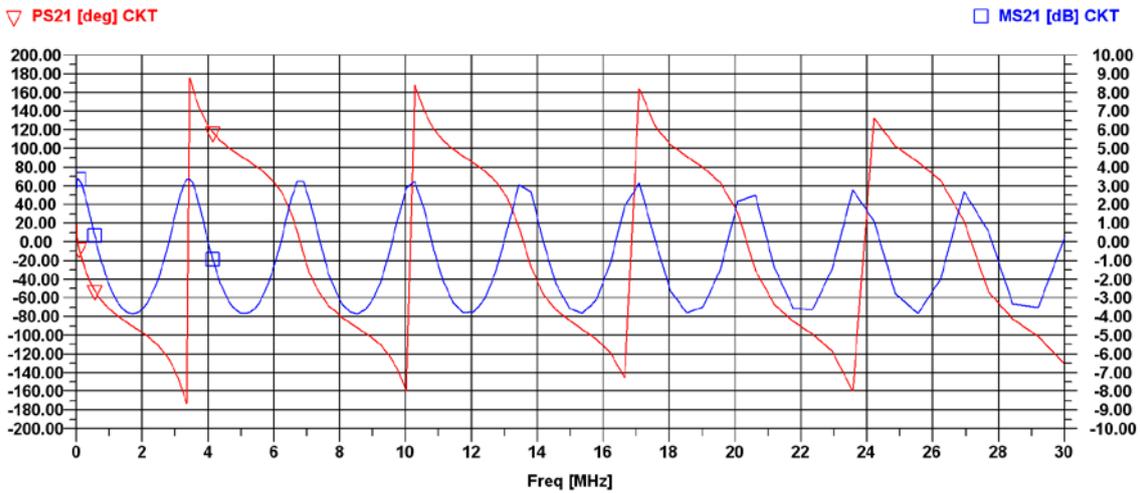
Figure 2.16   Phase (red) and magnitude (blue) response with frequency of receiving system with output buffer output Z = 1.8 ohms, 30 m of ideal 50 ohms co-ax connecting to a receiver input Z = 21 ohms

The mismatch between the ideal 50 ohm transmission line and 21 ohms input resistance of the receiver has reflected some of the RF signal back to the output of the buffer stage where the mismatch between the transmission line and the 1.8 ohms buffer output impedance re-reflects this signal back to the receiver. The magnitude of the reflected signal diminishes with each reflection as a portion of the signal is dissipated in the output of the buffer amplifier and the receiver. The termination at the ends of the co-ax that is closest to 50 ohms will dissipate most of the reflected signal. The 1.8 ohms output impedance of the buffer amplifier will therefore re-reflect most of the reflected energy back to the receiver where it is algebraically summed with the current forward wave. As can be seen in figure 2.16, the result is a significant gain and phase variation with frequency.
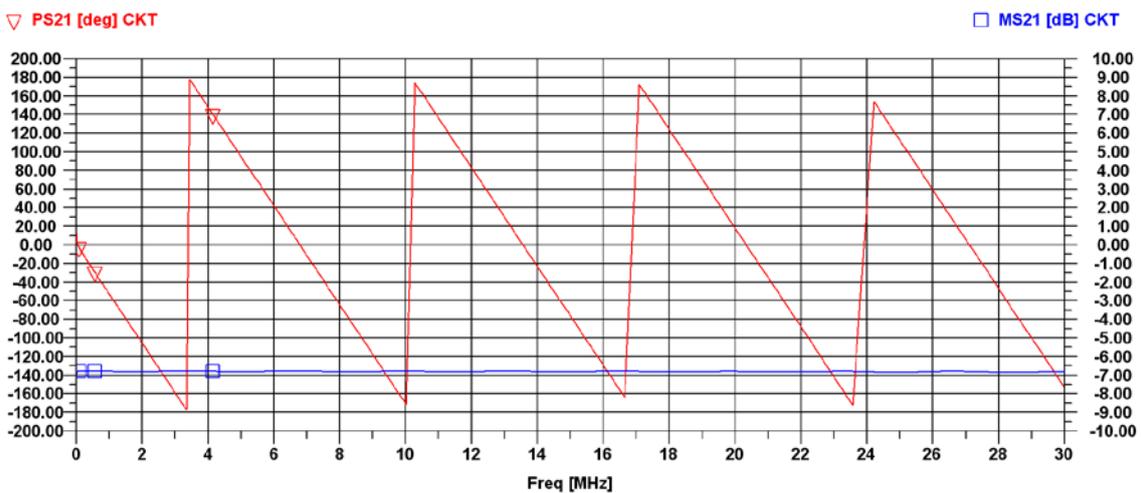

Figure 2.17   Phase (red LH side, degrees) and magnitude (blue RH side, dB) response with frequency of receiving system with output buffer output Z = 50 ohms, 30 m of ideal 50 ohm co-ax connecting to a receiver input Z = 21 ohms

The simulation was reconfigured to reflect figure 2.11, an active antenna with a broadband output impedance of 50 ohms and a receiver input resistance of 21 ohms. The simulation carried out again resulting in the response of figure 2.17. Here the reflected wave from the receiver whose input resistance is 21 ohms is totally absorbed by the 50 ohm output impedance of the buffer amplifier resulting in a 0.05 dB ripple with frequency at the receiver input; note the smooth phase response.

2.3.3.1   A Note About Passive Broadband Receive Antennas

While this simulation was aimed at active antennas it is also relevant to passive broadband receive antennas; for example, the elevated feed monopole. The input impedance of these antennas, which is the output impedance when used to feed signals to a receiver, varies with frequency and may never be 50 ohms throughout its operational frequency range. It is for this reason that a broadband pre-amplifier should be placed directly at the antenna's output terminals so the 50 ohms output impedance of the amplifier drives the co-axial cable which feeds the receiver. While an amplifier's output impedance is usually affected by its input termination and the magnitude and type of feedback employed within the amplifier, the variation in the amplifier's output impedance with frequency will be considerably less than that of the antenna to which it is connected to and also much closer to the required 50 ohms. This would minimise unwanted changes in phase and magnitude should the receiver's input impedance vary with frequency.

2.4   +15V Voltage Regulator

Some three terminal voltage regulators are quite noisy. Their dynamic output impedance in the vicinity of an ohm makes it quite a daunting task to filter out the internal noise generated from such a low impedance source. The LM317LZ three terminal regulator has an acceptable noise level especially when used with a PNP current boost transistor. PNP transistors are known for their low $1/f$ noise [ref. 13 pg. 353].

The +15V voltage regulator seen in the final design has been designed to be low noise. Its regulated output is filtered via separate RF chokes before being connected to the input and output buffer stages. This ensures the noise performance of the final active antenna is limited by those two stages and not by the noise contribution of the voltage regulator. A steering diode (S1G) is used for reverse polarity protection.

2.5   The Final Circuit

Figure 2.18 is the final circuit of the active antenna. The entire prototype circuit was built into the lid of a IP66 rated diecast aluminium box measuring 64x58x34mm using "in air three dimensional point to point wiring" on a copper ground plane made from a piece of FR4 PCB material, see figure 2.19.

Figure 2.20 shows the recessed BNC connector. This protects the connector from damage caused by rain. Note the 1 mm hole drilled in the high point of the recess to allow the air

pressure to equalise with temperature changes. An earlier version of this active antenna which did not have this hole sucked in rain water when the cool rain rapidly dropped the temperature of the box. Figure 2.21 is the completed assembled antenna. The machining on the right is to locate the U bolt used to attach the antenna to a metal post.
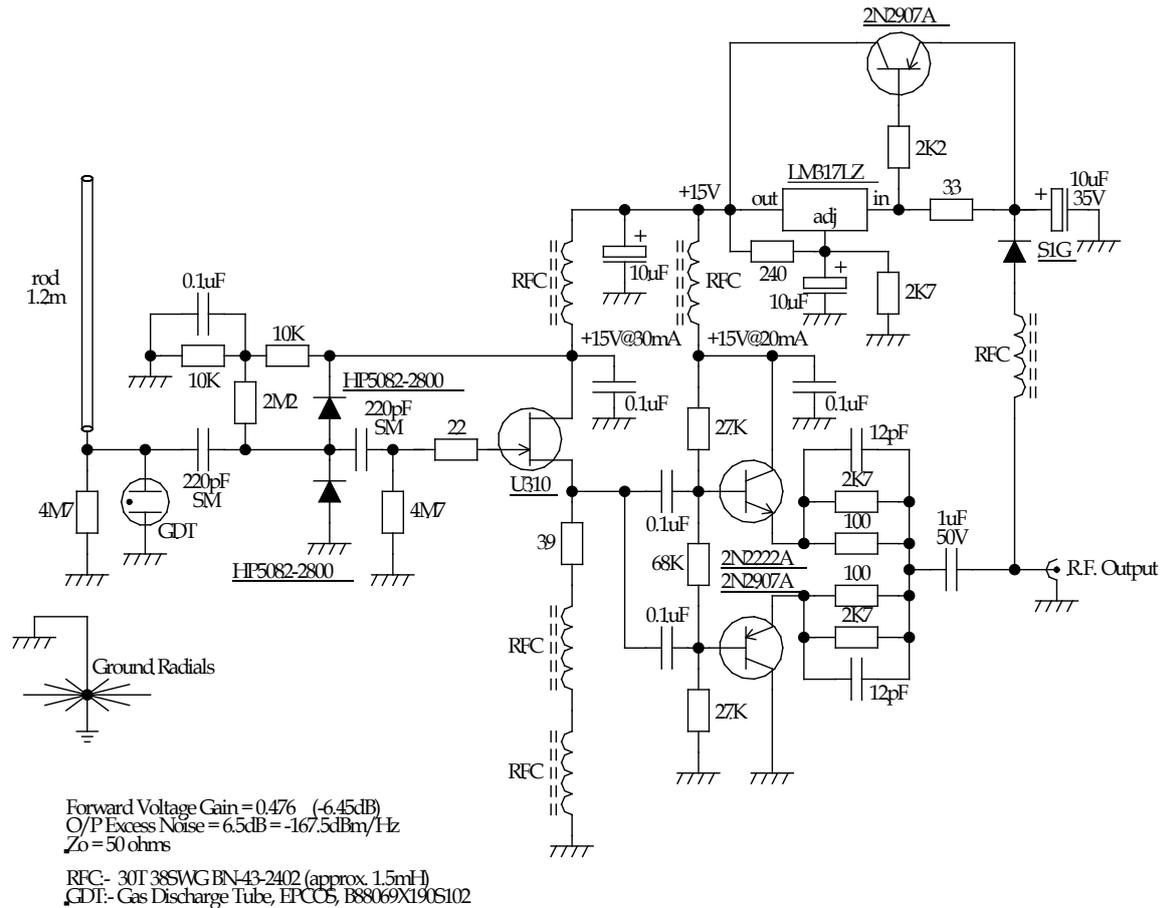

Figure 2.18   *The final circuit of the active antenna*

Figure 2.19   Point to point wiring of the active antenna. The two yellow series connected silver mica capacitors are in the foreground

Figure 2.20   View from the bottom showing the recessed BNC connector

Figure 2.21   Assembled antenna with rod

2.6 Installation Recommendations

There are three general methods that can be used to install an active monopole antenna at a site.

1. ground level on a metallic mat or radial ground plane.

2. elevated above the ground with the active antenna's outer metal case electrically connected to the top of a metal pole with no metallic ground plane at its base, figure 2.22.

3. elevated above the ground via a metal pole with a radial type or wire mesh ground plane at the pole base, figure 2.23.

If a radial ground mat is used then it should consist of a minimum of 16 radials. For the antenna setup described in figure 2.23 which was used in all field trials, h_1 = 1.2 m, h_2 = 1.2 m and h_3 = 0.3 m. The length of each of the 16 radial wires forming the ground plane was $ah_1 + h_2 \approx 1.7$ m; where the factor a is derived in section 2.6.1.

An RF current choke, figure 2.25 and 2.26, should be inserted in series with the co-axial line at the point where the ground radials end. This effectively turns the outer braid of the co-ax between the output of the active antenna and the current choke into a 17th ground radial. A current choke is also installed on the outside of an earthed co-ax cable entry panel into the RF hut, see figure 2.24. The twisted pair of enamel covered wires used to make the co-axial cable RF current chokes should have a characteristic impedance the same as that of the co-axial cable being used to connect the active antenna to the receiver. The RF current choke at the cable entry panel prevents RF currents induced into the outer braid of the co-axial cable from flowing towards the earthed cable entry panel inducing a differential voltage between the inner and outer conductors of the co-ax. The outer braid of the co-ax would otherwise act like a horizontal monopole antenna with the high current point being at the earthed cable entry panel. Figure 2.27 is the response of the active antenna installation before the RF current chokes were fitted. In figure 2.28 the RF current chokes have been fitted as per figure 2.24. Notice the suppression of the two resonances and also the general suppression of unwanted currents below approximately 6.5 MHz.
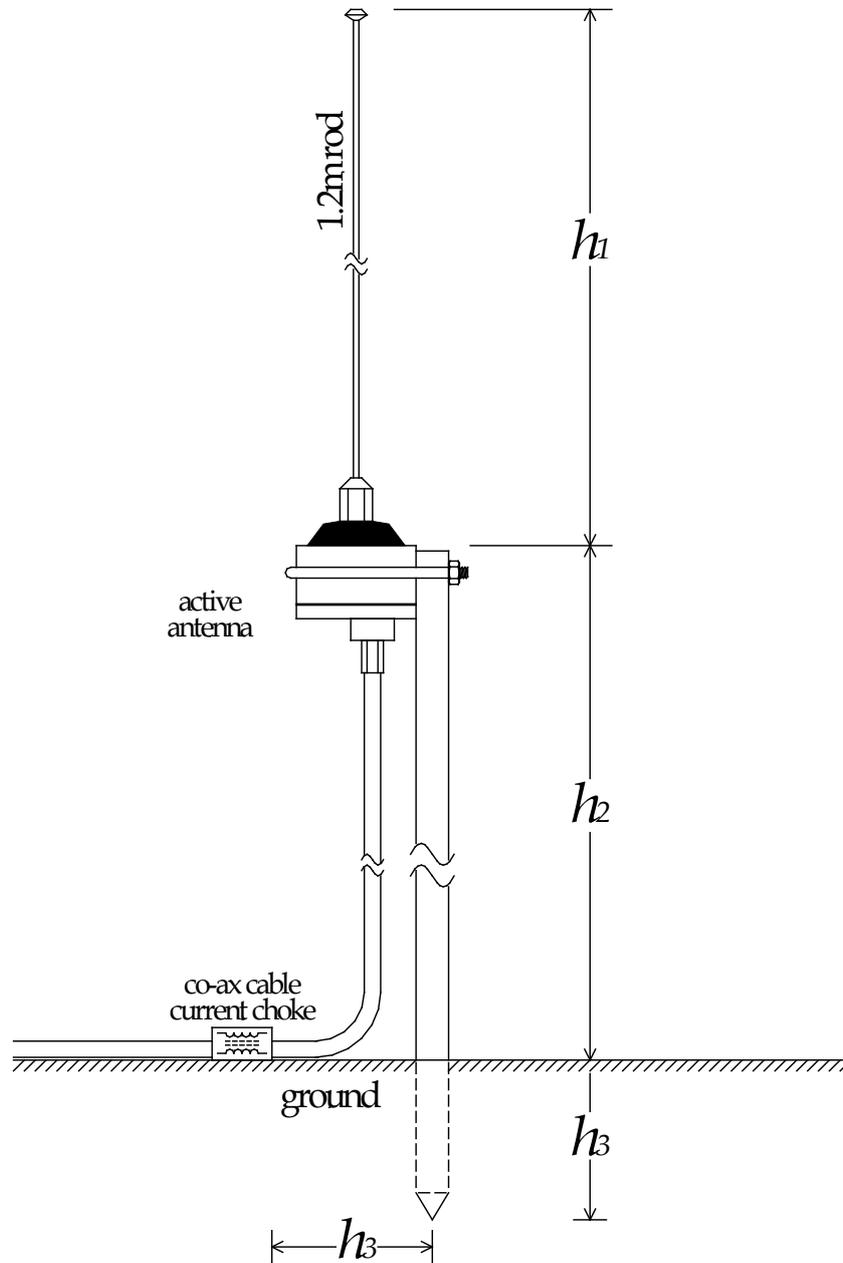
Figure 2.22   Recommended installation for a given mast height and rod length for an active antenna; note the position of the co-axial current choke.
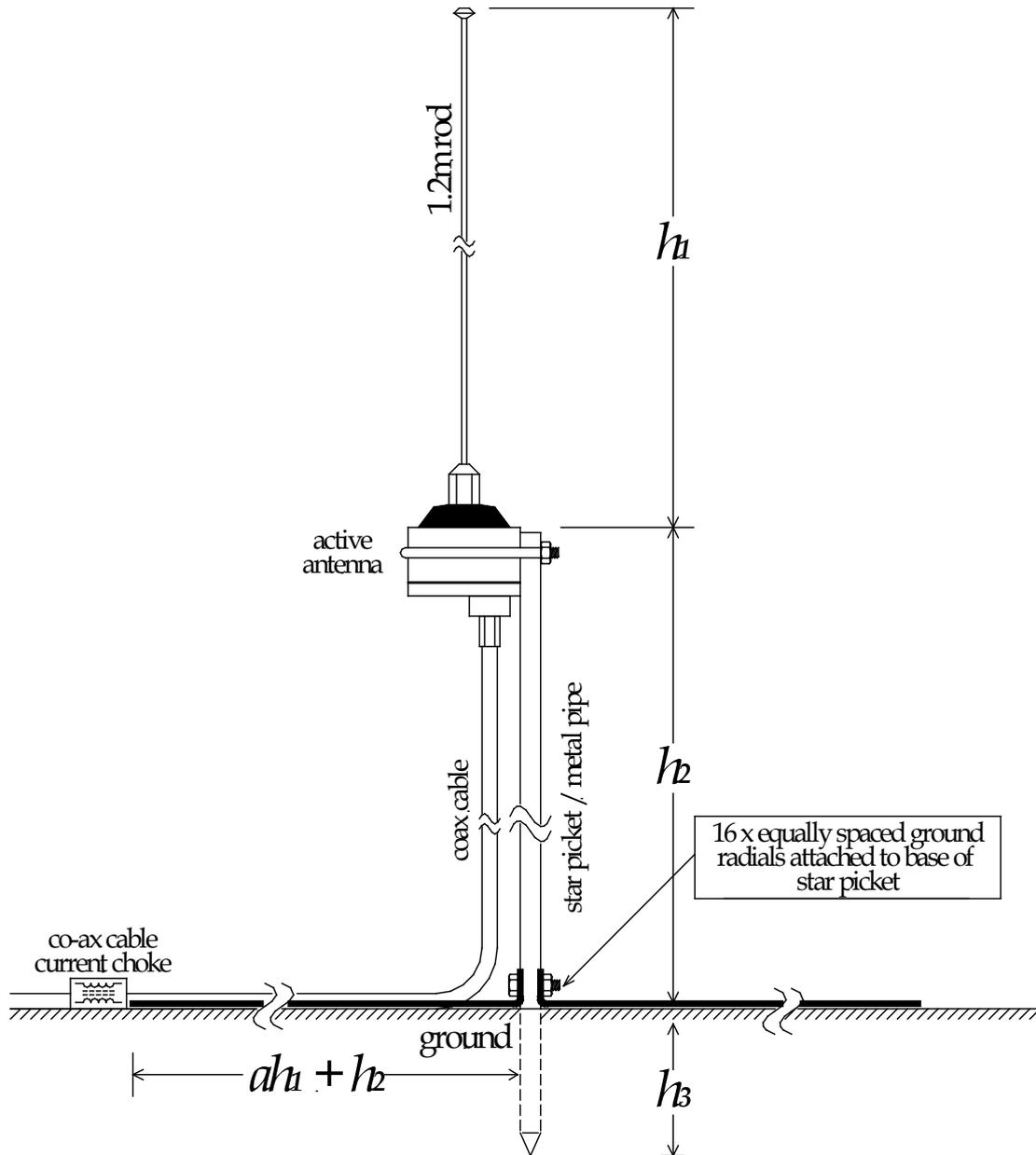
Figure 2.23   Recommended lengths of ground radials for a given mast height and rod length for an active antenna
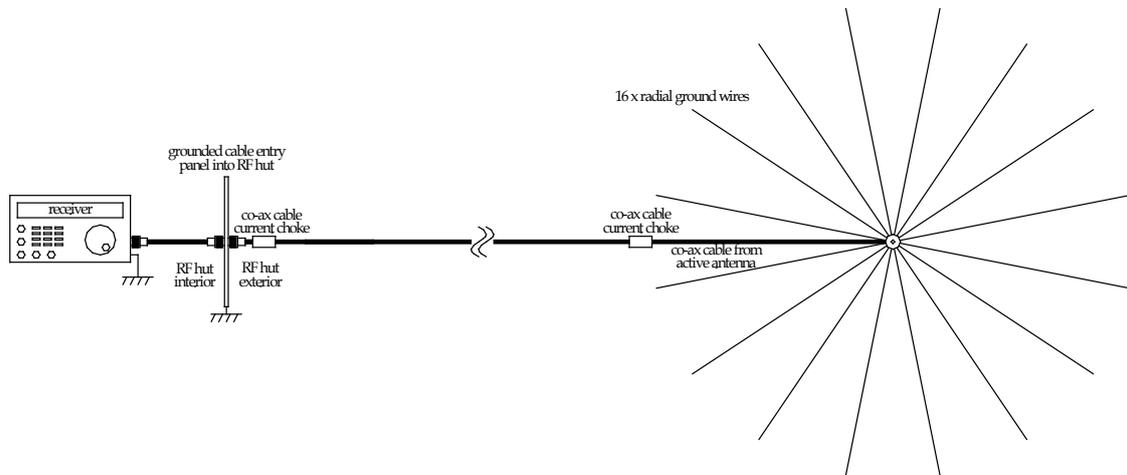
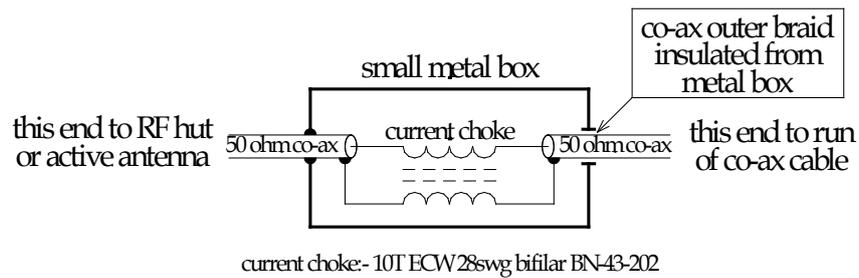*Figure 2.24   Recommended installation for an active antenna*

Figure 2.25   Construction details of current choke

Figure 2.26   RF current choke housed is a small metal box. Note the isolated BNC connector on the right hand side
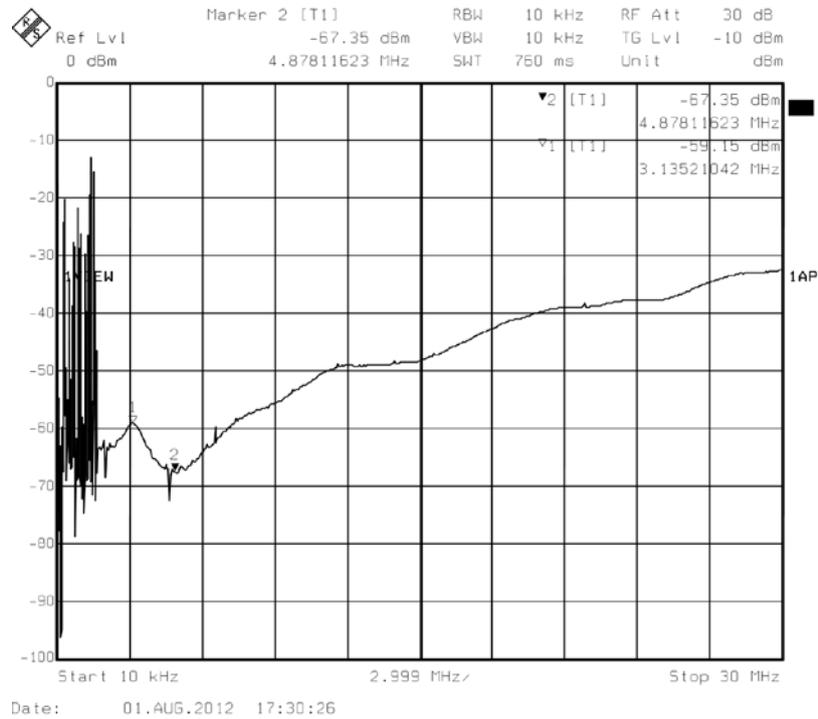
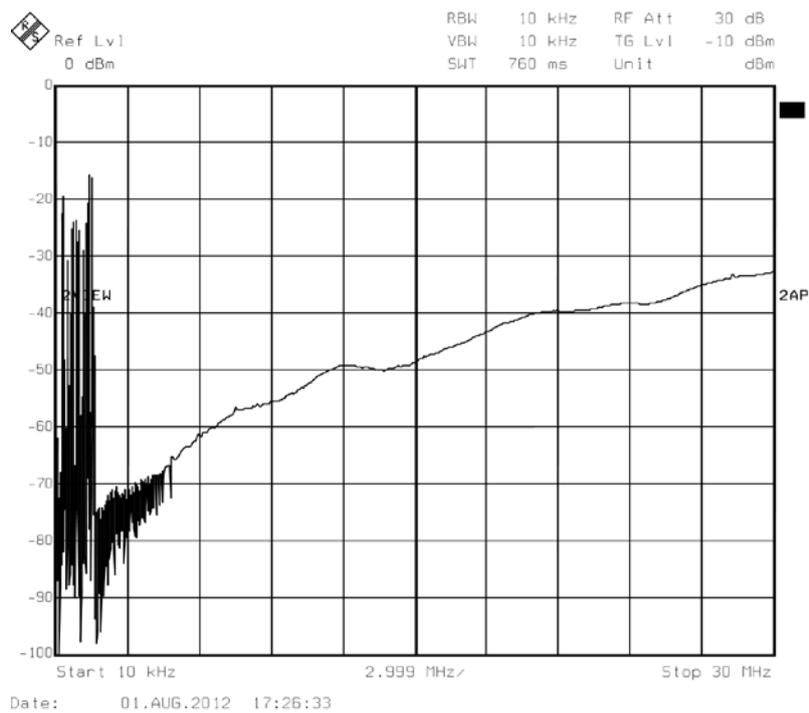*Figure 2.27   Active antenna response with no RF current chokes fitted*

Figure 2.28   Active antenna response with co-axial current chokes fitted at ground radial end and entrance to RF hut. Notice the suppression of currents below 6.5 MHz
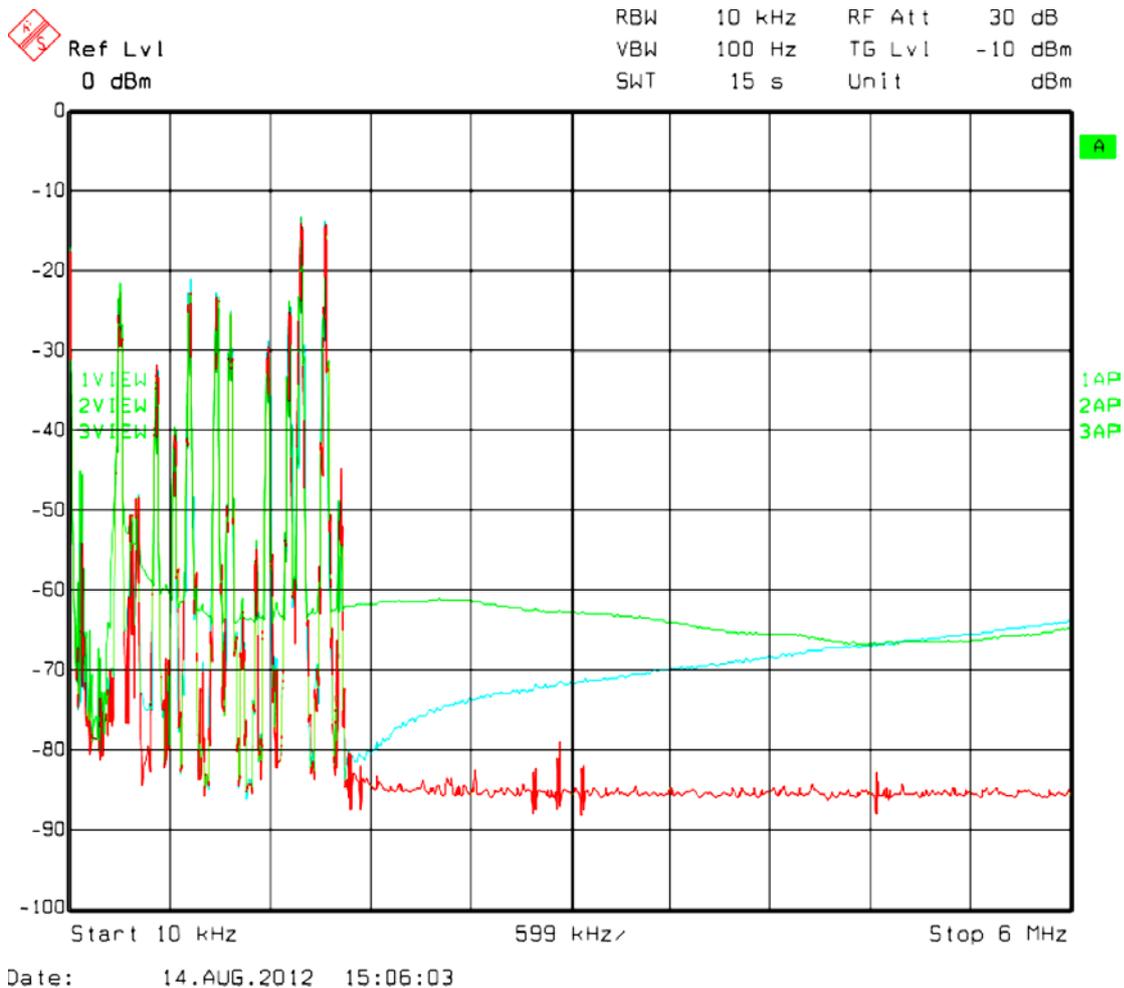
Figure 2.29   Active antenna response from 10 KHz to 6 MHz with (cyan trace) and without (green trace) the co-axial current chokes fitted. Notice the magnitude of the AM stations are relatively unchanged. The red trace is the response with the Tx loop disabled.

Figure 2.29 was included to demonstrate effectiveness of the RF current chokes and how the magnitudes of the AM radio broadcast radio stations below 1.8 MHz are relatively unchanged with their insertion. The magnitudes of these signals are dependent upon the state of the AM modulation at the instant of sampling. Of importance is the level of noise in the space between the signals. The test set-up was based on a 20 m/side equilateral triangle with the RF hut, broadband Tx test loop [ref. 12, appendix F] and active antenna at each of the points of the triangle. There was 23 metres of co-ax cable between the Tx loop and the RF hut and 20 metres of co-ax between the active antenna ground radial end and the hut. The active antenna's co-ax was connected to the cable entry panel; the co-ax from the Tx loop entered the RF hut through a crack in a doorway on the opposite side of the hut to the cable entry panel. The Tx loop's co-ax cable had 'clip-on' ferrite beads functioning as RF chokes at 1 metre spacing to control radiation from the cable's outer braid from the Tx loop inducing currents into its own feeding co-ax.

At 6 MHz and below all points in the test set-up are in the near field of the Tx loop. This would simulate an active antenna's receiver and associated co-axial feed in an electronic rich installed environment, i.e. racks of electronic equipment some with switch mode power supplies. The green trace shows signals being induced into the outer braid of the co-ax creating a differential voltage between the co-ax's inner and outer conductors. The cyan trace show's how effective the RF current chokes are at suppressing these near field signals induced into the outer braid. The red trace is the response without excitation from the Tx loop.

2.6.1 Antenna and Mast Resonance

The height of the star picket/metal pipe used as a mast to elevate the active antenna above ground has its lower portion imbedded in the ground with ground radials attached at the air/ground interface. This mast acts as a grounded vertical monopole antenna which has its first resonance where the total apparent length equals 0.25λ. The mast is made up of two heights; the above ground height and the below ground height. The full length of this mast is used unaltered in the resonance calculation. The only effect the portion of the mast that is below ground has is to lower the Q of resonance and this is dependent upon the soil type and moisture content of the ground.

The metallic rod used for the antenna however appears electrically to be much shorter than its physical length. This is due to the high input impedance of the rod's termination being predominately capacitive appears in series with the rod and has a shortening effect. The shortening factor a which the rod's physical length is multiplied by, can be readily calculated for any active antenna by finding the resonance when the active antenna is attached to a mast of known total length. Horizontal ground radials being orthogonal to the vertical mast and antenna rod have very little effect on this resonance. Therefore the search for resonance can be carried out with or without the ground radials attached.

Figure 2.30 is the response of the active antenna via a vertically polarized Tx loop up to 100 MHz ; ¼ wave resonance is at 38 MHz and ½ wave resonance is the dip at 75.35 MHz. Equation (2.14) is used to find the shortening factor a.

$$a = \frac{\left(\dfrac{300}{4f_{1/4}} - h_2 - h_3\right)}{h_1} \tag{2.14}$$

where: $f_{1/4}$ = 38 MHz = peak in the active antenna's response indicating the antenna installation's ¼ wave resonance, marker 1 from figure 2.30 (MHz)

h_1 = physical height of antenna rod (m)
h_2 = physical height of mast above ground (m)
h_3 = physical height of mast below ground (m)
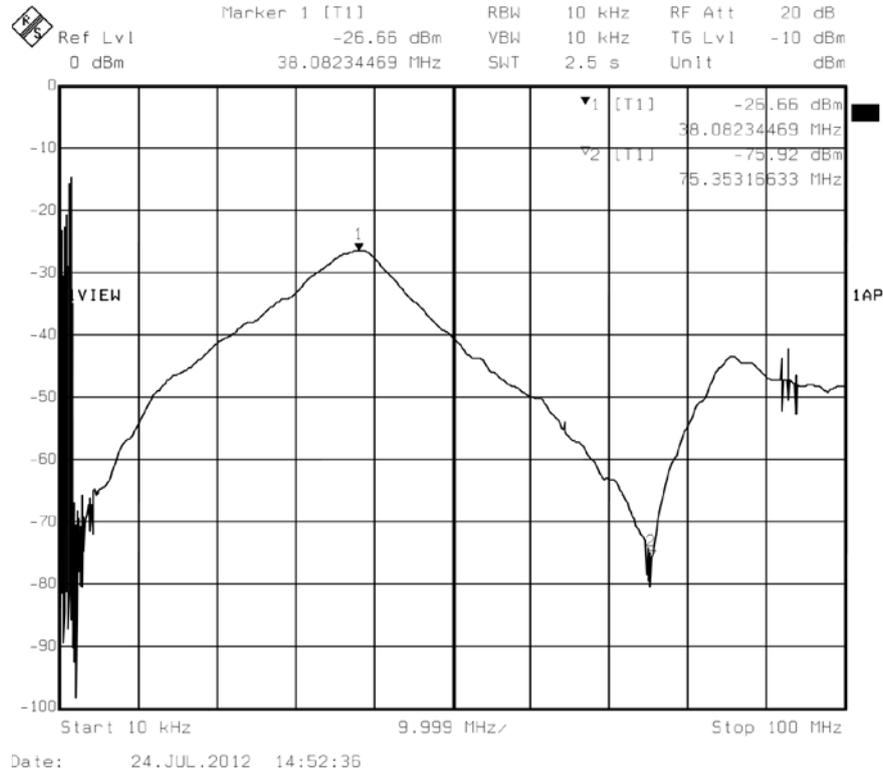a = a factor peculiar to the design of a particular active antenna

Figure 2.30 *Maximum response at 38MHz where the total apparent length of the active antenna appears as a resonant 1/4 wavelength monopole and minimum response at 75.3MHz where the antenna is 1/2 wavelength long*

Once the shortening factor is known then equation (2.15) can be used to calculate the frequency where an active antenna's response is maximum for any given mast height and rod length. This equation was checked by replacing the fixed length antenna rod of 1.2 m with a 0.57 m to 3.5 m telescopic rod. Random lengths of the telescopic rod were chosen and the calculated resonance was compared with the measured resonance. There was agreement to within a few percent between the calculated and measured resonances.

$$f_{1/4} = \frac{300}{4(ah_1 + h_2 + h_3)}$$

(2.15)

where: $f_{1/4}$ = ¼ wave resonance of an active antenna operating in the HF band (MHz)

         a = 0.3947 = a factor peculiar to the design of this active antenna as calculated via equation (2.14)

Equation (2.15) can be used with any active antenna for use in the HF band as long as the rod's terminating impedance is a very high resistance in shunt with a small value of capacitance and the shortening factor a, equation (2.14), has been calculated for that particular antenna design.

3. Calculating the Expected Output Noise

Calculation of the expected output noise will be done in three sections as depicted in figure 3.1. The EMF (ElectroMotive Force or open circuit voltage) associated with each noise generator will have the following designations.

> *E1* = noise voltage generated by the combined resistance appearing between the FET's gate and ground

> *E2* = noise voltage generated within the FET and,

> *E3* = noise voltage generated by the output buffer stage.

All three noise sources will have their magnitudes projected as they would appear at the terminated output of the active antenna and be represented as V1, V2 and V3 respectively. The internal noise voltage generated by a receiver directly connected to the output of the active antenna can be readily calculated from the manufacturer's specifications and added to the active antenna's internally generated noise to find the system's overall noise performance.
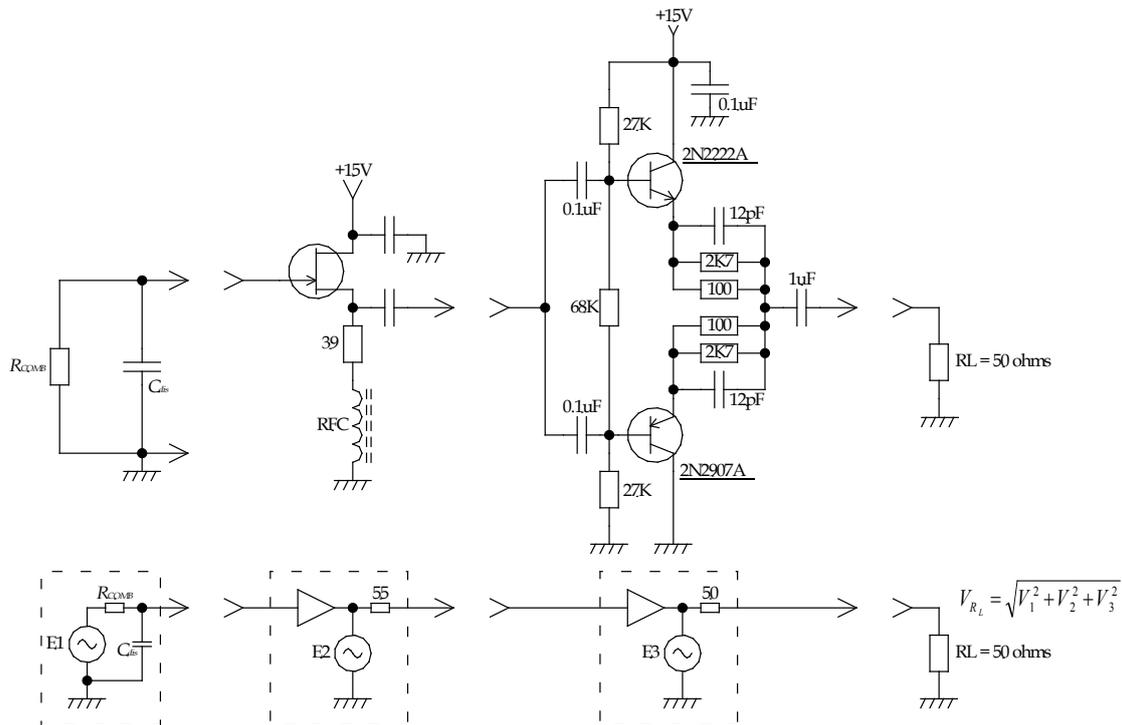

Figure 3.1   *The three sources of noise in the Active Antenna*

3.1 V_1 the noise generated by the circuitry connected to the JFET's gate

The open circuit noise voltage generated by the combined resistance connected between the JFET's gate and ground is [ref. 10, ch. 4]:

$$E_1 = \sqrt{4kTBR_{COMB}}$$

(3.1)

where: k = Boltzmann's constant = 1.380622E-23 Joules/Kelvin
T = 290 = absolute room temperature (K)
B = receiver bandwidth (Hz)
R_{COMB} = combined gate to ground resistances (Ω)

For the design being considered, R_{COMB} is = 4M7 $\|$ 2M2 $\|$ 4M7 = 1M13 ohms. The distributed capacitance to ground of the whip antenna plus the total input circuit capacitance form a frequency dependent voltage divider with the R_{COMB} reducing its internally generated open circuit noise voltage. In order to calculate the voltage drop across the capacitance, whose voltage drop is coupled to the gate of the FET, it is first necessary to know the impedance of the series $R_{COMB}C_{dis}$ circuit presents to the internal noise generator. The impedance of a series RC circuit can be calculated by equation (3.2).

$$Z_{RC} = \sqrt{R_{COMB}^2 + X_{dis}^2}$$

(3.2)

where: Z_{RC} = impedance of a series connected resistor and capacitor (Ω)
f = frequency (Hz)
R_{COMB} = 1M13 = combined gate to ground resistance (Ω)
C_{dis} = distributed rod and circuit capacitance (F)
$X_{dis} = \dfrac{1}{2\pi f C_{dis}}$

The ratio of voltage dropped across the capacitor becomes:

$$voltage\ ratio = \frac{1}{2\pi f C_{dis} Z_{RC}}$$

(3.3)

and multiplying this ratio by the measured forward voltage gain of the active antenna gives:

$$ratio = 0.47863 \frac{1}{2\pi f C_{dis} Z_{RC}}$$

(3.4)

where 0.47863 = $anti\,log\left(\dfrac{-6.4}{20}\right)$ = measured forward voltage gain of active antenna.

Combining equations (3.1) and (3.4) gives equation (3.5) for calculating V_1, the noise voltage generated by the $R_{COMB}C_{dis}$ circuit at the input of the FET which appears at the terminated output of the active antenna.

$$V_1 = 0.47863\sqrt{4kTBR_{COMB}} \left(\frac{1}{2\pi f C_{dis} Z_{RC}} \right)$$

(3.5)

C_{dis} is not easy to measure directly; the measuring instrument's reading rendered unstable by its own test signal being partially radiated and also corrupted by signals received by the length of the antenna rod. The capacitance can, however, be indirectly measured through resonating it with a known inductor. By placing a known inductor between the base of the rod and the antenna's metal case the antenna's input circuit will take the form of a Colpitts oscillator. If there is sufficient positive feedback then oscillation will occur, otherwise a pronounced sharp increase in the antenna's internally generated noise will be observed at the antenna's output. It is a simple matter to measure the frequency of the oscillation/sharp increase in noise at the output of the antenna using a spectrum analyser.

Figure 3.2 shows the active antenna's input circuit with the 7 uH inductor added. In figure 3.3(a) the input circuit has been simplified to only show those components that play a major role. C_{dis} in (3.5) is the active antenna's distributed input capacitance. It is comprised of C_a the rod capacitance, C_c the distributed wiring capacitance and finally C_{gs}, C_{gd} and C_{ds} are the JFET's internal distributed capacitances. These components have been rearranged in (b) to more easily identify the Colpitts oscillator configuration.
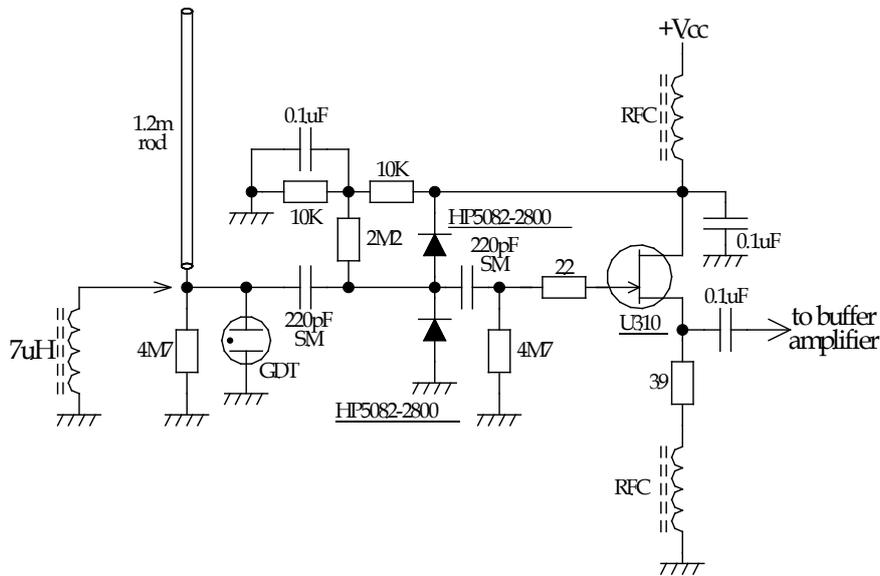

Figure 3.2   Input circuit of the Active Antenna with the 7 uH inductor added between the base of the rod and the antenna's outer metallic case
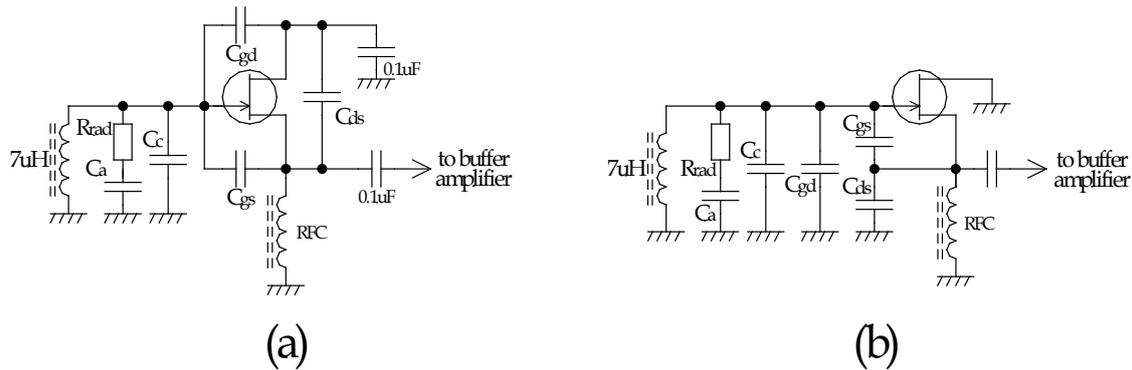
Figure 3.3   (a) The simplified input circuit, and (b) the components rearranged to highlight the Colpitts oscillator
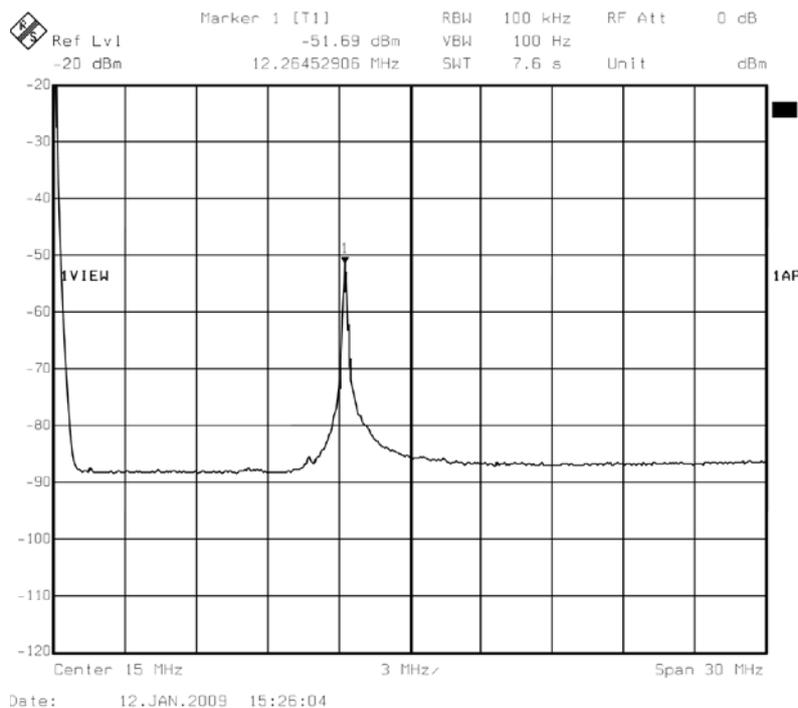
Figure 3.4   Measured peak in antenna noise floor at 12.26 MHz caused by the 7 uH inductor resonating with the antenna's distributed capacitance. +30 dB amplifier used between the antenna output and the analyser input.

The antenna with the 7 uH inductor was placed on an aluminium ground plain and installed onto the floor of a screened room. The screened room was necessary to prevent received HF signals from masking the expected peak in the antenna's noise floor. Figure 3.4 is a plot taken of the output of the active antenna; note the sharp increase in the noise floor seen at 12.26 MHz. The distributed capacitance was calculated using equation (3.6)

and found to be 24.1 pF. This result can be used in equation (3.5) to find V_1, the frequency dependent noise generated by the JFET's gate circuitry.

$$C_{dis} = \frac{1}{4\pi^2 f^2 L}$$

(3.6)

where: C_{dis} = 24.1E-12 = measured distributed rod and circuit capacitance (F)
f = resonant frequency (Hz) = 12.26 MHz from figure 3.4
L = known inductance = 7 uH

3.2 V_2, the noise internally generated by the JFET

The dominant source of noise in a common drain JFET is that generated by the drain/source equivalent noise resistance (r_n) [ref. 3, pg. 54]. Its open circuit noise voltage can be calculated by:

$$E_{o/c} = \sqrt{4kTBr_n}$$

(3.7)

$$r_n \approx \frac{0.67}{g_{fs}}$$

(3.8)

where: r_n = JFET's internal drain/source noise resistance (Ω)

This resistance should be as low as possible to minimise the magnitude of the noise voltage and is usually achieved by biasing the JFET close to saturation. Once biased, the g_{fs} can be calculated from a measurement made of the JFET's r_{ds} using a network analyser and manipulation of equation (2.10). In-house measurements made on a U310 JFET manufactured by Vishay and picked at random returned a g_{fs} value of 18.02 mS at 30 mA. This gives a value for r_n of 37.2 ohms.

The noise voltage generated by r_n is increased by the device's excess noise. Excess noise is usually cited by the manufacturer as the device's noise figure in dB (F_{dB} = 1.5 dB typical at 105 MHz for the U310 JFET, [ref. 8]). The manufacturer usually measures the noise figure when the JFET is conjugate matched to a 50 ohm system. The output impedance of the U310 JFET (r_{ds} is approx. 55 ohms) in the active antenna is not matched to the high input impedance of the buffer amplifier. This high input impedance of the buffer amplifier allows the JFET's output noise voltage to approach the FET's internal noise generator's open circuit voltage. Also, the gate terminal of the FET is terminated to a very high impedance. The manufacturer's cited noise figure for the U310 FET is no longer valid and needs to be measured under the conditions in which the FET is to be used. This measurement returned a value of 5.28 dB. The measured noise figure is first converted into a voltage ratio:
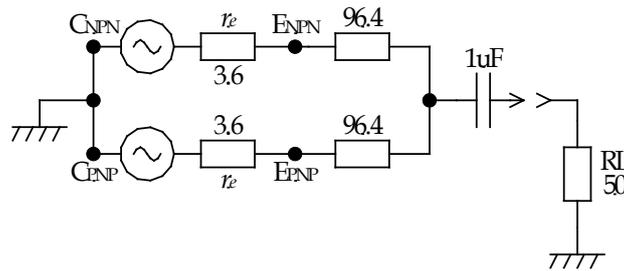
$$voltage \; ratio = anti\, log_{10}\left(\frac{F_{dB}}{20}\right)$$

(3.9)

then multiplied by the open circuit noise voltage of r_n. The open circuit noise voltage appearing at the source of a common drain U310 JFET is therefore:

$$E_{FET} = anti\, log_{10}\left(\frac{F_{dB}}{20}\right)\sqrt{4kTBr_n}$$

(3.10)

where: F_{dB} = 5.28 dB measured excess noise figure for the U310 JFET under conditions in which it is being used
            r_n = 37.2 ohms from equation (3.8)

3.2.1   The effect the input of the buffer amplifier on E_{FET}

Figure 3.5 is the schematic of the output buffer circuit. The drain/source resistance from the JFET has been calculated using equation (2.10) and found to be approximately 55 ohms and is used as the buffer amplifier's input termination. The output of the buffer amplifier is terminated with a load resistance of 50 ohms. With the input and output of the buffer amplifier correctly terminated, work can now begin calculating the sources of noise.

Figure 3.5   The buffer amplifier circuit

Figure 3.6   Simplified buffer amplifier output circuit, C is the collector and E the emitter

The 68K bias resistor in the input circuitry of the buffer amplifier has its noise shorted out by the two 0.1 μF coupling capacitors and hence plays no part in the noise calculations. The other two 27K bias resistors are effectively connected in parallel via the chassis and positive rail forming a single 13K5 equivalent resistor. The input resistance of the two individual transistors will be approximately equal and in parallel with each other. The manufacturer's data sheets for both transistors give their current gain (β) in the range of 100 to 300. A value of 150 will be used as a typical figure. The expected input resistance of each transistor is then:

$$r_{in} = \beta\left(R_E + r_e\right)$$

(3.11)

where: r_e = 3.6 = measured internal emitter resistance (Ω)
β = 150 = typical value of transistor current gain

R_E is best calculated with the aid of figure 3.6, a simplified schematic of the output stage in which E_{NPN} and E_{PNP} are the emitter and C_{NPN} and C_{PNP} are the collector leads of the respective transistors. Here it can be seen that R_E = 96.4 ohms + (the parallel combination of R_L (50 ohms) and the series connection of 96.4 ohms plus the 3.6 ohms of the other transistor's r_e) = 130 ohms. Applying this number for R_E in equation (3.11) returns a value of approximately 20K ohms. The inputs of the two transistors are connected in parallel giving a combined transistor input resistance of 10K ohms. The parallel combination of the 13K5 equivalent bias resistance and the 10K transistor input resistance gives a total input resistance for the buffer amplifier of 5K7 ohms. This calculation agrees with an earlier measurement carried out in section 2.3 using a network analyser which returned the parallel combination of 5700 ohms and 16 pF capacitance.

The following calculations are carried out to show the effects the JFET output impedance and the buffer input impedances have on each other. The excess noise generated by the buffer amplifier will not be considered at this stage.
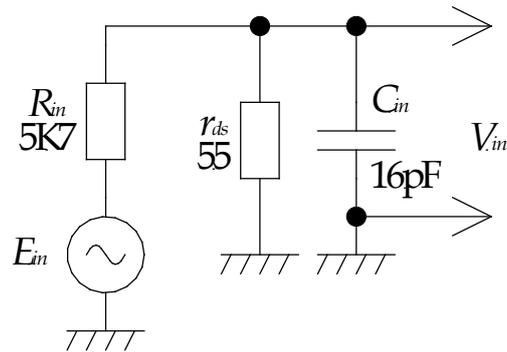
Figure 3.7   Noise generated by the effective input resistance of the buffer amplifier

The open circuit noise voltage generated by the 5K7 equivalent input resistance of the buffer stage is:

$$E_{in} = \sqrt{4kTBR_{in}}$$

(3.12)

where: Rin = 5K7 = total combined resistance seen at the input of the buffer (Ω)

The JFET's r_{ds} (equation (2.10)) and the input resistance and capacitance of the buffer form a divider network reducing the open circuit noise voltage of the 5K7 equivalent input resistor, see figure 3.7. The impedance presented by the parallel combination of r_{ds} and C_{in} is [ref. 15, pg. 28]:

$$Z_{rds\,/\!/\,Cin} = \frac{r_{ds}X_{in}}{\sqrt{r_{ds}^2 + X_{in}^2}}$$

(3.13)

$$X_{in} = \frac{1}{2\pi f C_{in}}$$

where r_{ds} = 55 ohms = internal resistance between the FET's drain and source
C_{in} = 16pF = measured buffer amplifier's distributed input capacitance
f = frequency (Hz)

and the voltage ratio resulting from the series connection of the 5K7 resistor and this impedance becomes:

$$voltage\ ratio = \frac{Z_{rds\,/\!/\,Cin}}{R_{in} + Z_{rds\,/\!/\,Cin}}$$

(3.14)

Combining equations (3.12) and (3.14)

$$V_{in} = \sqrt{4KTBR_{in}} \, \frac{Z_{rds \, // \, Cin}}{R_{in} + Z_{rds \, // \, Cin}}$$

(3.15)

The second source of noise and the most dominate of the two is that from the JFET. The simplified schematic is shown in figure 3.8.
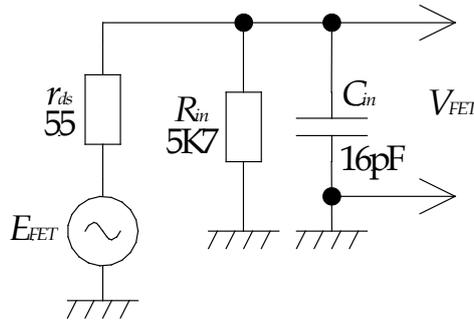

Figure 3.8   Noise generated by the FET appearing at the input of the buffer amplifier

The magnitude of E_{FET} has already been calculated via equation (3.10). In order to calculate the magnitude of V_{FET}, the voltage appearing at the input of the buffer amplifier, it is first necessary to know the impedance presented by the parallel combination of R_{in} and C_{in}. Equation (3.16) is used to calculate this impedance.

$$Z_{Rin \, // \, Cin} = \frac{R_{in} X_{in}}{\sqrt{R_{in}^2 + X_{in}^2}}$$

(3.16)

$$X_{in} = \frac{1}{2\pi f C_{in}}$$

and the voltage ratio resulting from the series connection of the 55 ohm r_{ds} resistor and this impedance becomes[2]:

$$voltage\ ratio = \frac{Z_{Rin \, // \, Cin}}{r_{ds} + Z_{Rin \, // \, Cin}}$$

(3.17)

combining equations (3.10) and (3.17) to obtain the magnitude of V_{FET}.

$$V_{FET} = anti\log_{10}\left(\frac{F_{dB}}{20}\right)\sqrt{4KTBr_n} \, \frac{Z_{Rin \, // \, Cin}}{r_{ds} + Z_{Rin \, // \, Cin}}$$

(3.18)

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[2] Note that r_{ds} is used in (3.17) as it represents the output resistance of the JFET and not r_n which is used to calculate the expected magnitude of the noise.

V_2 which appears at the terminated output of the active antenna is the summation of the two non-coherent noise sources multiplied by the gain of the output buffer amplifier, equation (3.19).

$$V_2 = anti\,log_{10}\left(\frac{G}{20}\right)\sqrt{V_{FET}^2 + V_{in}^2}$$

(3.19)

where G = -6.4 = measured forward gain of the active antenna (dB)

3.2.2   Justifying the fixed forward gain of -6.4 dB and the particular construction of the RFC

The JFET's internal r_{ds} resistance of 55 ohms and the JFET's load impedance described in equation (3.16) when used in the voltage gain equation (2.5) returns an expected voltage gain variation for the JFET of 0.974 at 3 MHz to 0.849 at 30 MHz. Multiplying these figures by the expected voltage gain of the output buffer stage of 0.5 gives an expected gain of -6.25 dB at 3 MHz and -7.77 dB at 30 MHz. Actual measurements of total forward gain for the active antenna are -6.1 dB at 3 MHz, -6.2 dB at 15 MHz and -6.3 dB at 30 MHz. A fixed forward gain of -6.4 dB was chosen to represent the worst case. The near constant forward gain is attributed to the 'inductive kick' associated with the use of the very large, low internal loss RFC seen in series with the JFET's 39 ohm source resistor. While it is tempting to replace this time consuming hand wound twin RFC with a single of-the-shelf perhaps even an SMD type inductor, it should be borne in mind that the very low Q of such inductors will negate the inductive kick effect and introduce the expected calculated gain with its 1.52 dB variation and not the measured 0.2 dB variation across the 3 MHz to 30 MHz band. When the high Q hand wound twin RFC is used the numerical voltage gain of the JFET is considered to be unity due to the 'inductive kick' and the voltage gain of the buffer amplifier is considered to be numerically 0.47863 or -6.4dB.

3.2.3 V_3 the noise generated by the output buffer stage active devices

No noise specification could be found in any of the various manufacturers' data sheets for the 2N2907A transistor. The noise generated by the buffer circuit will have to be directly measured so that its impact on the overall active antenna design can be assessed. Figure 3.9 is a set of plots of the noise output from the output buffer stage only as seen in figure 3.5.
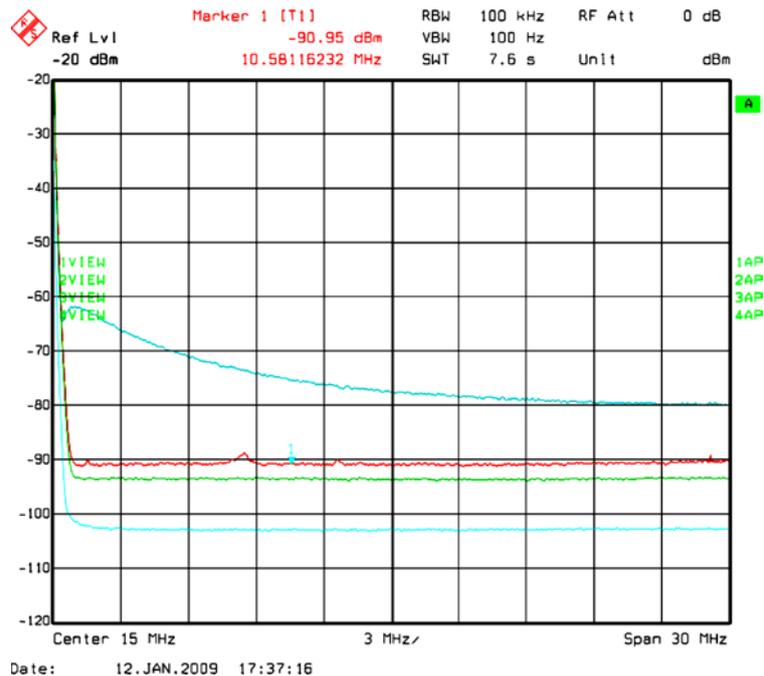
Figure 3.9   Measured noise output from the buffer stage only, +30 dB gain with a 1.5 dB noise figure post amplifier used on all but the Cyan trace.
Blue: input of buffer amplifier input unterminated.
Red: input to buffer amplifier input terminated with 55 ohms.
Green: +30 dB post amplifier only with a 50 ohm input termination.
Cyan: spectrum analyser input terminated with 50 ohms indicating its internal noise figure of 21 dB.

A +30 dB amplifier with a 1.5 db noise figure was used between the output of the buffer amp and the spectrum analyser. This was needed to overcome the measured 21 dB noise figure of the spectrum analyser. The blue (top) trace is the buffer amplifier with no input termination and shows the noise generated by the equivalent 5K7 ohm input resistance and the influence of the 16 pF distributed capacitance at its input[3]. The red trace with the marker at 10.58 MHz is with a 55 ohm resistor connected between the input and ground simulating the r_{ds} with no excess noise from the JFET. The noise generated by the buffer amplifier's input circuitry and the effects of the distributed input capacitance are swamped by the r_{ds} of 55 ohms. The marker at 10.58 MHz shows a noise level of -90.95 dBm for 100 KHz of bandwidth. The green and cyan traces are used to confirm the red and blue traces represent the unit under test and are not test equipment limited.

The circuit output noise voltage generated by the active devices used in the output buffer stage can be calculated by subtracting the forward gain of the test amplifier from the spectrum analyser's measured total output noise level thus referencing the total noise to

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[3] The blue trace in figure 3.9 is larger than what is predicted by equation (3.15). This has been attributed to the positive feedback via the capacitance between the transistor's emitter and base (aka, bootstrapping) when the input is unterminated.

the input of the test amplifier. Subtract from this the noise contribution of the test amplifier and the remainder is the noise from the active antenna, see equation (3.20).

$$V_3 = \sqrt{\left(0.22361\, anti\, log\left(\frac{dBm - G_{dB}}{20}\right)\right)^2 - \left(anti\, log\left(\frac{F_{dB}}{20}\right)0.5\sqrt{4kTBR}\right)^2}$$

(3.20)

where: V_3 = 108.4nV = rms noise voltage generated within the buffer amplifier
appearing at its terminated output (V)
dBm = -90.95 dBm for 100 KHz bandwidth, marker from figure 3.9
G_{dB} = +30 = post active antenna test amplifier gain (dB)
0.22361 = the terminated rms voltage for 0 dBm in a 50 ohm system (V)
F_{dB} = 1.5 = measured noise figure of +30dB test amplifier
k = Boltzmann's constant = 1.380622E-23 Joules/Kelvin
T = 290 = absolute room temperature (K)
B = 100KHz = receiver bandwidth from figure 3.9 (Hz)
R = 50 = termination resistance (Ω)

The left-hand-side of (3.20) is the total noise of both the active antenna and the test amplifier as measured at the output of the test amplifier converted to its equivalent at the test amplifier's input. The right-hand-side is the open circuit noise of a 50 ohm termination multiplied by 0.5 to convert it into its terminated form then multiplied by the numerical voltage ratio derived from the test amplifier's 1.5 dB measured noise figure. The total noise contribution generated internally within the active antenna to be added to the internal noise generated within any receiver that the active antenna is connected to therefore becomes:

$$V_{rms} = \sqrt{V_1^2 + V_2^2 + V_3^2 + \left(anti\, log\left(\frac{F_{dB}}{20}\right)0.5\sqrt{4kTBR}\right)^2}$$

(3.21)

where: V_1 = results of equation (3.5)
V_2 = results of equation (3.19)
V_3 = results of equation (3.20)
F_{dB} = noise figure of any receiver connected to output of active antenna (dB)
k = Boltzmann's constant = 1.380622E-23 Joules/Kelvin
T = 290 = absolute room temperature (K)
B = same bandwidth used to calculate V_1, V_2 and V_3 (Hz)
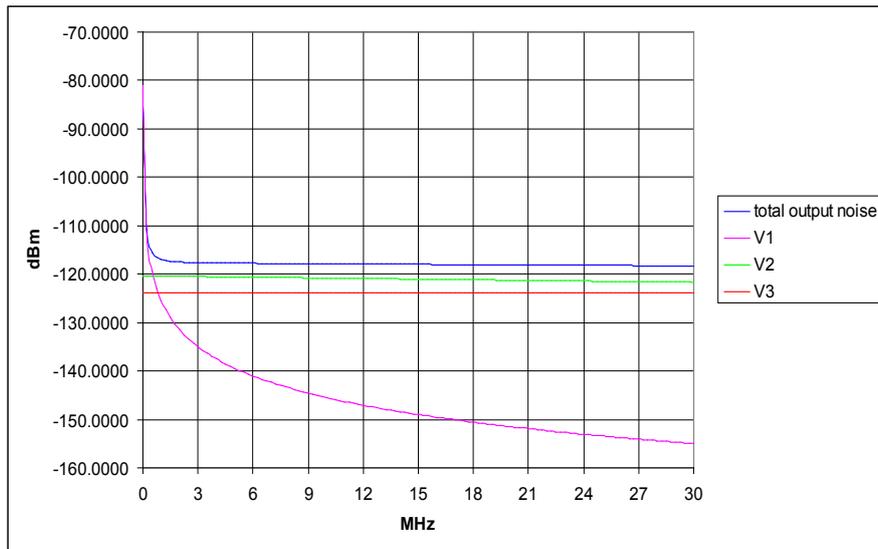R = 50 = termination resistance (Ω)

Figure 3.8 Blue: Calculated output noise of -118.0dBm at 15.5 MHz, 100 KHz bandwidth, 50 ohm termination, no post amplification.

Pink: V₁ noise from the parallel combination of 1M13 resistor and 24.1 pF distributed capacitance circuitry connected to the JFET's gate before the JFET's intrinsic noise is added. As with the blue trace, 100KHz bandwidth, 50 ohm termination, no post amplification.

Green: V₂ noise generated internally by the FET appearing at the output. As with the blue trace, 100KHz bandwidth, 50 ohm termination, no post amplification.

Red: V₃ noise generated by the buffer output stage. As with the blue trace, 100KHz bandwidth, 50 ohm termination, no post amplification.

The noise of the active antenna in isolation can be calculated by letting F_{dB} in equation (3.21) equal zero. This, however, is an approximation as any receiver that the active antenna is connected to will make a contribution to the total measured noise and influence the measured system performance.

Equations (3.5), (3.19), (3.20) and (3.21) were entered into an Excel spreadsheet to produce the plot as seen in figure 3.8. The blue trace is the expected output as seen on the spectrum analyser for the test setup. Notice it flattens at approximately 1 MHz and above. This is where the noise generated by the JFET's gate resistors, seen as the pink trace, is being attenuated with increasing frequency by the 24.1 pF distributed shunt capacitance and is where the noise generated by the JFET's internal r_{ds} and the output buffer stage predominates.

In figure 3.9 the +30 dB gain test amplifier with a 1.5 dB noise figure has also been added so a direct comparison can be made between calculated (figure 3.9) and measured (figure 3.10) noise in a system bandwidth of 100 KHz.
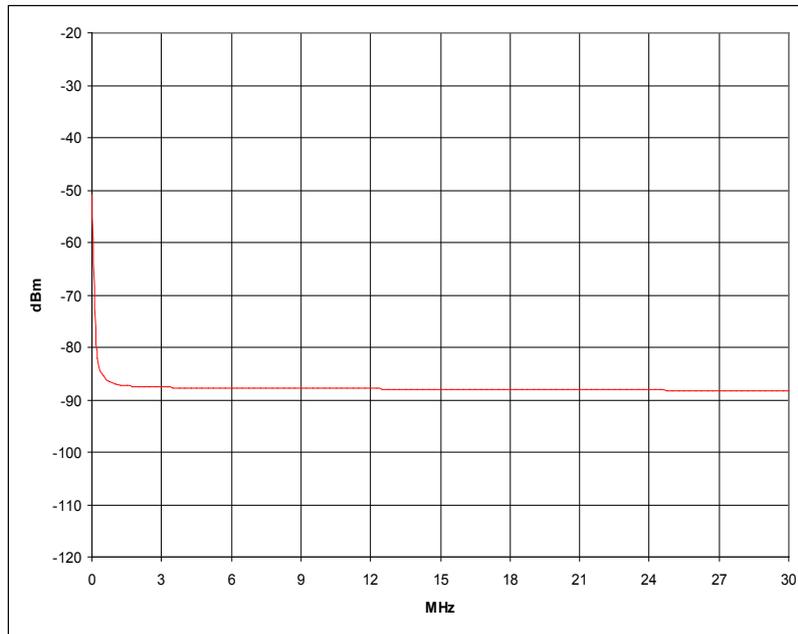
Figure 3.9   Calculated internally generated noise at -88.0 dBm @ 15.5 MHz appearing at the output of the Active Antenna, +30 dB post antenna gain with a 1.5 dB noise figure, 100 KHz bandwidth has been added.
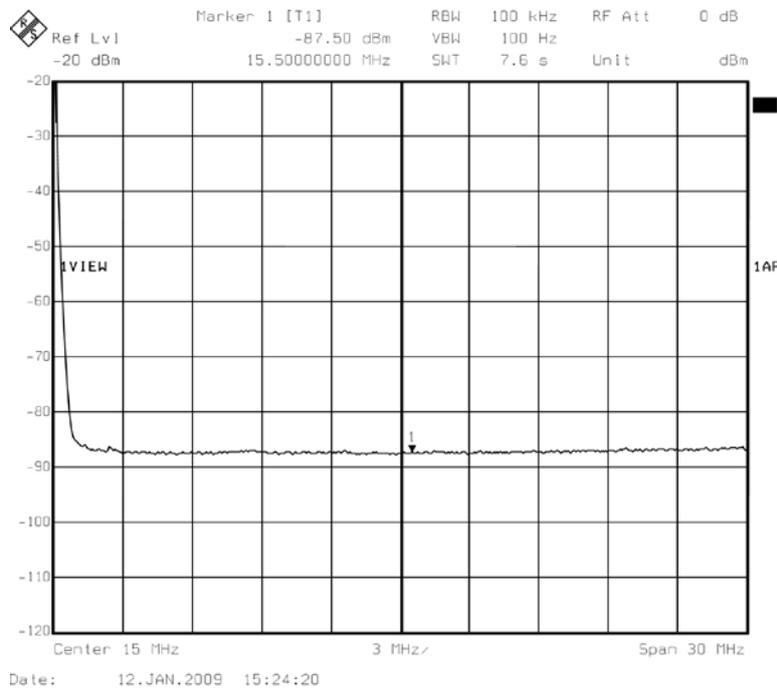

Figure 3.10   Measured internally generated noise (-87.5 dBm @ 15.5 MHz) appearing at the output of the Active Antenna, +30 dB post antenna gain with a 1.5 dB noise figure, 100 KHz bandwidth.

Figure 3.10 is the measured output noise generated by the active antenna when installed on a ground plain in a screened room to exclude external EM fields. Note the very close agreement with the red trace of figure 3.9.

3.3   Calculating the output excess noise

Excess noise[4] in a 50 ohm system is defined as noise above the thermal noise generated by a correctly terminated 50 ohm termination resistor at room temperature, usually 290 K. The excess noise internally generated by the active antenna's circuit appearing at its 50 ohm terminated output can be calculated by letting $F_{dB} = 0$ in equation (3.21). Using the marker from figure 3.10, equation (3.22) returns a measured excess noise figure of 6.5 dB.

$$F_{dB} = 174 + dBm_{ant} - 10Log_{10}B - G_{dB}$$

(3.22)

where: dBm_{ant} = calculated using equation (3.21) or read directly from the
spectrum analyser measurement of figure 3.10 (dBm)
G_{dB} = +30 dB = gain of low noise amplifier between output of active
antenna and input of spectrum analyser (dB)
B = same bandwidth used to calculate or directly measure dBm_{ant} (Hz)

Figure 3.11 is a graph showing the calculated excess noise versus frequency for the active antenna when connected to a broadband 50 ohm termination; no external EM fields are being received by the antenna rod.
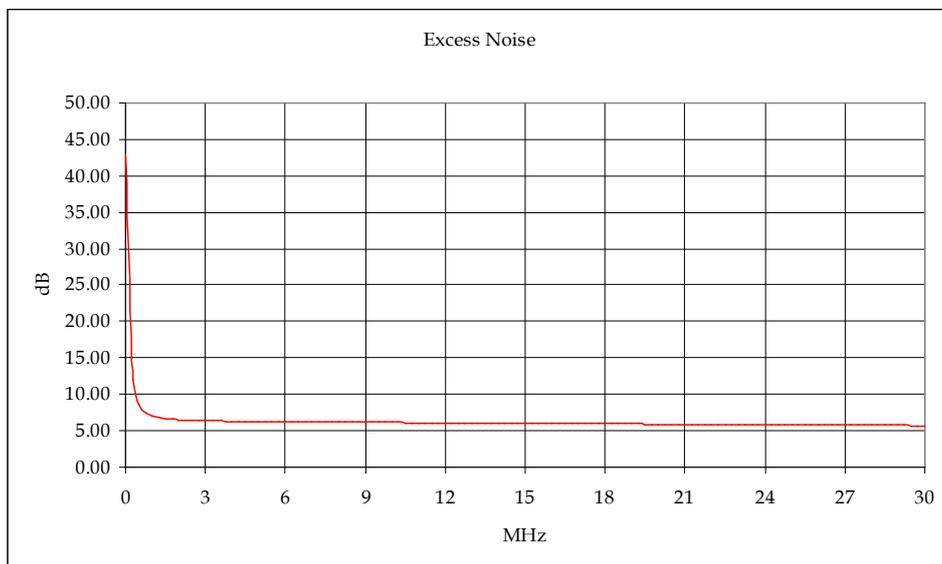

Figure 3.11   *Calculated excess noise of 6.01 dB at 15.5 MHz appearing at the terminated output of the active antenna. The 6.5 dB measured in figure 3.10 includes the effects of the 1.5 dB excess noise of the +30 dB test amplifier*

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[4] More commonly known as noise factor when expressed as a ratio or noise figure when in dB's.

The active antenna's calculated excess noise is in the vicinity of the measured noise figures of some commercially made HF receivers for amateur use. Military grade HF receivers usually have higher noise figures, in the range of 10 to 15 dB, with higher input intercept points. Some amplification between the active antenna's output and military grade receiver input may be necessary in quiet sites if the active antenna's sensitivity is to be maintained.

3.4 Summary of internal noise generated within the active antenna

Equations (3.5), (3.19), (3.20) and (3.21) were implemented into an EXCEL spreadsheet and the variables manipulated to find the main causes of internal noise generated by the active antenna's circuitry. The main findings are:

- The higher the effective total shunt resistance connected between the JFET's gate circuit and ground, the lower the frequency at which the noise generated by this resistance predominates for a given distributed shunt capacitance. For the current design this noise predominates below 500 KHz.

- The larger the distributed shunt capacitance of the JFET's gate circuitry the lower the frequency at which the noise generated by the shunt resistance predominates for a given shunt resistance. This shunt capacitance also includes the antenna rod capacitance to ground.

- Use a JFET that has the lowest r_{ds} keeping in mind that low r_{ds} generally means higher internal device capacitances, especially the JFET's gate to drain/source capacitance, which will shunt to ground the wanted RF signals. Biasing the JFET close to its saturation to lower the r_{ds} will also generate internal heat. A trade-off may be necessary between the heat generated and an acceptable value for r_{ds}. This is the dominant source of noise from 500 KHz and above.

- Use only metal film resistors in construction as these have the lowest internally generated excess noise.

- Use only **silver mica** capacitors for the series connected 220 pF capacitors in the gate circuit of the JFET. The very high impedances of the gate circuit make it sensitive to the noise generated by the high shunt resistance associated with the capacitor's dielectric. Some low pF ceramic capacitors have been found to be very noisy when used in this high impedance application.

4. Performance Testing

4.1 Frequency Response

Figure 4.1 (a) and (b) shows two possible methods of connecting the test tones from a signal source to the active antenna for the purpose of measuring frequency response and inter-modulation performance.
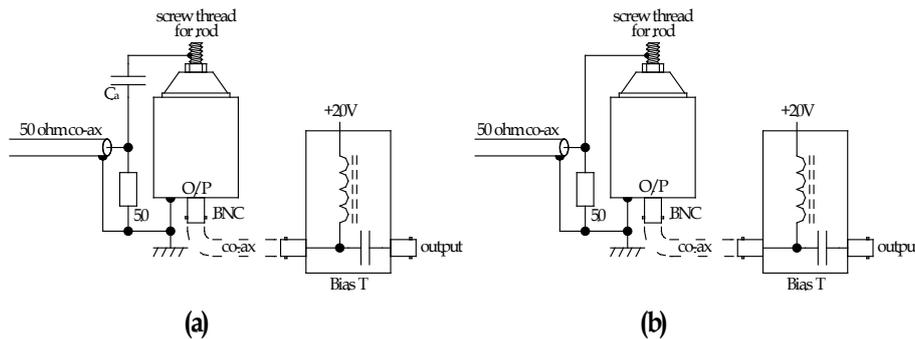

Figure 4.1   Two possible methods of connecting to the active antenna

Both methods were tried to see if there were any effects on the results. C_a, which represents the rod capacitance to ground, was calculated using equation (2.2) and found to be 13 pF at 3 MHz. Apart from the capacitive divider produced by C_a and the distributed input capacitance of the active antenna reducing the magnitude of the test signal to the input of the active antenna by approximately -6 dB, there was no difference in the measured output intercepts between the two methods. The test connection shown in (b) will be used as this setup did not need a high intercept point amplifier to boost the two tone test signal to the required level.

Figure 4.2 is a plot of the forward gain starting from 10 KHz to 100 MHZ. The vertical scale is 1 dB/division. The marker at approximately 7 MHz shows a forward gain of -6.46 dB.
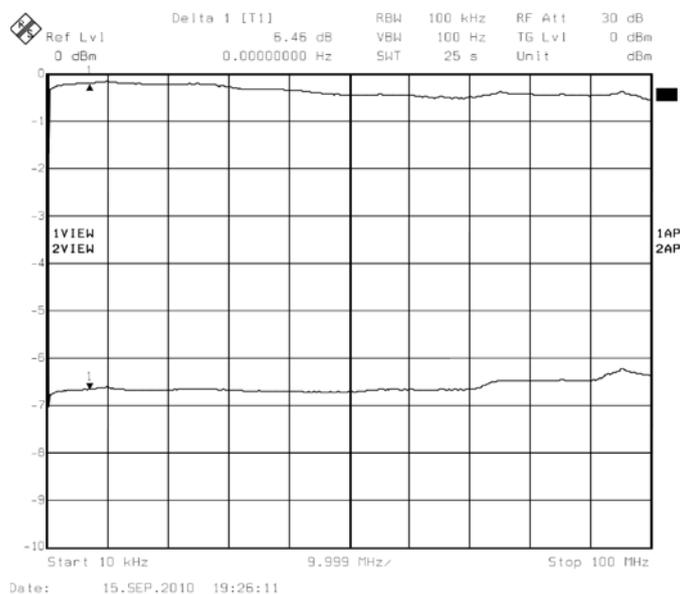
Figure 4.2　　*Top trace: through path loss of test set-up and reference from which gain is calculated*
Bottom trace: active antenna output showing a measured forward gain of -6.4 dB across the HF band

4.2　Measured Output Intercept Points

Figure 4.3 is the test setup used for the inter-modulation tests. Two non-harmonically related test tones of 8 MHz and 11 MHz were chosen. A 12 MHz lowpass filter attenuates the harmonics from the two tones. The 50 ohm termination at the input connection to the active antenna prevents any standing waves from developing in the coax between the output of the 12 MHz lowpass filter and the input of the active antenna. The 50 ohm termination also ensures that the 12 MHz lowpass filter is properly terminated. The 10 dB attenuator between the output of the antenna and the spectrum analyser is needed to provide a proper termination to the output of the active antenna when a filter is later inserted between it and the spectrum analyser when measuring the distortion products. A level of 0 dBm/tone at the output of the active antenna was for convenience in calculating intercept points. Figure 4.4 shows the test systems measured output level of -10 dBm/tone (0 dBm/tone at the output of the active antenna). The displayed distortion products at 3 MHz and 5 MHz are generated internally by the spectrum analyser and are not from the active antenna.

A 6.5 MHz 7th order lowpass filter was inserted between the 10 dB attenuator and the spectrum analyser to reduce the level of the two tones without affecting the level of the distortion products generated by the active antenna, see figure 4.5. A check was made by adding extra internal attenuation within the analyser to confirm the displayed distortion products at 3 MHz and 5 MHz were not generated or contributed to by the spectrum analyser itself.
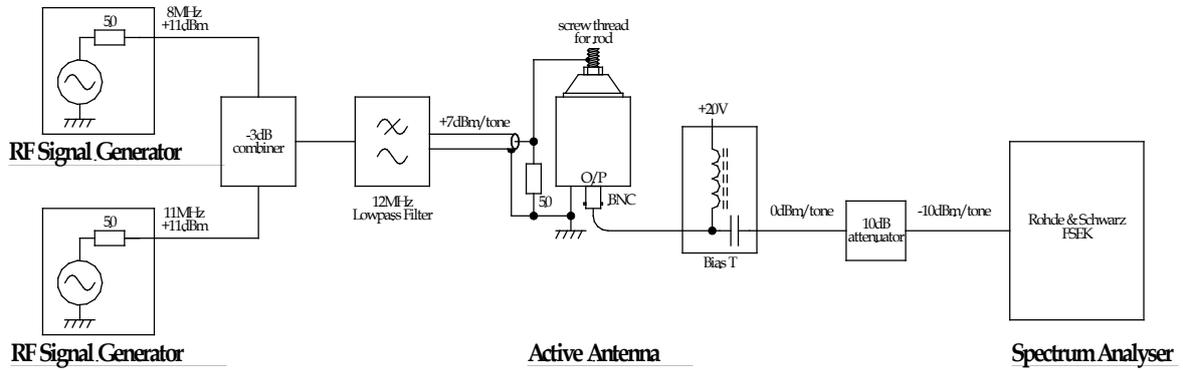
Figure 4.3   Test set-up for measuring the levels of the 8 MHz and 11 MHz two tone test signal
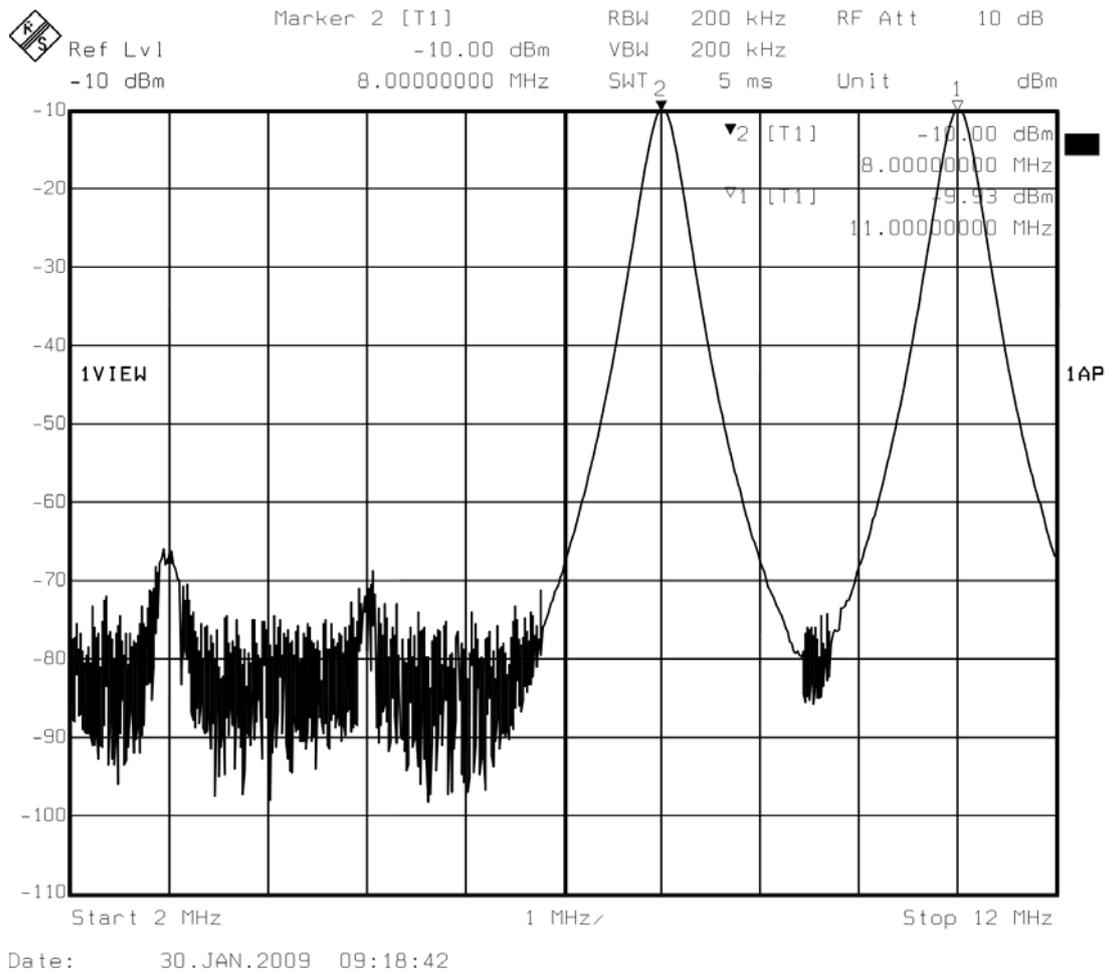

Figure 4.4   Two tone test, 8 MHz and 11 MHz, 10 dB external attenuation used
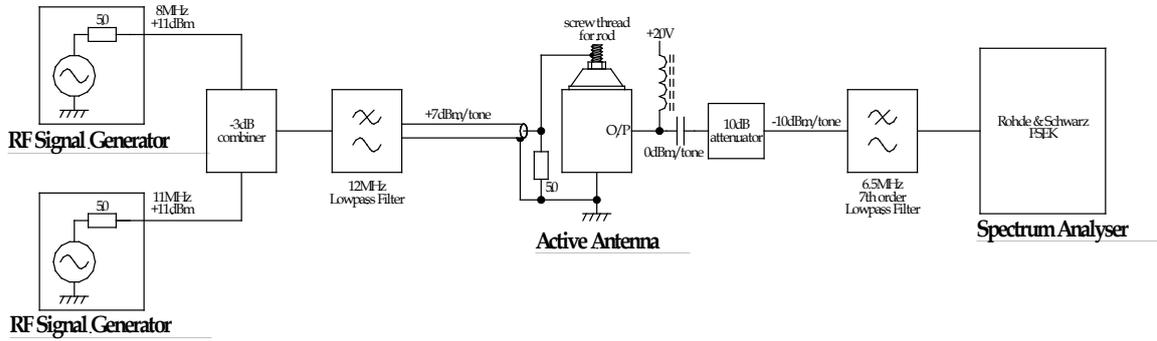
Figure 4.5   *8 MHz and 11 MHz two tone test set-up for measuring the levels of the inter-modulation products*
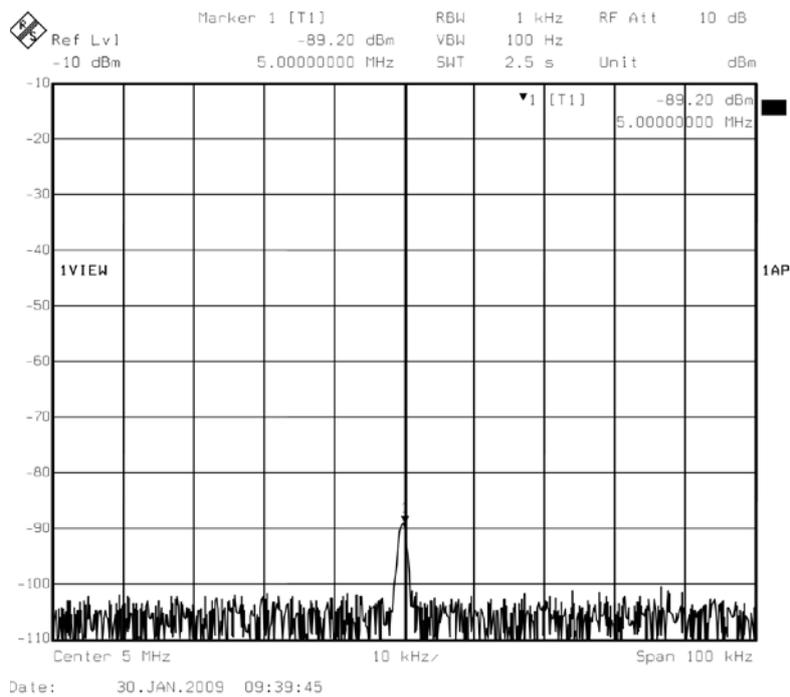

Figure 4.6   *Third order tone at 5 MHz, 10 dB external attenuation. Marker at -89.2 dBm (-79.2 dBm at output of antenna) equates to an OPIp3 = 39.6 dBm*
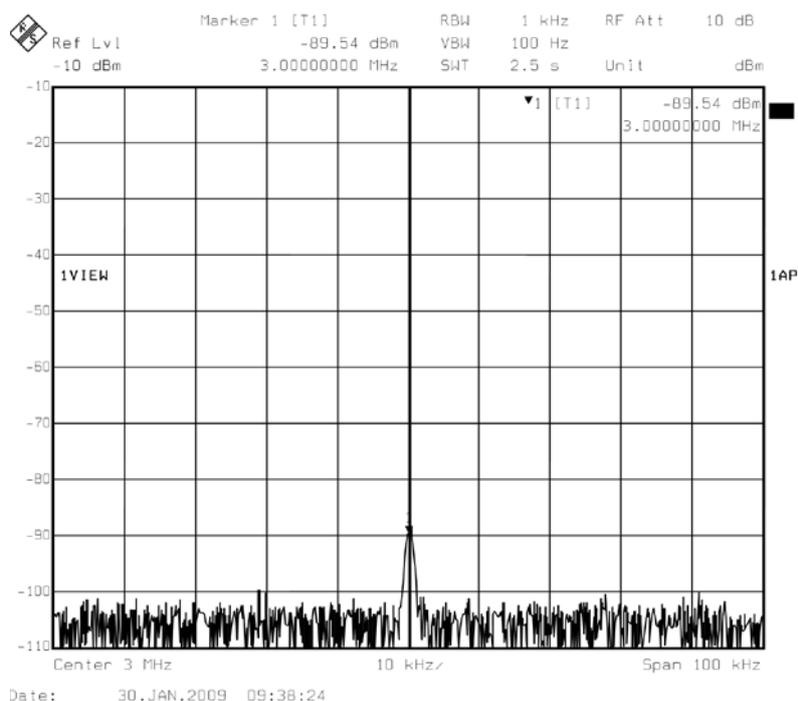
Figure 4.7   Second order tone at 3 MHz, 10 dB external attenuation. Marker at -89.54 dBm (-79.54 dBm at output of antenna) equates to an OPIp2 = 79.5 dBm

Figure 4.6 shows a measured output level of -89.2 dBm (-79.2 dBm at the output of the active antenna) for the third order product. The intercept point can be calculated by the general formula [ref. 5, pg. 103]:

$$Ip_n = P_o + \left(\frac{P_o - P_I}{n-1} \right)$$

(4.1)

where: Ip_n = nth order output intercept point (dBm)
P_o = output level of two equal tones (dBm)
P_I = output level of IMD product of nth order (dBm)
n = order of distortion product, 2 for second order, 3 for third order

With a two tone level of 0 dBm and the measured third order product of -79 dBm equation (4.1) returns a value of OPIp3 of +39.6 dBm. Likewise, figure 4.7 shows a measured second order distortion product at 3 MHz of -89.5 dBm (-79.5 dBm at the output of the active antenna). Applying equation (4.1) to this second order product returns a measured OPIp2 of +79.5 dBm.

4.3   Calculated and Measured Antenna Factor

The antenna rod was removed from its mount on the active antenna and the active antenna's input capacitance was resonated with a 7 uH inductor connected between the

screw thread of the antenna mount and the metal case. The active antenna's circuit oscillated at 16.028 MHz. Equation (3.6) was used to calculate the input capacitance C_{in} of 14 pF. The antenna rod is 1.2 m long and 2.45 mm in diameter at its base. Equation (2.2) was used to calculate an expected C_a of 13 pF. These two capacitances form a capacitive divider to the voltage generated in the rod by the E field reducing the rod's open circuit voltage by a factor of 0.48. The forward voltage gain of the active antenna was measured using the setup of figure 4.1(b) and found to be 0.479. Total voltage gain between the open circuit voltage generated in the rod to that appearing at the terminated output of the active antenna is 0.48 x 0.479 = 0.23. The calculated antenna factor (k) is therefore:

$$k = \frac{E}{V_{ter}} \quad = \frac{1}{0.23} \quad = 4.35$$

(4.2)

and expressed in decibels:

$$K = 20 \log k \quad = 12.8 dB$$

(4.3)

The active antenna was installed in an antenna test site where an E field generated from a distant vertical monopole was calculated using the response of a Rohde & Schwarz HE010 active antenna and the manufacturer's cited typical antenna factor of 17 dB for that antenna. The calculated E field strengths were then used in conjunction with the terminated output voltage of the active antenna under test to confirm the antenna factor calculation. Figure 4.8 is the measured antenna factor calculated from the collected data. There is close agreement between the calculated and measured antenna factors.
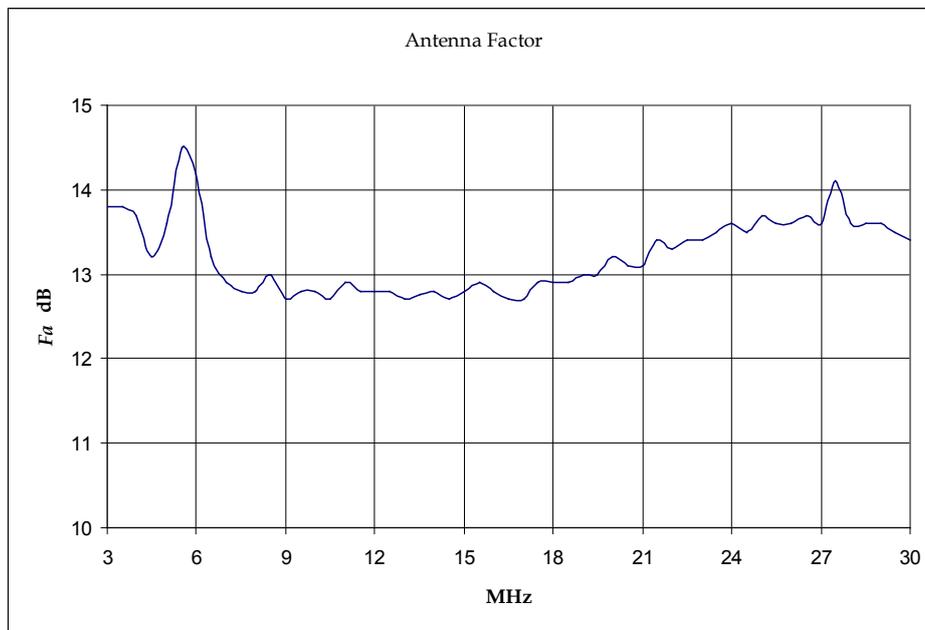

Figure 4.8   *Measured antenna factor*

4.4   Calculating the Active Antenna's Noise Figure

Reference 6 (page 3) gives the following equation for calculating the vertical r.m.s. noise field strength in dB(μV/m) for a short vertical monopole above a perfect ground plane.

$$E_n = F_a + 20\log f_{MHz} + 10\log B - 95.5 \quad \text{dB}(\mu\text{V/m}) \tag{4.4}$$

The equivalent noise field strength needed to be equal to the internal noise generated by the active antenna can be calculated by rearranging equation (4.4) to solve for F_a :

$$F_a = E_n - 20\log f_{MHz} - 10\log B + 95.5 \quad \text{dB} \tag{4.5}$$

E_n can be found by calculating the terminated noise generated by a standard 50 ohm resistor, expressed in dB(μV), then adding to it the active antenna's excess noise figure and the active antenna's antenna factor (K).

$$E_n = 20\,log\left(500000\sqrt{4kTBR}\ \right) + F_{dB} + K \quad \text{dB}(\mu\text{V/m}) \tag{4.6}$$

where: F_a = active antenna noise figure (dB)
f_{MHz} = centre frequency (MHz)
k = Boltzmann's constant = 1.380622E-23 Joules/Kelvin
T = 290 = absolute room temperature (K)
R = termination resistance = 50 (ohms)
B = bandwidth in (Hz)
F_{dB} = excess noise figure from equation (3.22) (dB)
K = antenna factor from equation (4.3) (dB)
500000 = 1 000 000/2 = multiply by 1 000 000 to convert from volts to μV
then divide by 2 to convert from μV open circuit to μV terminated

Equations (4.5) and (4.6) were entered into an EXCEL spreadsheet and the active antenna's noise figure plotted, see figure 4.9.
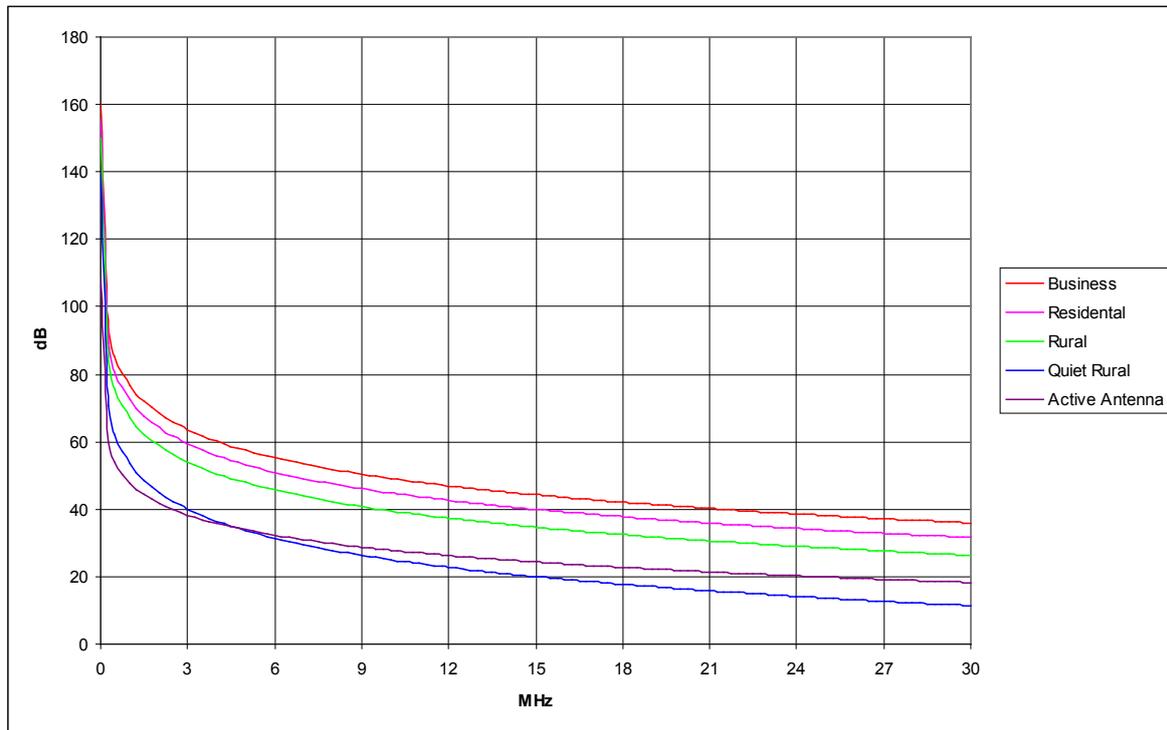
Figure 4.9   The calculated antenna noise figure for the active antenna with a 1.2 m rod 2.45 mm diameter and the expected values of man-made noise for comparison

The ITU report [ref. 6, pg. 10 and pg. 14] gives the following equation and table for calculating the expected man-made noise levels for European receive sites in the HF band.

$$F_{am} = c - d \log f$$

(4.7)

where F_{am} = median value of man-made noise power in the 0.3 to 250 MHz frequency range for a short vertical lossless grounded monopole antenna (dB)
f = frequency (MHz)
c and d = as per *Table I*

Table I values for the constants c and d obtained from [*ref. 6*] *to be used in equation (4.7)*

Environmental Category	c	d
Business	76.8	27.7
Residential	72.5	27.7
Rural	67.2	27.7
Quiet Rural	53.6	28.6

Noise levels in rural Australia have been measured and found to be as much as 10 to 15dB lower than the values of equation (4.7) [14, pg. 12], however, the ITU report is regularly

cited in reports and texts and seems to be the reference to which comparisons are made. Expected noise levels for the four European environmental sites are also plotted in figure 4.9 for comparison.

4.5   Spurious Free Dynamic Range (SFDR)

The SFDR for an active antenna can be calculated by using the frequency dependent noise figure and the input intercept points. The input intercept point for the active antenna is:

$$\overset{n}{I} = OPIp^n + K$$
(4.8)

where: $OPIp^3$ = +39.6 dBm = measured output intercept point, figure 4.6
$OPIp^2$ = +79.5 dBm = measured output intercept point, figure 4.7
K = antenna factor (in dB) from equation (4.3) = 12.8 dB

The second and third order SFDR can be calculated by [5, pg. 108]:

$$3rd\ order\ SFDR = \frac{2}{3}\left(\overset{3}{I} - kT/Hz - 10\log B - F_a \right)$$
(4.9)

$$2nd\ order\ SFDR = 0.5\left(\overset{2}{I} - kT/Hz - 10\log B - F_a \right)$$
(4.10)

where: $SFDR$ = spurious free dynamic range (dB)

$\overset{3}{I}$ = system third order input intercept point (dBm)

$\overset{2}{I}$ = system second order input intercept point (dBm)

kT/Hz = thermal noise in a 1Hz bandwidth
        = -174 dBm for a 50 ohm system
B = bandwidth (Hz)
F_a = noise figure, from equation (4.5) (dB)

Figure 4.10 shows the results of equations (4.9) and (4.10) for a 3 KHz communication system bandwidth.
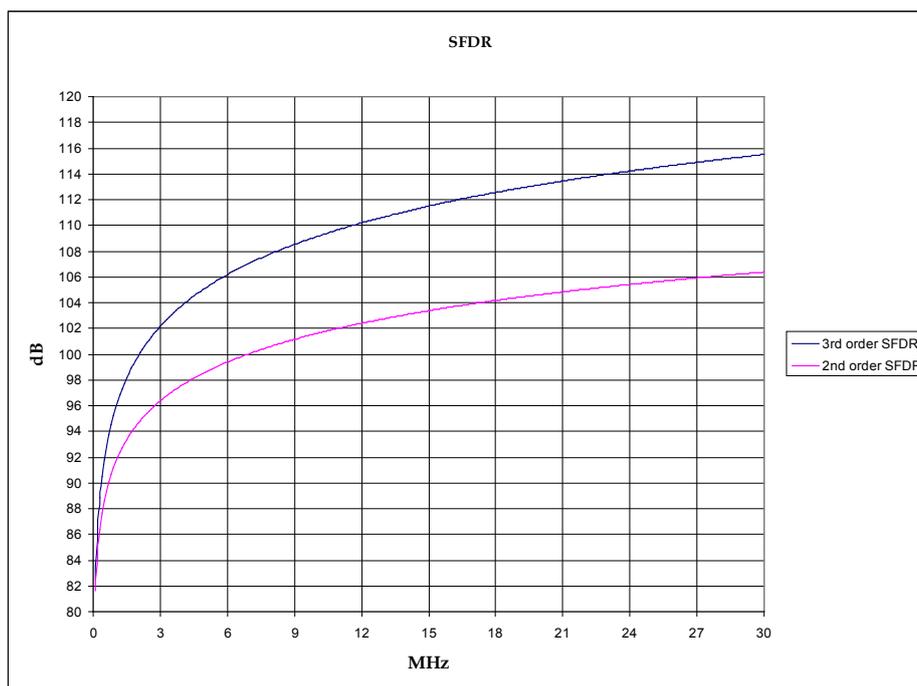
Figure 4.10   Calculated 3rd and 2nd order SFDR for a 3 KHz system bandwidth

The data of figure 4.10 can be converted into an *E* field strength/tone measured in volts/metre by multiplying the terminated noise voltage generated in a 50 ohm resistor by the voltage ratio of both the calculated antenna noise figure F_a and the *SFDR*, equation (4.11).

$$V / m = 0.5\sqrt{4kTBR} \quad anti\,log_{10}\left(\frac{F_a}{20}\right) \quad anti\,log_{10}\left(\frac{SFDR}{20}\right)$$

(4.11)

where: k = Boltzmann's constant = 1.380622E-23 Joules/Kelvin
T = 290 = absolute room temperature (K)
B = 3000 = typical bandwidth for a HF communication system (Hz)
R = 50 = active antenna termination resistance (Ω)
F_a = active antenna noise figure, from equation (4.5) (dB)
SFDR = spurious free dynamic range from equation (4.9) or (4.10) (dB)

The results of equation (4.11) are displayed in figure 4.11. This is the level that two equal strength *E* fields will have to be in order to generate a spurious product at the same voltage level as the antenna's internally generated noise floor.
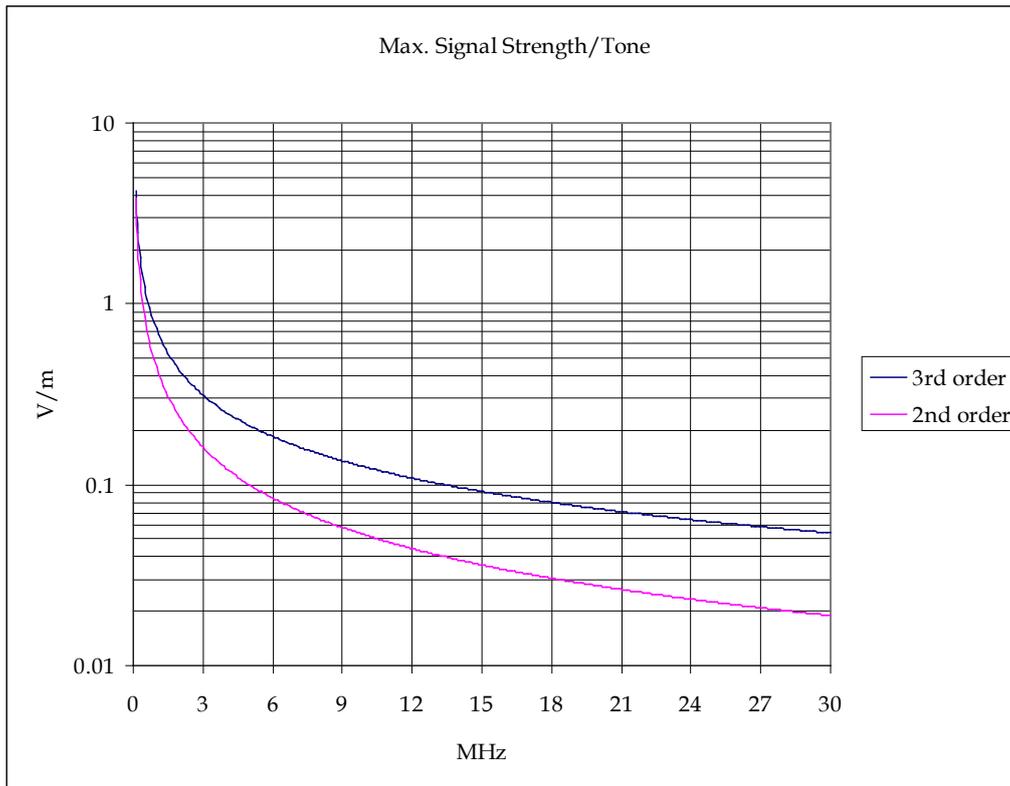
Figure 4.11   Calculated strengths of two equal magnitude E fields that would create a spurious product at the same voltage level as the antenna's internally generated noise floor

4.6   In Summary

The design of a high performance HF active antenna has been presented. The main source of inter-modulation products has been identified and steps taken to reduce their magnitudes are described and implemented. The internal noise generated within the design, which can limit the antenna's sensitivity, was also examined in detail. The effect of near-field external noise on the lower frequencies of the HF band was also shown and methods to reduce their influence via the use of RF current chokes presented.

Recommendations were given on how to install these types of antennas in an antenna site with the resonance formed by the mast together with the active antenna readily calculated. Operating an active antenna at frequencies higher than this resonance will result in a monotonic decrease in sensitivity with increasing frequency.

The antenna factor was calculated from first principles and confirmed via a field experiment. The antenna factor was then used to calculate the active antenna's noise figure thus allowing comparisons to be made with the expected levels of man-made noise cited in an ITU report.

The completed design is a compact receive only active antenna covering the frequency range from LF to HF. It is broadband with a constant antenna factor over this frequency

range. The presented active antenna can be used to measure the vertical E field strength of a signal, perform general surveillance work (either fixed or vehicle mounted) or used to create a quickly deployable antenna array for the geo-location of signals in the HF band.

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for the High Frequency Band

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17. ABSTRACT The design of an active antenna with an operational frequency range from low- to high-frequency is presented. The main cause of inter-modulation distortion is identified and recommendations are given to minimize their generation. A detailed analysis is carried out into the various sources of noise generated within the active antenna and how their summation affects the total noise seen at the output. Lightning protection is also discussed.			