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**Development of a pulse
modulated microwave power
flux density measurement
system and the evaluation of
field strength probes with
pulse modulated fields.**

K P Holland

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Development of a pulse modulated microwave power flux density
measurement system and the evaluation of field strength probes
with pulse modulated fields

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ABSTRACT

This report describes the development of a measurement system to calibrate power flux density and field strength probes for use in pulsed applications such as radar. The additional uncertainties due to the measurement of the peak and average power have been analysed and the system has been used to measure the performance of typical probes. Those probes employing thermocouple detection showed very good performance, as did a frequency selective probe. A diode based probe, although exhibiting large variations in sensitivity depending on the pulse parameters and applied level, can now be corrected by calibration at NPL. In common with the more expensive frequency selective probe, its fast response and peak-hold facility offers scope for the measurement of rotating radars.

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1 Introduction

The interest in measurement of Power Flux Density and Field strength has grown over the recent past with the public concern over exposure to electromagnetic radiation and the publication of the Stewart report [1] on mobile phones and health. The assumption that effects upon humans are most likely to be thermal in nature has prompted health organisations such as Health Protection Agency (HPA) in the UK, and the World Health Organisation (WHO) internationally, to specify safety exposure levels for fields to which people can be exposed. The International Commission for Non Ionising Radiation Protection (ICNIRP) is the formally recognized non-governmental organization in non-ionising radiation protection for the WHO, the International Labour Organization (ILO), and the European Union (EU). Since 2004, the HPA [2] recommended that the UK should adopt the 1999 ICNIRP guidelines [3].

The European EMF Directive [4] adopts the safety levels as recommended by ICNIRP and becomes UK law on 30 April 2008. Therefore, a need exists for a means of traceably measuring these fields.

This project recognised that PFD probes are currently only calibrated in a continuous wave (CW) environment but in most cases are used to measure modulated signals, multiple frequency signals, or pulsed fields.

Pulsed fields from radar installations have been recognised by the major automotive manufacturers, Ford and GM, as having the most potential for disrupting electronic control systems and have imposed strict EMC testing of all prototype vehicles and ancillary equipment using pulsed signals.

The ICNIRP guidelines make 27 references to pulsed fields and require that for frequencies exceeding 10 MHz the peak power flux density as averaged over the pulse width should never exceed 1,000 times the reference levels. Thus for duty cycles of 0.1% or less, which is quite normal for radar type pulses, the peak limit is reached before the average reference level. The reasoning behind this peak limit is to prevent any auditory response [5] and is in addition to the accepted thermal hazard on the basis that any perceived response to EM fields is potentially hazardous and should be avoided. It is important therefore that a calibration system is able to directly measure the peak field intensity rather than relying on a calculation based on the duty cycle and the average value.

It has been established through work undertaken in the 2000 – 2003 Electrical Programme on power metrology for wireless communications that it is not necessarily valid to assume that the response of an instrument to pulsed or modulated signals can be inferred from the response to continuous wave signals. Radiation hazard probes currently available utilise two types of detection systems and it is already well known and documented that those employing diode sensors can indicate a much higher or a much lower field strength than is actually present when non-CW signals are present.

The work described in this report builds upon the existing NMS facilities and has successfully lead to a pulse power measurement capability within the lifetime of the programme. One of the important aims of the project was to evaluate a selection of probes that are typically used to measure in pulsed environments, in order to determine the most suitable subset of measurements to enable a meaningful calibration with sufficient accuracy for a given environment.

2 Pulse Definitions

Continuous Wave RF is uninterrupted RF of constant amplitude and frequency, such as that from an oscillator. Amplitude modulated (AM), frequency modulated (FM), and phase modulated (PM) signals contain additional frequency components but the RF is continuously present. Pulsed RF, on the other hand, is bursts of RF with no RF present between bursts. The most general case of pulsed RF consists of pulses of a fixed pulse width or (τ), which come at a fixed time interval, or period, (T). For clarity and ease of this discussion, it is assumed that all RF pulses in a pulse train have the same amplitude. Pulses at a fixed interval of time arrive at a rate or frequency referred to as the pulse repetition frequency (PRF) of so many pulses per second. Pulse repetition interval (PRI) and PRF are reciprocals of each other.

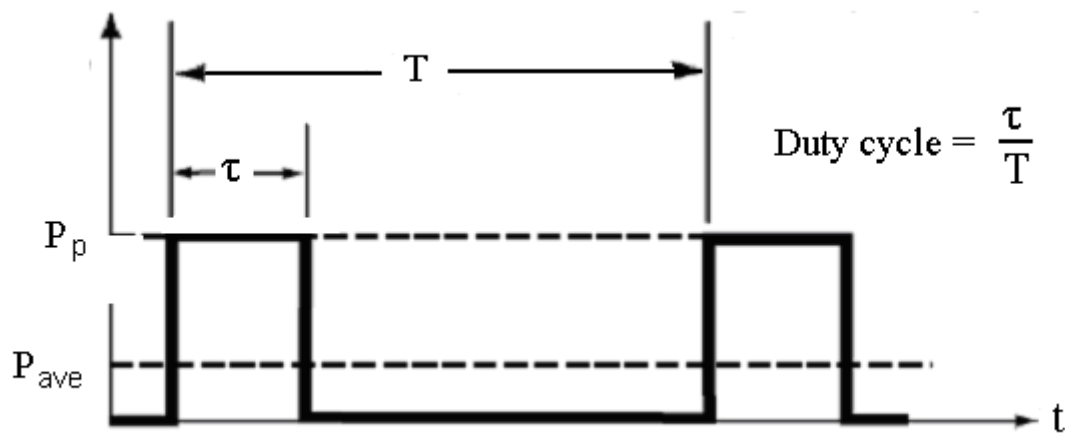


Figure 1 Parameters of a pulse

Power measurements are classified as peak power, P_p , or average power, P_{ave} . Most power measurement methods measure the heating effects of the RF energy to obtain an average power but a new generation of sampling power sensors can provide both peak and average power. The Duty Cycle is τ/T and for perfectly rectangular pulse shapes is the same ratio as P_{ave}/P_p . Practical pulses have finite rise and fall times and overshoot, making them far from rectangular so in order to measure the duty cycle accurately the pulse width is defined as the time interval between the 50% amplitude point on the rising edge to the 50% amplitude point on the falling edge, where the amplitude is the Average Pulse Power. The Average Pulse Power (also known as Pulse Power), is defined as the power computed by using the statistical data from pulse-top amplitude power and time measurements. The term Peak Pulse Power is defined as the highest point of power in the pulse envelope, usually at the first overshoot; it might also occur elsewhere across the pulse top if parasitic oscillations exist. These definitions are from the IEEE Standard [6].

Figure 1 shows an ideal pulse train. The pulse width is 1 unit of time and the period is 4 units. In this case the Duty Cycle is $\tau/T = 1/4 = 0.25$ or 25%.

A more typical case for radar would be a PRF of 1,000 and a pulse width of 1.0 microsecond. The Duty Cycle is 0.001 or 0.1%.

3 Overview of existing systems for calibration of PFD and field strength probes

PFD and Field Strength probes are primarily designed to have a similar sensitivity over a wide range of frequencies and levels. This is achieved in most cases by using antennas that are much smaller than a wavelength at their highest operating frequency. This also achieves another important requirement for field probes in that they couple loosely with the field thus causing minimal field perturbations. They are not designed to be calculable but are adjusted to read true within stated limits by a calibration process. Their wide frequency range (10 Hz to 50 GHz) precludes a single calibration system. In general, calibration below 300 MHz is carried out in guided wave enclosures known as Transverse Electromagnetic (TEM) Cells [7] and at higher frequencies using anechoic enclosures and calibrated antennas.

3.1 TEM Cells

Crawford TEM Cells are, in principle, coaxial lines expanded by input and output tapers to form a rectangular outer shielding conductor with a flattened inner conductor called the septum. They are operated in the transverse electromagnetic mode, so that both the E-field and H-field components generated between the septum and outer conductor have the characteristics of a wave propagating in free space. The field strength can be calculated from the dimensions of the cell, its impedance at the measurement plane and the input power. TEM cells have an upper frequency limit, determined by the size of the cell, below which, the fields set up in the central region are essentially uniform which makes them ideal for calibrating probes. However, once the upper frequency limit is exceeded, higher order modes can propagate, which give rise to large spatial variations in the fields. For electrically small probes the NPL Tapered TEM cell can be used up to 2.5 GHz. The Tapered cell is equivalent to the input taper of a Crawford type cell, but is terminated with microwave absorber. The accurately made tapered profile, ensures that almost all of the input power goes into the TEM mode, so that errors associated with higher order modes are kept to a minimum.

3.2 Anechoic Enclosures

For the highest accuracy, a large, fully lined, anechoic chamber is required above 300 MHz, and at this frequency, it can be practically realised in terms of the physical size and cost of the facility. An anechoic chamber is a room lined with pyramidal shaped radio absorbing material (RAM). The size of the RAM is proportional to the largest wavelength at which it will efficiently absorb. In a true reflection free environment, a calculable, uniform field can be generated using a calibrated coaxial or waveguide coupler system feeding a horn antenna. Knowing the power fed into the antenna and its gain, one can readily calculate the power flux density and hence field strength at a known distance from the horn antenna. The antenna, the coupler and the power sensor have limited bandwidths and calibration parameters that depend on the frequency of operation. A range of antennas and couplers are therefore required and need to be calibrated at a sufficient number of points to enable interpolation to be carried out with minimal additional uncertainty. A system block diagram is shown in Figure 2.

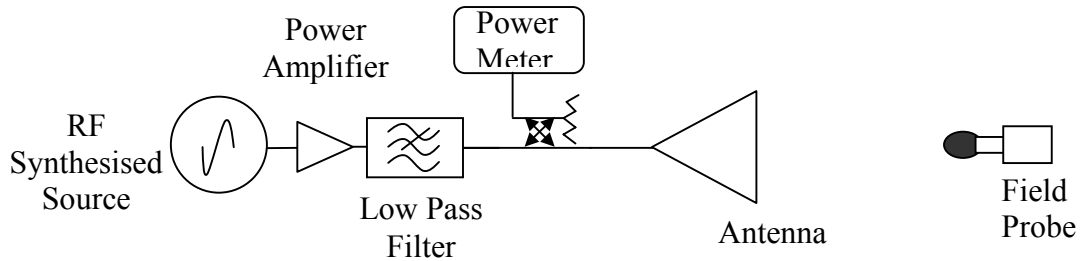


Figure 2 Probe calibration in an anechoic enclosure

The generating and power monitoring equipment and the antenna are mounted on a carriage, which moves along the long axis of the rectangular chamber to set the required distance or range between the antenna and the probe for calibration. The antenna is calibrated in a much larger anechoic chamber at NPL using the three-antenna gain extrapolation method [8,9]. This method can also accurately predict gains at finite distances. Finite range gain is suitable for small receiving apertures such as field probes and enables higher field levels to be generated at distances of less than 1 metre.

4 Design of a Pulsed System

In order to extend the measurement system for pulsed operation additional equipment was required and the existing equipment was assessed for the effect of pulse modulation.

4.1 Amplifiers

Existing CW amplifiers could be used provided a signal generator was available with pulse modulation capability. However the peak power would be limited to the maximum CW power and for typical radar pulses, this would not be sufficient to provide realistic average powers. Specialised broadband pulse amplifiers are available that provide high peak powers but they have a limit on the average powers, allowing a maximum duty cycle of around 4-6 %. Radar frequencies range from 300 MHz to 40 GHz, so in order to cover a significant number of radar applications two pulse travelling wave tube amplifiers (TWTA) having a nominal peak power of 1.5 kW were obtained. The maximum output power delivered to the antenna was measured and the power flux density and field strength calculated for distances of 0.75, 1.5 and 3 metres. The results for 0.75 metres are given in Table 1.

Table 1
Maximum peak and average levels for three duty cycles calculated
at a distance of 0.75 metres.

Frequency GHz	Peak power Flux Density W/m ²	Average Power Flux Density W/m ²			Peak Field Strength V/m	Average Field Strength V/m		
		Duty Cycles				Duty Cycles		
		4% (max)	1.0%	0.1%		4% (max)	1.0%	0.1%
2.5	8996	360	90	9	1842	368	184	58
3	10145	406	101	10	1956	391	196	62
4	13878	555	139	14	2287	457	229	72
5	7519	301	75	8	1684	337	168	53
6	7304	292	73	7	1659	332	166	52
7	4119	165	41	4	1246	249	125	39
8	4582	183	46	5	1314	263	131	42
9	6318	253	63	6	1543	309	154	49
10	5783	231	58	6	1477	295	148	47
11	8336	333	83	8	1773	355	177	56
12	10226	409	102	10	1963	393	196	62
13	6660	266	67	7	1585	317	158	50
14	8049	322	80	8	1742	348	174	55
15	8231	329	82	8	1762	352	176	56
16	8548	342	85	9	1795	359	180	57
17	8832	353	88	9	1825	365	182	58
18	8777	351	88	9	1819	364	182	58

The ICNIRP reference levels from 2 – 300 GHz are given in Table 2. Note that the peak level is specified as the peak power flux density as averaged over the pulse width. This level is reached before the average level only where the duty cycle is less than 0.1%. Over most of the frequency range greater than 50% of the peak reference level for the General Public and greater than 10% of the levels for Occupational Exposure can be achieved.

Table 2 The ICNIRP reference levels from 2 – 300 GHz

ICNIRP Standard	Average		Peak	
	W/m ²	V/m	W/m ²	V/m
Occupational	50	137	50000	4384
General Public	10	61	10000	1952

Higher field levels are possible by reducing the distance between the probe and the antenna. The power density increases as the decrease in distance squared but the effective gain of the antenna reduces. By measuring at a lower level at a calibrated distance and measuring at the same indicated level at a reduced distance the effective reduction in antenna gain can be calculated.

The amplifiers were also tested for pulse response and signal purity. Harmonics of the carrier frequency are removed using low pass filters frequency matched to the waveguide sizes. The amplifiers have a low-level sampled output, which was connected to a Hewlett Packard coaxial detector type 33330C. The rise time of the

detector is given as 4 ns when connected to an Agilent 54622A sampling oscilloscope specified as having a 50 ohm, 14 pF input. The rise time of the oscilloscope is given as 3.5 ns and the sampling period is 5 ns with a horizontal resolution of 3 ns per division. Using this system, rise and fall times of greater than 5 ns can be measured. The pulse output from the pulse generator has a minimum transition time of 5 ns but is adjustable up to 200 nS. The amplifiers modulate a continuous wave input from the generator by applying a positive going DC pulse to the grid modulator input of the TWT. The rise time of the pulse output from the TWT was found to be 50 ns and the pulse top showed amplitude spikes when the amplifier was not driven into saturation. These spikes had a 5 μ s period (200 kHz) and were identified by the manufacturer as emanating from the switched mode power supplies. As the spikes are not coherent with the PRF, the power meter will correctly average these during the measurement of the average power, but in order to measure the appropriate peak power in accordance with ICNIRP reference levels it was necessary to compute the Average Pulse Power. It was found that the amplitude of the modulating pulse at the input of the TWT amplifier affected the relationship between the input pulse and the output pulse envelope. Too high an input resulted in significant overshoot on the rising edge of the output pulse and reduced the delay between the rising edges of the input and output pulse. This delay is an important parameter in obtaining correct average pulse power measurements. The amplifiers also have a delay between the falling edge of the modulating pulse and that of the output. This effectively stretches the pulse width. As it is important to be able to program the pulse width to better than 1% this must also be taken into account during the measurement set-up. The optimum drive level for both amplifiers was 1.3 V, the leading edge delay was typically 750 ns and the falling edge delay was typically 200 nS. Note that since the completion of this project, the manufacturer of the TWT amplifiers has indicated that the pulse overshoot can be removed by modifying the grid modulator circuit.

4.2 Power Measurement

Thermocouple and diode sensors have been used in the present PFD calibration facility. Power sensors are designed to measure power available to a load matched to the characteristic impedance of the transmission line. Thermal sensors measure the temperature rise due to the power absorbed in a resistive load. Diode sensors convert the alternating voltage across a transmission line to steady DC voltage.

4.2.1. Thermal Sensors

The thermocouple sensor converts the minute temperature rise of a matched RF load into a DC voltage using the Seebeck effect. The modern thermocouple power sensor combines semiconductor and microwave thin-film technologies to form a highly sensitive averaging detector with an inherent response time of 120 μ S. However, further averaging time is needed to produce accurate measurements especially down to their lowest measurement level of 0.3 μ W (– 30 dBm). Because of their inherent ability to sense power with true square-law characteristics, thermocouple sensors will always be best for handling signals with complex modulations or multiple tones. Provided that the response time is significantly longer than the longest PRI or modulation period, they always respond to the true average power of a repetitive waveform, whether modulated, pulsed or composed of multiple signals. They are also rugged, stable, and reliable. Thermocouple sensors have been compared with thermistor sensors and have been found to be linear over a 30 dB dynamic range to

better than 1%. For the measurement of pulsed power, the relatively long time constant requires many cycles of the pulse frequency to reach a stable reading and the pulse repetition interval must be much less than the response time. The thermal power sensor has no way of directly measuring the peak amplitude of the signal so care must be taken to ensure that high peak powers do not damage any part or the sensor. Thermal sensors may not be sufficiently sensitive for low duty cycles where average power is very small. To achieve an average power of 1 mW at a duty cycle of 1/1000 would require a peak power of 1 Watt. If a 100 mW sensor was used and the pulse width was set incorrectly then the sensor could be easily be overloaded. The dynamic range of the thermocouple sensor depends on the duty cycle as well as the pulse amplitude.

4.2.2. Diode Power Sensors

Low-barrier Schottky diodes (LBSD) are used to achieve very wide power ranges from -70 dBm (100 pW) to -20 dBm (10 μ W) from 10 MHz to 18 GHz.

Mathematically, a detecting diode obeys the diode equation [10]

$$i = I_s (e^{av} - 1) \quad \text{Equation 1}$$

where $a = q/nKT$, i is the diode current, v is the net voltage across the diode, I_s is the saturation current and is constant at a given temperature. K is Boltzmann's constant, T is absolute temperature, q is the charge of an electron and n is a correction constant to fit experimental data.

Equation 1 can be expressed as a power series to better analyze the rectifying action for small signals.

$$i = I_s \left(av + \frac{(av)^2}{2!} + \frac{(av)^3}{3!} + \dots \right)$$

Equation 2

The second and other even-order terms of this series provide the rectification. For a range of small signals, only the second-order term is significant so the diode is said to be operating in the square-law region. In that region, the output is proportional to the RF input voltage squared. A full analysis of the large signal response has been described [11] which shows that Equation 2 is no longer valid and under restricted operating conditions the diode detection law progresses through the quasi-square-law region, or transition region and then moves into the linear detection region (output voltage proportional to input voltage). The square law region extends to an input voltage of 22 mV, which is equivalent to a power level into 50 ohms of -20 dBm. The latest generation of diode detectors can be made with transfer characteristics that are highly stable and reproducible so that corrections for the transition region and into the linear region can be made within the sensor instrumentation, thus enabling the full 90 dB dynamic range to be realized.

It is important to distinguish between averaging diode sensors and diode sensors that are suitable for pulsed signals. All diode sensors output a single polarity voltage in response to the instantaneous voltage amplitude on the transmission line, but in order to calculate the average power flow in the transmission line the relationship between this voltage and the instantaneous power flow must be known. Also the averaging must be applied over a time period greater than the lowest modulation period or about five times the filter time constant. In this way all diode sensors operating in their square-law region will accurately measure the average power regardless of the waveform provided that the peak voltage remains within the square law region.

The new generation of wide dynamic power sensors correct for errors in their non square-law region by using the average detected voltage and a lookup table. Corrections are applied assuming that the waveform is sinusoidal because the relationship between the output voltage and the average voltage for a sinusoidal signal is constant. For modulated, multiple frequencies and pulsed signals this relationship is not known and therefore this type of power sensor is unsuitable unless the peak voltage remains in the inherent square-law region. Newer diode sensors, [10] employ two switchable diode stacks that keep to the square law over an 80 dB range to $\pm 3.5\%$. However the range switching level (-10 dBm) can only be determined using the average power, which can result in the incorrect range being selected for low duty cycles where the average power is correct for the low range, but the peak power requires the higher range.

4.2.3. Diode Pulse Power Sensors

Pulse Power Sensors take the correction process one stage further in that they apply a correction for the non square-law errors based on each instantaneous sample of the transmission line voltage. The detected (video) is digitized at a rate of 20 million samples/second and a digital signal processor (DSP) corrects for linearity based on the amplitude, temperature and frequency response using a three dimensional look-up table stored in a programmable memory located in the sensor housing. The DSP applies averaging over user defined time intervals (known as gates), relative to a trigger pulse derived from the video or supplied externally. The DSP then calculates the peak and average level and the peak to average ratio over each gated period. The ability to measure peak power, average pulse power and average power makes the pulse power sensor ideal for the accurate measurement of radar type pulses.

4.3 Application of the Pulse Power Sensor

A choice was made to use the Agilent E9327A because at the time, it had the highest video bandwidth (>5 MHz) available and the fastest sampling DSP. It operates over the frequency range 50 MHz to 18 GHz and the correctable dynamic range for the peak power at this bandwidth is specified as -40 to +20 dBm. To avoid burn out of the diode detector the peak power must not exceed +30 dBm for >10 μ s and the average power must never exceed +23 dBm. The calibration accuracy is specified as $\pm 2.7\%$ and the linearity, after correction, from -40 to 0 dBm is 4% and 2% from 0 to +20 dBm.

4.3.1. Linearity

The Power sensor linearity was assessed by comparison with a known thermocouple power meter at 5 and 18 GHz. The frequency of operation was entered into the pulse power meter, as this was needed to enable the appropriate linearity correction from the sensor EPROM. The correction based on temperature and the power level is then

also made automatically. The pulsed sensor was set to the ‘normal’ pulsed mode but the signal was continuous wave to enable the most stable comparison of the two power sensors. The pulsed sensor was set to measure the average power over a 100 μ s gated period with continuous free run triggering. The final average power was obtained by taking the arithmetic mean of 256-gated periods. The output of a signal generator was split two ways using a matched power splitter each arm feeding a power sensor. The generator power was varied between indicated readings on the pulsed sensor of 0.05 to 100 mW. The linearity of the thermocouple sensor was known to better than 0.2%. Figure 3 shows that the linearity relative to 1 mW indicated on the pulse power sensor is within $\pm 1.5\%$, which is well within the manufacturer’s specification.

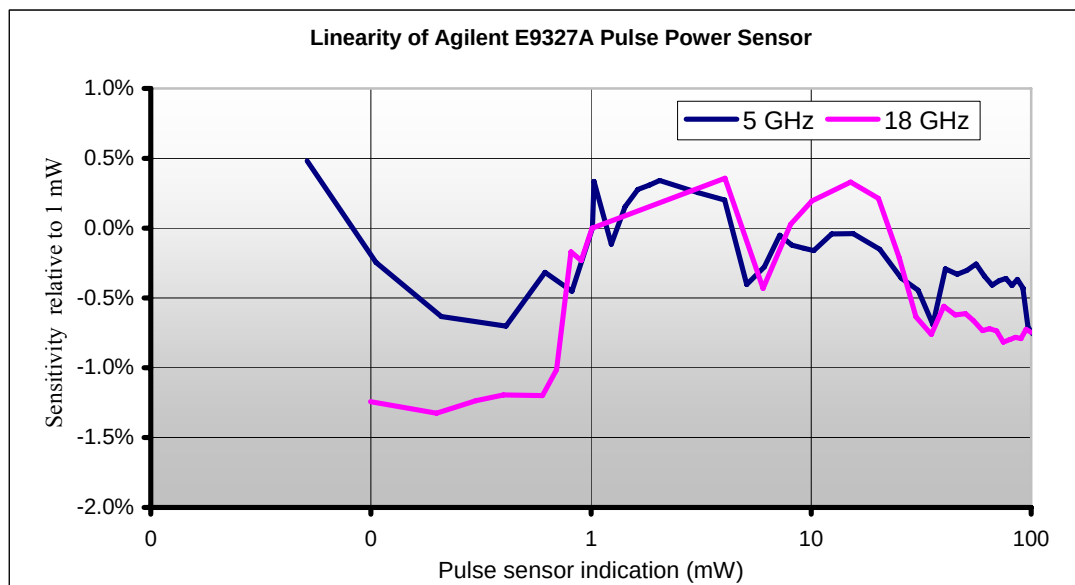


Figure 3

Linearity of Agilent E9327A Pulse Power Sensor measured against a Thermocouple Power Sensor

4.3.2. Sensor Bandwidth

The pulse power sensor has four video bandwidths. The pass band shapes achieved by three of the video bandwidth settings are flat with very sharp cutoffs to ensure accurate power measurement within the specified band. A fourth, ‘Filter Off’, provides approximately 3dB roll off at the maximum sensor bandwidth (5 MHz) and is best suited for accurate trace capture, removing any ringing effects caused by the sharp cut-off filters used in the Low, Med and High settings.

4.3.3. Measurement of Peak and Average Power

Pulsed signals are often defined in terms of Duty Cycle, Pulse Width and Pulse Repetition Frequency. Field probes almost always display average Power Flux Density levels or RMS Field Strengths, and so the independent variable for calibration should be based on average values of indicated or actual field strength. The relationship between these parameters only holds if the peak field level refers to

the average pulse power, which is the pulse top amplitude averaged over the pulse width. Figure 4 shows the measurement gate timing required to measure the average pulse power and average power of a practical pulsed waveform. T_p is the trigger point provided by the pulse generator and is very close to the rising edge of the output pulse. This is not a reliable reference point for measurement because very small delay variation will result in significant changes in the sampled amplitude. Although the power meter continuously stores data in a buffer memory, it is the measurement trigger that starts the process of retaining and analysing the data during the gated periods. The timing of the gates are relative to the trigger event and setting a negative trigger delay allows a time reference and data to be used before the measurement trigger occurred. $T_p - T_D$ gives the trigger delay, which is matched to the amplifier or signal generator if a CW amplifier is in use. The start of Gate1 (G1) is set to be well ahead of the rising edge and the duration of Gate1 is set to the pulse repetition interval. Gate2 is set to start 100 ns (5 samples) from the 50% rise time point and the duration of Gate2 is set to 200 ns less than the pulse width. The average of each time-gated data set provides the pulse power and average power measurements required for the calibration system.

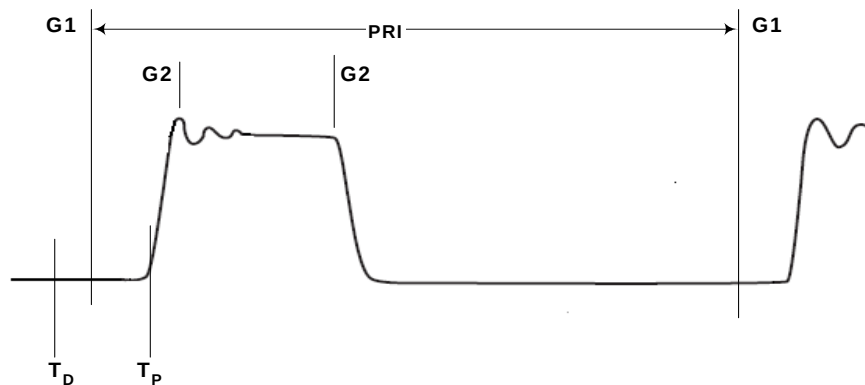


Figure 4 Measurement timing of pulse envelope

4.3.4. Pulse width measurement

The E4417 power meter has a memory for each gated period, which stores the instantaneous power readings from the pulse sensor, at a maximum rate of 20 Msamples/sec (50 ns samples). The memory size is 200,000, which corresponds to a 10 ms period. This would capture one cycle of a pulse train with a PRF of 100 Hz, which is more than is required for radar. By setting a longer capture period the sampling rate is reduced. Setting a shorter period results in less stored points, because the sample rate cannot be increased beyond 20 MS/s. The memory size can be manually decreased to 1000 points for faster data transmission.

A software routine was written to enable the extraction of the data from the DSP to allow external computation of the average power and average pulse power (the average power during the pulse). The pulse repetition frequency and pulse width are also determined to provide real-time verification of the transmitted pulse characteristics. The routine requires the precise PRF, the DSP sampling rate and the approximate pulse width to be input prior to the computation. The sampling rate of the power meter was verified by setting the power meter to output its internal clock

from the Trigger output BNC connector on the back panel. The specification is 10 MHz $\pm$ 0.01% and this was confirmed by measurement.

The average power is simply the sum of the instantaneous power samples divided by the number of samples in the Pulse Repetition Interval (PRI). The pulse width is the interval between the 50% amplitude points of the rising and falling edges measured from the average pulse power.

The program calculates the average pulse power and pulse width using successive approximation. The peak power is obtained by finding the maximum power sample over the PRI and this is used to provide a first estimate of the 50% pulse height of the rising edge and falling edges. The pulse width is then the number of samples multiplied by the sample period before the amplitude falls to 50% or less. Having established the data points corresponding to the rising and falling edge the average pulse power is obtained by averaging the power samples over the period between the 50% rising and falling edges. The average pulse power is then used as the new reference for the 50% amplitude points to determine the pulse width more accurately.

The program was used to measure a pulsed waveform having a PRF of 1 – 10 kHz and pulse width of 1-10 μ S. Differences between the average powers measured directly by the internal averaging routine and the external software was less than 0.2%. The accuracy of the pulse width was similar except for narrow pulses where there are insufficient samples over the pulse width. A 1 μ s pulse contains 5 samples and unless the first sample is perfectly synchronised with the 50% point on the rising edge of the pulse then an error of $\pm$ 1 sample (20%) could occur in the measurement of pulse width. For this reason, the pulse width can be verified more accurately using an oscilloscope. The data transfer to the computer took between 10 and 40 seconds depending on the PRI. This would considerably slow down the measurement system so having verified the power meter's internal power averaging routines, these were used in the calibration system. The pulse width is set from the pulse generator, corrected for amplifier delays and verified using an oscilloscope.

The data from the DSP memory is returned in IEEE 488.2 arbitrary block program data and a special conversion routine in Visual Basic was developed to enable the fast transfer of data to a computer. The verification program for the conversion process is given in Appendix 1.

4.4 Pulse Bandwidth

A true continuous wave (CW) occupies zero bandwidth and frequency dependencies of measuring equipment are removed by calibrating at sufficient number of frequencies to enable interpolation. A pulsed modulated carrier such as radar introduces a spread of energy in the form of a series of discrete frequencies, harmonically related to the pulse repetition frequency, either side of the carrier frequency. The discrete frequencies occupy 'sidebands' which have an amplitude envelope with nulls and peaks according to the pulse width.

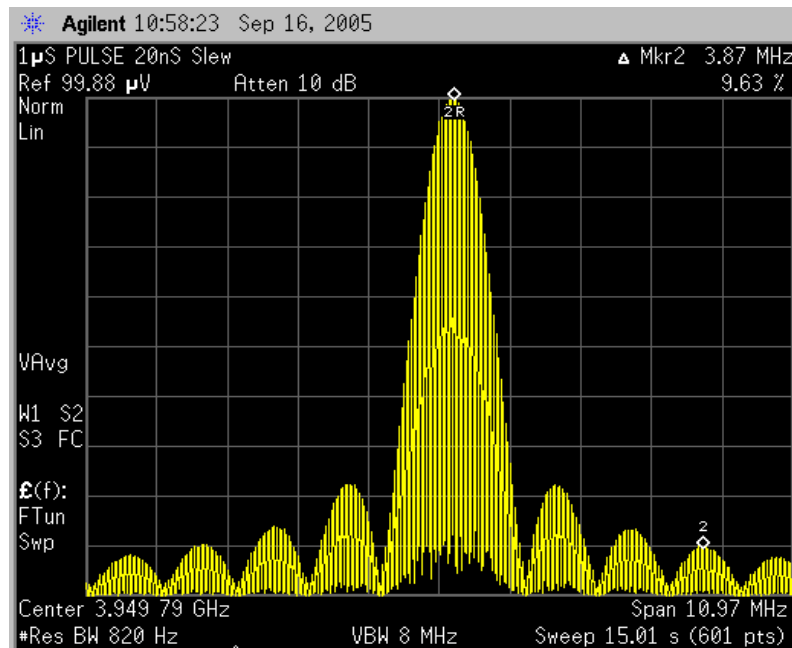


Figure 5 Spectrum of a 1 μs Pulse

Figure 5 shows the spectrum of a 1 μs Pulse with a PRF of 1 kHz. The nulls and peaks in the pulse envelope occur at 1 MHz intervals, which corresponds to 1/Pulsewidth.

The amplitude envelope is in the form of $(\sin(x)/x)$ where the amplitude of one of the discrete frequencies f , is related to its amplitude x , by

$$x = \pi(f - f_0)\tau$$

where: τ = Pulse width (s)

f_0 = Carrier Frequency (Hz)

The theoretical pulse has infinitely fast rise and fall times which spreads the energy over an infinitely wide bandwidth albeit at ever decreasing amplitudes. The practical pulse is rise-time limited by electronic constraints and the need to reduce bandwidth occupancy in order to conserve the usage of the RF and microwave spectrum.

Figure 6 [12] provides a useful representation of the energy spread in terms of two offset frequencies $F1$ and $F2$ either side of the carrier frequency. The limit of $F1$ is where the power can be considered almost constant and depends mainly on the pulse width if this is significantly greater than the rise-time ($\delta1$) and fall-time ($\delta2$). The powers in the sidebands then fall at a rate of 20 dB per decade of frequency until $F2$ away from the carrier frequency. $F2$ is a function of the rise and fall times only and at this point the amplitude falls away at 40 dB per decade.

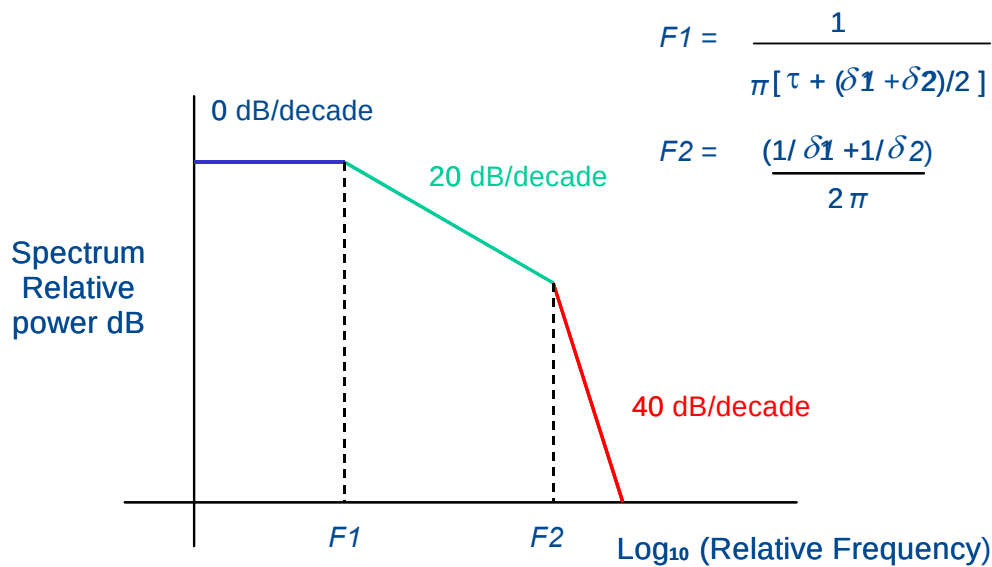


Figure 6

Energy density relative to carrier frequency for a pulse of
 $\delta1$ = Rise time, $\delta2$ = Fall time, τ = Pulse width (at 0.5 peak amplitude).

For the current measurement system the transition times of the pulse are typically no faster than 20 ns and the smallest pulse width is 1 μ S.

This approximates to $F1 = 315$ kHz and $F2 = 16$ MHz. The spectral power will be 20 dB down at $315 + (10 \times 315)$ kHz = 3.465 MHz. The measured spectrum in Figure 5 above shows that at 3.9 MHz from the carrier frequency the sideband voltage has fallen to 9.6% of its maximum, which is 20.4 dB reduction in power. This is in good agreement with the prediction based on the analytical approximation above.

4.5 Power monitoring directional couplers

The power monitoring couplers connect between the amplifiers and the antennas and output a small sample of the transmitted power that is measured by a power sensor. The coupling calibration factor is the ratio of the sampled power to the output power and varies with frequency across its useable bandwidth. The WG10 coupler is a cross-guide coupler for compactness and has the greatest variation with frequency. For a CW signal the worst-case variation in coupling factor is 0.6 dB for a 50 MHz change (2.55 to 2.6 GHz). As predicted above, for a 1 μ s pulse, the relative amplitude of the sidebands would be 20 dB down on the carrier frequency at 3.3 MHz and if the CW coupler calibration values were to be used then the error in the coupling factor for this frequency would be +0.07 dB based on the CW value at the carrier frequency. However, as the error in coupling factor increases with carrier offset, the amplitudes of the sidebands reduce thus minimising the cumulative error in the coupling factor.

If the coupling factors for CW operation were not sufficiently accurate for pulsed operation then it would be necessary to calibrate each coupler for particular pulse widths, which would be extremely undesirable. The effect of the pulse width on the coupling factor was calculated and measured for the worst-case energy spread when a 1 μ s pulse is used. The energy spread depends only on the pulse width and rise time

and therefore changing the pulse repetition frequency will only affect the average power. This relationship was used to optimise the measurement of the coupling factors when pulsed waveforms were used because average power meters would not be sufficiently sensitive to low duty cycle pulses.

Figure 7 shows the calculated cumulative error of a 1 μ s pulse having rise and fall times of 20 nS. The graph was produced using the $(\sin x/x)^2$ function to calculate the power envelope, assuming a perfect rectangular pulse. The error in the coupling factor was applied to the pulse envelope until 16 MHz away from the carrier. Above this frequency, the amplitude was reduced by 40 dB per octave to allow for the finite rise and fall times. The error amplitude was then integrated over the full pulse bandwidth. The error can be positive or negative depending on the gradient of the coupling factor function. Above 16 MHz the amplitude of the sidebands rapidly reduce due to the finite rise and fall times and therefore the total error eventually becomes constant.

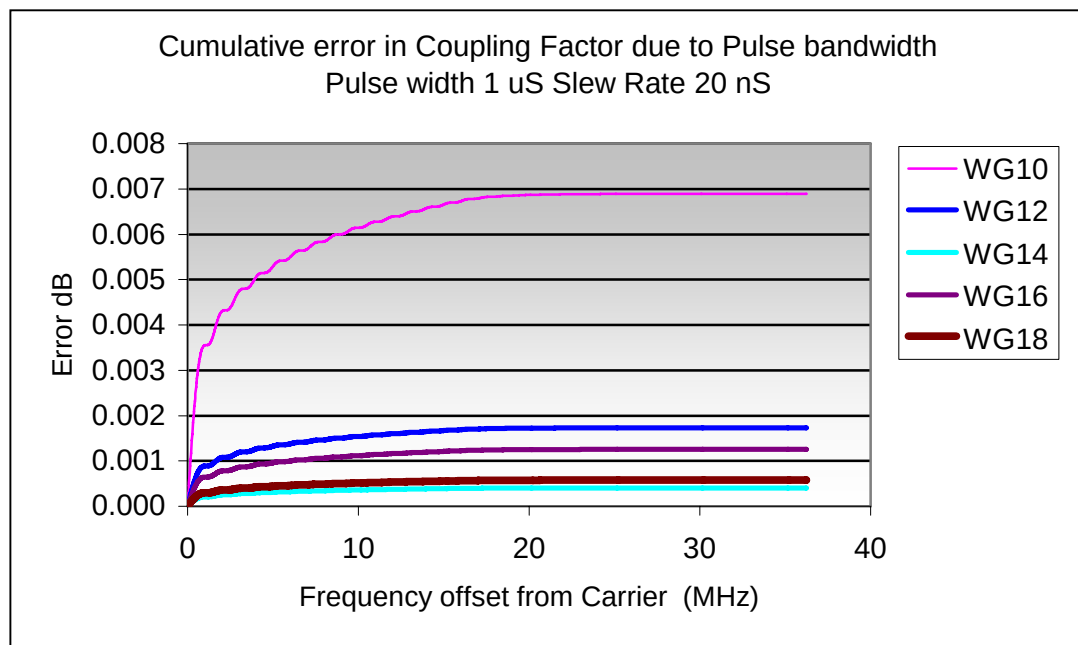


Figure 7

Cumulative Error in Coupling Factor due to Pulse Bandwidth

Similar calculations were done for the antenna gains. The gain variations with frequency are less than for the couplers. Again the WG10 antenna had the greatest variation with frequency at 0.2 dB for a 50 MHz frequency change. For a 1 μ s pulse, this is similar to the error produced by the WG12 coupler, which is ± 0.002 dB.

The reference power sensor, used for calibrating the couplers and sidearm power sensor, has a small frequency dependency with a worst case of 0.013 dB for a 50 MHz change. This is an order less than the antenna gain and can be neglected.

4.6 Measuring the Coupling Factors

In order to confirm the theoretical analysis, the couplers were calibrated against a thermal power sensor using pulsed and continuous waveform. These measurements were also required to calibrate the couplers using the Pulse Power Sensor. Note that the coupling factors relate the power indicated by the pulse power sensor to the average power indicated by a calibrated power sensor on the main output arm of the coupler. The antenna is removed prior to the calibration of the coupler.

The calibration set up is shown in Figure 8 . The Agilent E9327 diode power sensor was connected to the attenuated side arm of a waveguide coupler and a calibrated waveguide power sensor, consisting of an Agilent 8481H average power sensor and WG to coaxial transition, attached to the main arm output. Both of these sensors were connected to an Agilent E4417A power meter, which was set to read in ratio mode. Five such waveguide assemblies, covering the frequency band of 2.45 GHz to 18 GHz were measured.

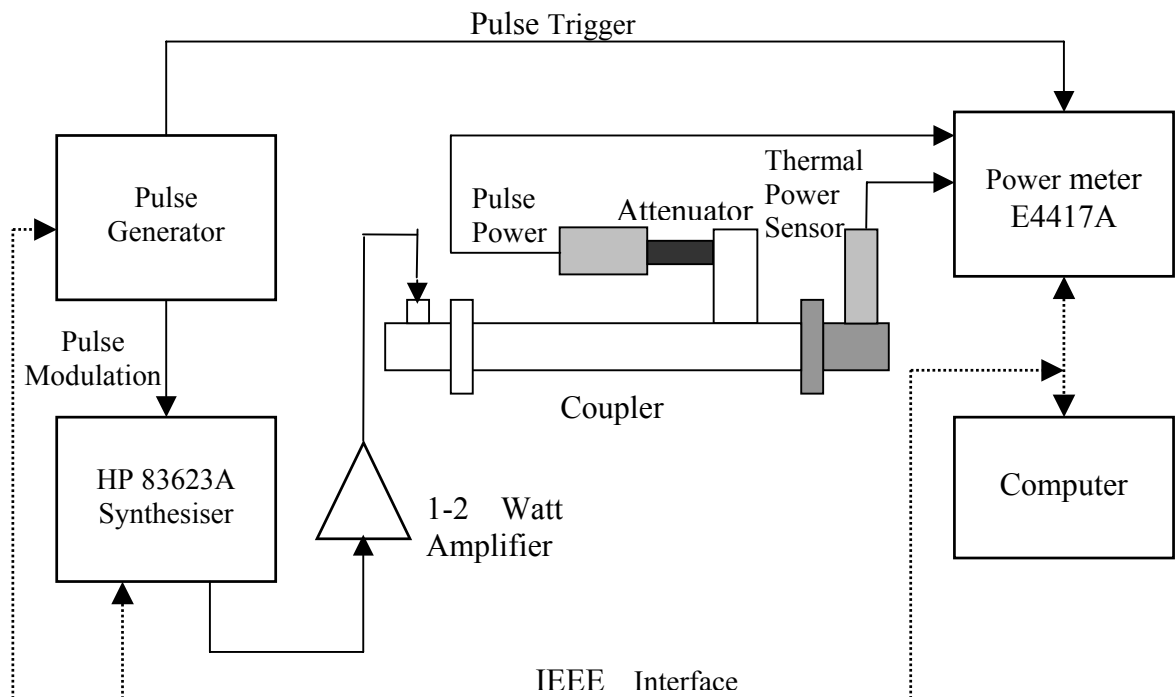


Figure 8. Measurement of system coupling factors

The thermal power sensor used as the reference has a maximum rating of 3 watts and had been calibrated at 10 mW and 1 Watt. The differences in calibration factors between these power levels are within 0.2% so the actual power level used is not critical. A 20 dB coupler and CW at 1W was first used to obtain a narrow band reference. Then various pulse waveforms of widths 1us and 10uS, with varying PRFs were used to assess the effects on the measured coupling factor. Figure 9 shows the coupling factors for pulsed waveform relative to that obtained using CW. A constant peak power of 1 W was used; therefore, the operating level for the CW reference sensor varied from 1 W down to 1 mW for the smallest duty cycle.

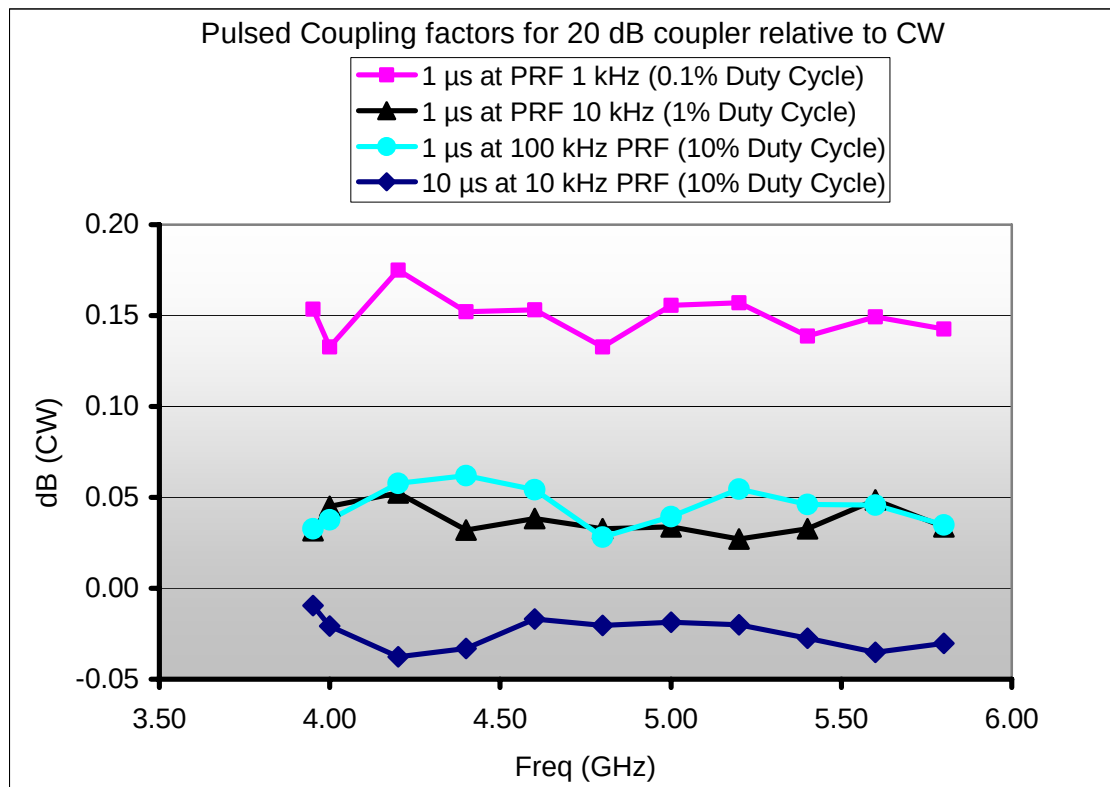


Figure 9 Coupling Factors for pulsed operation relative to CW

The data shows that for duty cycles of 1% and 10% using pulse widths of 1 and 10 μ s, the coupling factors were within ± 0.05 dB of CW values, but for a 0.1% duty cycle and 1 μ s pulse width, the difference is 0.15 dB. It would seem therefore, that low average power levels (proportional to the duty cycle) increase the pulsed coupling factors but the pulse width has little or no influence. Further work is required to fully investigate the effects that low duty cycles are having on the thermal sensor.

The couplers needed for the calibration system require coupling factors of between 43 and 46 dB to allow power monitoring up to 2.5 kW with a 100 mW power sensor. Using 1 Watt of peak power with a 1 μ s pulse width and a PRF of 100 kHz results in an average power at the reference sensor of 100 mW and a peak level at the pulse sensor of between 2.5 and 6 μ W.

Figure 10 shows the typical measured coupling factors for WG16 (8.2 – 12.4 GHz). The average difference between CW and pulsed waveform is 0.04 dB or 0.09% and the difference increases successively for WG sizes 12, 14, 16 and 18. WG18 has the largest difference of 0.08 dB (1.8%) and all the factors obtained using pulsed waveform were consistently less than that for CW. One possible explanation is that the integration of the noise by the pulsed power meter resulted in a slightly higher average power indication because the power was only 30 dB above the noise floor for this sensor. Using a duty cycle of 10% means that the signal is only present for 10% of the pulse period whereas the noise floor is integrated over the whole period. Note that the widest bandwidth (WG18) resulted in the largest differences between CW and pulse measurements.

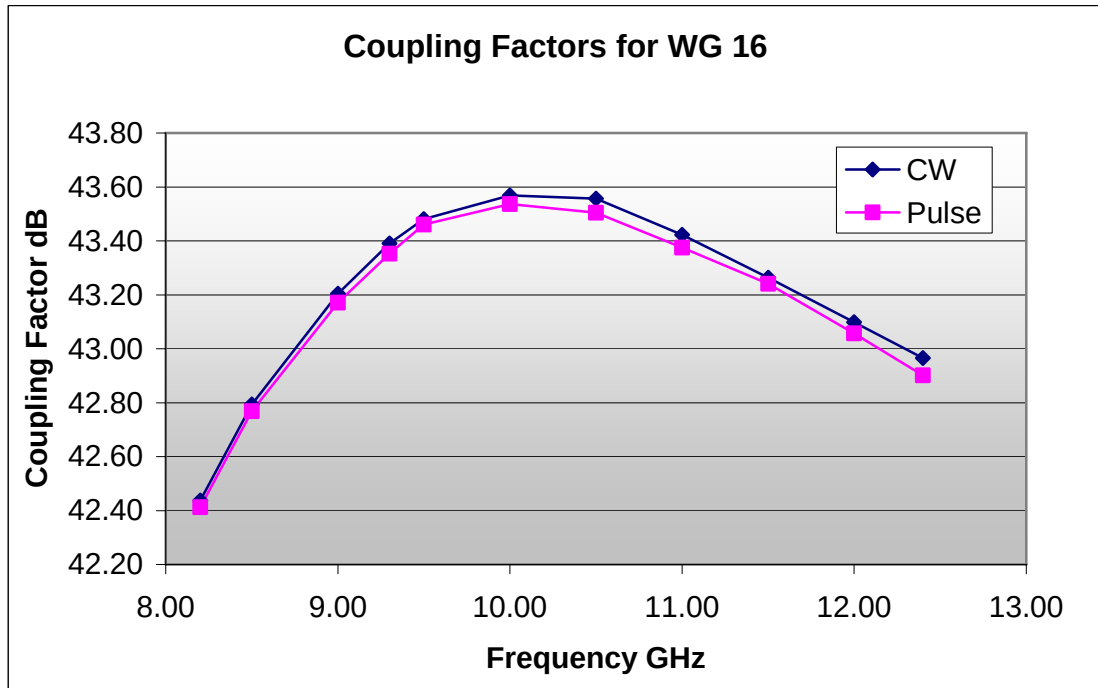


Figure 10 Coupling Factors for Waveguide 16

The uncertainty in the measurement of the coupling factors for CW is ± 0.2 dB and the typical repeatability is ± 0.05 dB. The levels of agreement between the CW and pulsed coupling factors and the improved signal to noise ratio for the CW measurements justify the use of the CW factors for the final calibration.

5 Field Probes and Sensors

The primary purpose of the calibration is to obtain correction factors for a range of frequencies and field levels. Note that the correction factor is the inverse of sensitivity.

Two correction factors are useful for pulsed calibrations:

$$CF(Ave) = \frac{\text{Average applied power flux density}}{\text{probe indication}} \quad \text{Equation 3}$$

$$CF(Peak) = \frac{\text{Peak applied power flux density}}{\text{probe indication}} \quad \text{Equation 4}$$

The sensitivity variation due to the field amplitude, known as the linearity, is not significantly affected by the frequency of the EM radiation for CW operation. Therefore, it is standard practice to calibrate over a wide range of frequencies at one or two field levels and at one frequency to measure over a wide range of field levels.

The sensors in RF probes are designed to respond to either the electric or magnetic field component. Below 300 MHz it is the magnetic field that is often measured because electric field sensors need to be large or feed very high impedances to maintain their sensitivity. At higher frequencies, the impedances of loop sensors become very high and they need to be very small to avoid cross coupling with the E-field and for these reasons, small dipoles become the preferred option.

A probe requires some form of antenna to couple into the desired field component. A single sensor responds only to one field polarisation and therefore must be orientated correctly to read the maximum value. Similarly, multiple dipole or 'crossed dipole' sensors arranged in a single plane will respond correctly when the direction of propagation is normal to the plane of the sensor. Three sensors organised in an orthogonal arrangement can be used to sum all field components, independent of polarisation and angle/direction of incidence; these are known as isotropic probes.

A probe also requires a detector to convert the high frequency energy into a measurable DC voltage.

5.1 Thermocouple Detection

Thermocouple based instruments generally use thin-film thermocouple pairs adhered to a plastic substrate as the detection element. They measure the minute temperature rise due to the currents flowing from the antenna into a load. This method of detection has the potential to provide a DC voltage output that is truly proportional to the average PFD with extremely wide bandwidths typically greater than 50 GHz. The disadvantages of thermocouple detection are their relatively long time constants (typically >1 sec) and lower burn out threshold due to non-uniform heating within the loads; especially when exposed to pulsed signals. They are also significantly more expensive than probes employing diodes for detection.

5.2 Diode Detection

Diode detectors are often used to terminate small loop antennas (H-field) or electrically short dipoles (E-field). Schottky diodes operate as efficient high frequency half wave rectifiers to convert the high frequency alternating voltage to a steady DC component..

The basic elements of a probe using diode detection are shown in Figure 11. The antenna is much smaller than a wavelength and can be represented by a resistor and capacitor R_A and C_A . A Schottky diode is represented by a small junction capacitance that is in parallel with C_P , dependent on the impressed voltage. The diode also has a non-linear resistance, which is dependant on the current flowing through the diode. The diode can be connected in shunt or series with the antenna but it is not the intention of this report to describe all the different configurations in detail. R_T and C_T form an averaging circuit that may be fixed or adjustable, sometimes referred to as the meter response time and R_L is the load resistance presented by the meter. The meter may be a voltmeter or a digital processor capable of carrying out conversions between units, performing data logging or averaging over long periods of time, for example the 6 minute averaging required by ICNIRP.

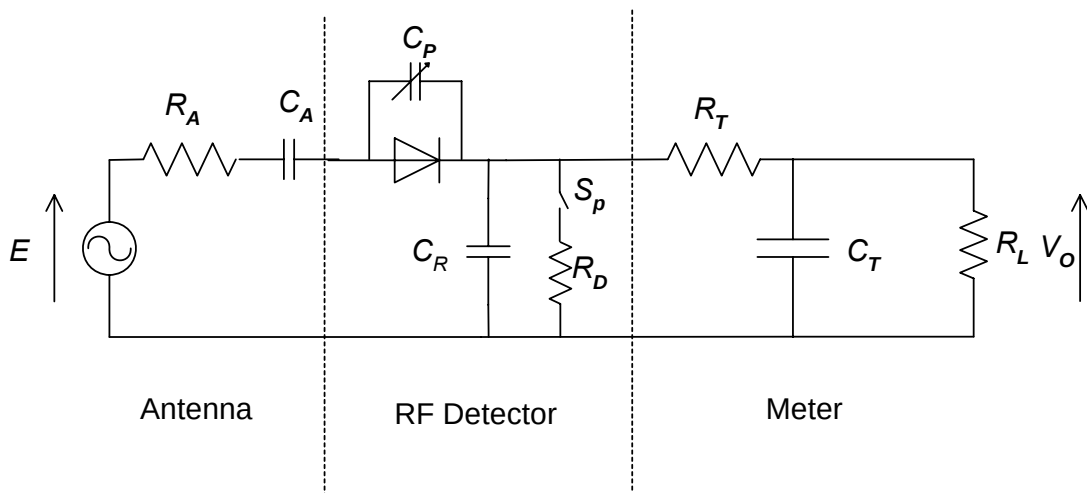


Figure 11 Elements of a typical diode detector probe

The antenna operates in a similar way for both CW and pulsed fields. A CW sinusoidal field causes current to flow through the antenna impedance formed by R_A and C_A which charges up C_P on the positive half cycles. The diode also has a resistance, which is dependent on the voltage across it, and over several cycles, there will be sufficient charge stored to produce a steady state voltage across C_R . Components R_T and C_T average the voltage over a period defined by the meter response time and a steady output, proportional to the peak amplitude of the field, will be achieved across R_L . S_p is a peak hold switch preventing C_R from discharging when the field strength is removed or reduced. When S_p is closed the detector output follows the field amplitude. For a CW signal, the meter sensitivity is designed to ensure that it reaches full scale indication within the square-law range of the diode or, for greater dynamic range, the meter circuitry can attempt to correct for the non square-law errors based on the average detected voltage. It is important to realise that in order for the probe to correctly interpret non-sinusoidal signals (signals containing harmonics or multiple frequencies) and modulated sinusoidal signals, the detector output envelope must be proportional to the square of the E-field and therefore all diode field probes are inherently power flux density meters. Conversion to E-field units is done by a square rooting process in the meter.

When the circuit is used to measure pulsed fields, the pulse width and PRF affect the various components to influence the way that the detection circuit and averaging circuit relate the field strength E to the voltage V_o , output to the meter.

5.2.1. Diode Detection Errors

Reference [11] describes the nonlinear behavior of diode detectors and shows that the smooth transition from square law to linear law is realistic only under restricted conditions for the load resistance, temperature and bias current. For other conditions the square law can progress to higher powers during the transition region. When a probe is used to measure pulsed fields the peak level often exceeds the square law region, resulting in higher or lower sensitivity depending on the law of the transition region. Furthermore as the transition region extends over a 20 dB dynamic range it is likely that the true linear region may never be reached. For example a probe having a maximum square law range of 800 V/m would operate in the transition region up to

8 kV/m. Over this range the sensitivity could decrease by up to 7 dB or increase by 19 dB depending on the detector load resistance [11].

5.2.2. Detector Response

The detector output has a rise time defined by $R_A C_R$, which may be many cycles or even many pulses of the carrier frequency. The discharge constant is $R_D C_R$ and must be large enough to remove the carrier frequency, so for a probe operating down to 3 MHz, the discharge time of the detector would need to be greater than 1.7 μ s. Also, in order to capture the true peak levels of discontinuous pulses (burst pulses), the rise time must be less than the pulse width, which can be 0.5 μ s, or less. However if the pulse shape is to be preserved then both rise and fall time constants must be similar to the rise and fall time of the pulse, which is typically 0.02 μ s. In practice the overriding aspect of performance is the ability to operate over a broad band and consequently some performance may be lost due to distortion of the detected pulse envelope.

5.2.3. Meter Response

In order to produce a steady state output proportional to the heating effect of the power flux density or field strength it is necessary to average the rapidly changing envelope voltage from the pulse modulation. For the circuit shown the averaging period is determined by the filter time constant RC which should be at least 5 times longer than the lowest pulse repetition interval (PRI) otherwise the output will be unstable. Most radars have PRIs less than 5 ms, so provided the averaging time is more than 30 ms, a stable reading will be achieved. However, the radar beam must illuminate the probe for at least the response time of the probe otherwise desensitisation occurs [13]. For example if the probe is used to measure rotating radar then there will be a reduction in the indicated field level approximating to:

$$A(dB) = 5 \log_{10} \left(1 + \left(\frac{t_{probe}}{t_{illum}} \right)^2 \right) \quad \text{Equation 5}$$

5.3 Frequency Selective Probes

Several frequency selective field probes are now commercially available. These probe systems retain the essential feature of being minimally perturbing, by employing electrically small antennas, but use a spectrum analysis approach to detect and display the measurements. The obvious advantage is that these probes can determine the frequency as well as the amplitude in a multi-signal environment; but another important feature is that the diode detector is always presented with sinusoidal waveform of the same frequency making detection more accurate. When measuring broadband signals such as pulsed or modulated carriers, care must be taken to ensure that the results are correctly interpreted for the operating mode. In Time Analysis mode the receiver is set to the carrier frequency and the amplitude displayed versus time.

Provided that $(0.2/Pulse\ Width) > Bandwidth > PRF$ then:

$$Peak\ Power = Measured\ Power \times (Pulse\ Width \times Impulse\ Bandwidth)^2$$

The Pulse Width and PRF can be measured in the time domain. The power averaged over one pulse repetition interval is given by:

$$\text{Average Power} = \text{Peak Power} \times \text{Pulse Width} \times \text{PRF}$$

The time domain mode may also have a time average parameter to obtain the 6-minute period required by ICNIRP.

The spectrum mode displays the amplitude of the individual spectral lines against frequency. The bandwidth must be adjusted to be less than the PRF to ensure only one spectral line falls in the pass band of the receiver. The amplitude of the central spectral line f_0 is measured and displayed by the SRM but this must be corrected for pulse desensitisation such that:

$$\text{Peak Power} = \text{Measured Power} \times \text{Desensitisation Factor}$$

where
$$\text{Desensitisation Factor} = \frac{1}{(\text{Pulse Width} \times \text{PRF})^2}$$

For example, a 2 μs pulse with a PRF of 50 kHz would have a desensitisation factor of 100, or 20 dB.

6 Probe Measurements

In order to investigate the magnitude of the errors described in paragraph 5, five commercially available field probes were tested using a range of pulse widths and PRFs. High peak levels were generated using pulse TWT amplifiers where low duty cycles were required and CW amplifiers were pulse modulated for duty cycles up to 100%. Broadband probes with diode detection, thermocouple detection, and frequency selective probes with diode detection were measured and the relative merits analysed.

6.1 Narda Type 33 (Thermocouple)

This probe is designed to operate from 300 MHz to 50 GHz and has a dynamic range of 0.17 to 1000 W/m^2 (8 to 614 V/m) for CW. The peak overload level is specified as 2000 kW/m^2 (27 kV/m) for a 1 μs pulse with duty cycle of 0.1%. The probe has three orthogonal sensors that are combined within the probe as root sum of squares (rss), to provide the resultant field independent of field polarisation and direction of propagation. The probe connected directly to an indicating meter, which was optically coupled to a computer for control and readout. The probe correction factors were measured using the CW calibration system and then with the pulse modulated system. The results are shown in Figure 12.

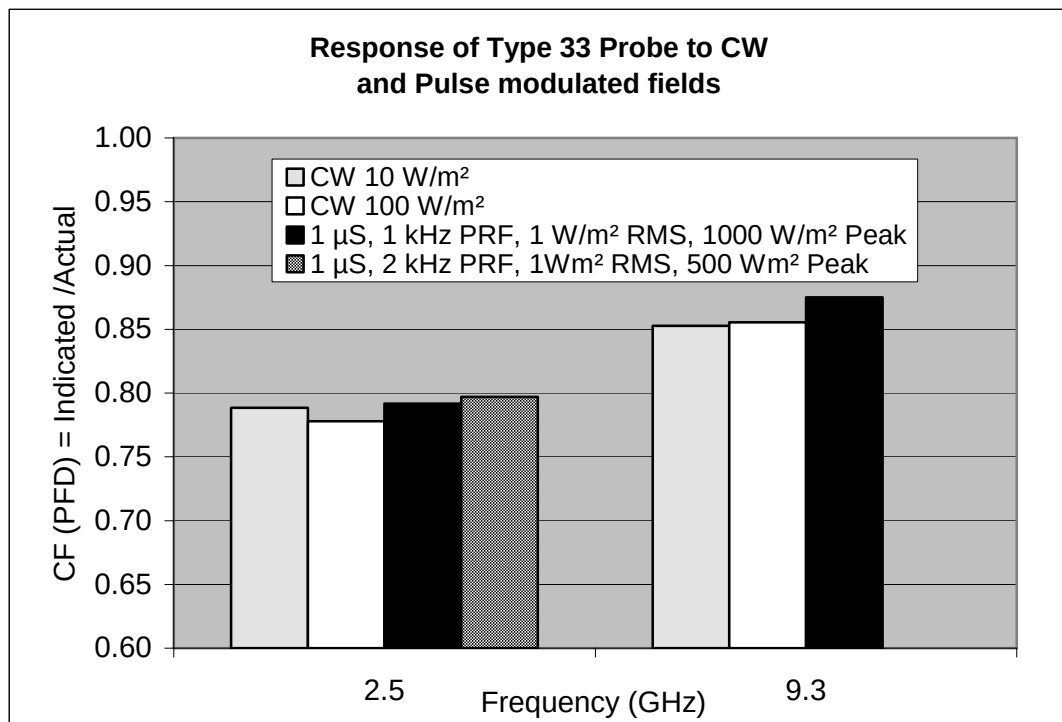


Figure 12

Response of Type 33 Thermocouple Probe to CW and pulse modulated fields

The probe correction factors agreed within 1% for duty cycles from 0.1 to 0.5 % for the pulse-modulated system. The measured difference between the pulsed and CW measurements is also very small, typically 1 – 3% which is comparable with the linearity deviation for CW and less than the uncorrelated uncertainties between the pulse and CW measurement methods. Some of the data spread for pulsed measurements can be attributed to the relatively low level.

The probe is only suitable for measuring stationary radars because the thermocouple detectors have a response time specified as 100 ms, which would result in desensitisation factors of greater than 25 dB.

The measurements have demonstrated that the within the uncertainty limits imposed by the measurement system, the Type 33 probe responded to the true field strength for both CW and various pulse-modulated waveform.

6.2 Amplifier Research Type FP7050 (Thermocouple)

This probe has an identical specification to the Type 33 probe. The probe connects to an optical sender unit, which is then optically coupled to the computer for readout and control.

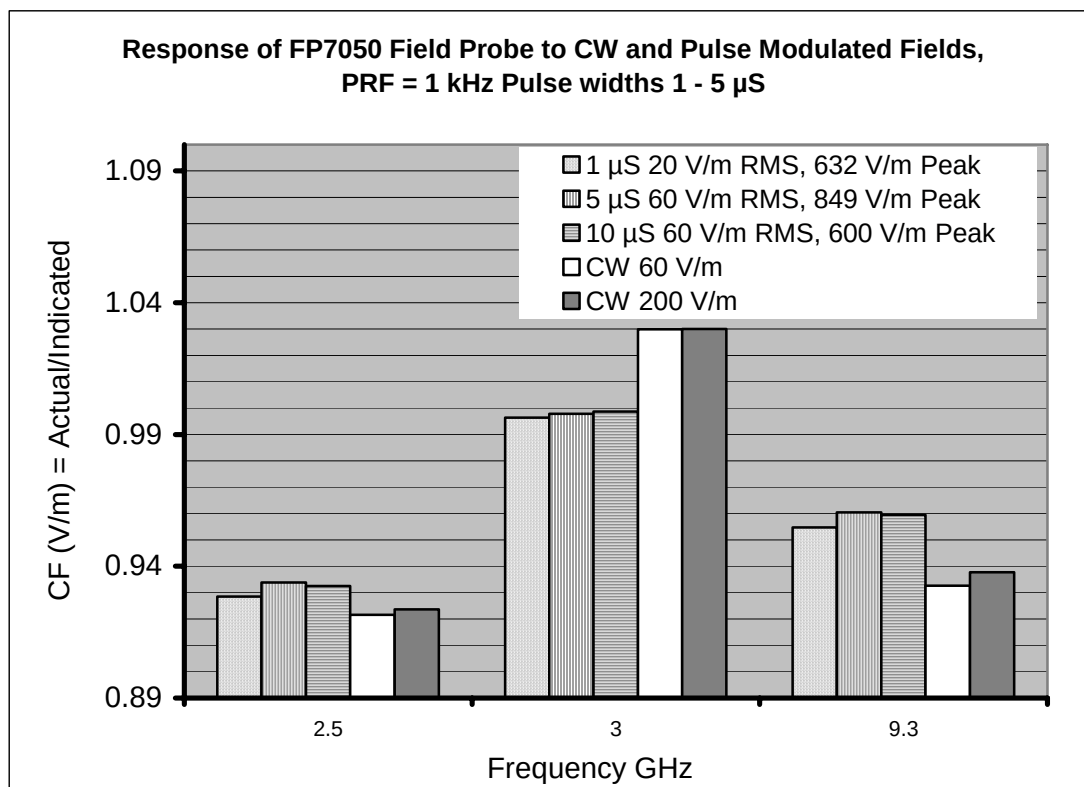


Figure 13

Response of FP7050 Thermocouple Probe to CW and pulse modulated fields

The probe correction factors agreed within 1% for duty cycles from 0.1 to 1 % for the pulse-modulated system. The measured difference between the pulsed and CW measurements is typically 1 – 3%, which is less than the uncorrelated uncertainties between the pulse and CW measurement methods. Some of the data spread for pulsed measurements can be attributed to the relatively low level.

The probe is only suitable for measuring stationary radars because the thermocouple detectors have a response time specified as 100 ms, which would result in desensitisation factors of greater than 25 dB.

The measurements have demonstrated that within the uncertainty limits imposed by the measurement system, the FP7050 responded to the true field strength for both CW and various pulse-modulated waveform.

6.3 Narda Type 9.2 (Diode)

This probe is designed to operate from 3 MHz to 18 GHz and has a dynamic range of 0.0038 to 2653 W/m² (8 to 1000 V/m) for CW operation. The peak overload level is specified as 700 kW/m² (16 kV/m) for a 1 µs pulse and duty cycle of 0.1%. The probe has three orthogonal sensors that are individually processed in the meter to provide the outputs from each axis and the resultant field independent of field polarisation and direction of propagation. The probe connects directly to an indicating meter, which was optically coupled to a computer for control and readout.

Initial measurements using the pulsed system showed large variations in the probe sensitivity for changing pulse parameters. For low duty cycles from 0.05 to 2 % the probe correction factor varied from 2 down to 0.1 compared with a CW value of 1.07 at 2.5 GHz.

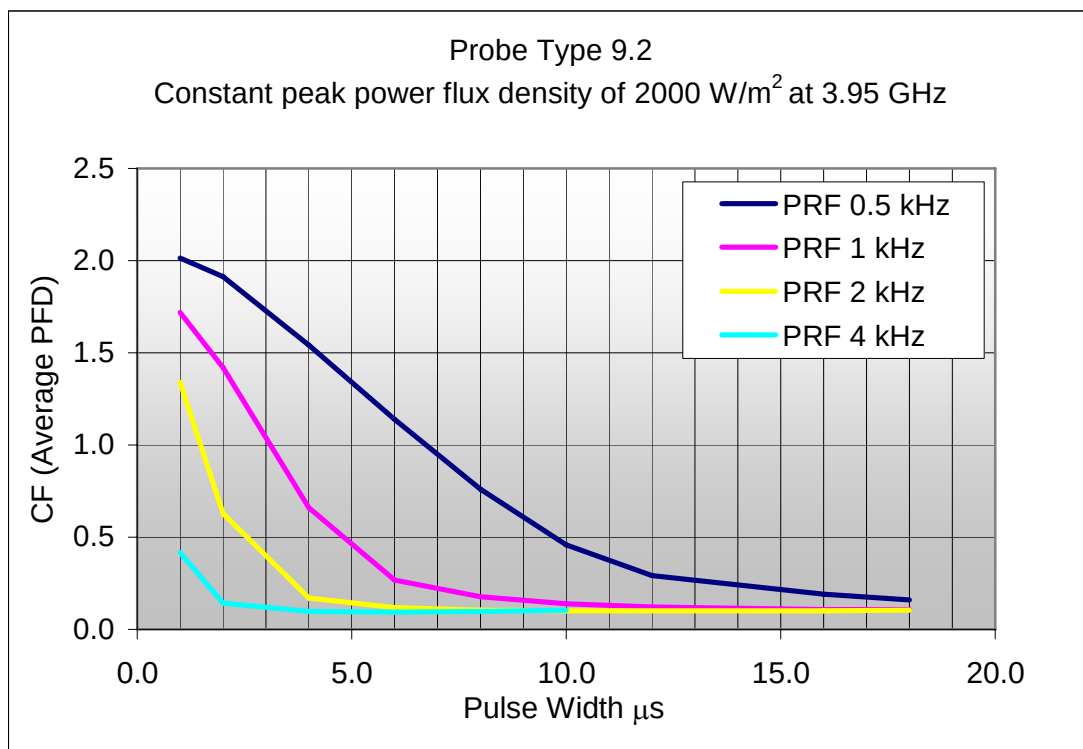


Figure 14

Average Correction Factors for Diode Probe Type 9.2 as a function of PRF for a constant peak Power Flux Density.

Figure 14 shows how the probe correction factor varies with increasing pulse width for a range of PRFs. The correction factor uses the power flux density averaged over the pulse period, which is equivalent to the average, or heating effect of the field. The peak field level was kept constant to remove any variations due to non-square law detection. The pulse train is continuous and the averaging period or response time is specified as 1 second, which is much greater than the pulse period so the averaging circuit will not influence the results. The variations in Correction Factor are therefore solely due to the pulse response of the detector. For small pulse widths and low PRFs, the probe is desensitised by the detector time constant. The rise time of the detector is such that the detected output never reaches the peak of the pulse. As the pulse width

increases, the detector output reaches the peak amplitude of the pulse but the detector discharge time is too long to correctly average the pulse envelope. This causes a higher than reading and the Correction Factor reduces.

Figure 15 shows the same data plotted relative to the Duty Cycle. The close overlay of the three PRF data sets shows the pulse response to be closely linked to the Duty Cycle, which is effectively the normalized average level. The data diverges for Duty Cycles below 1 % where the pulse widths are less than the detector rise time.

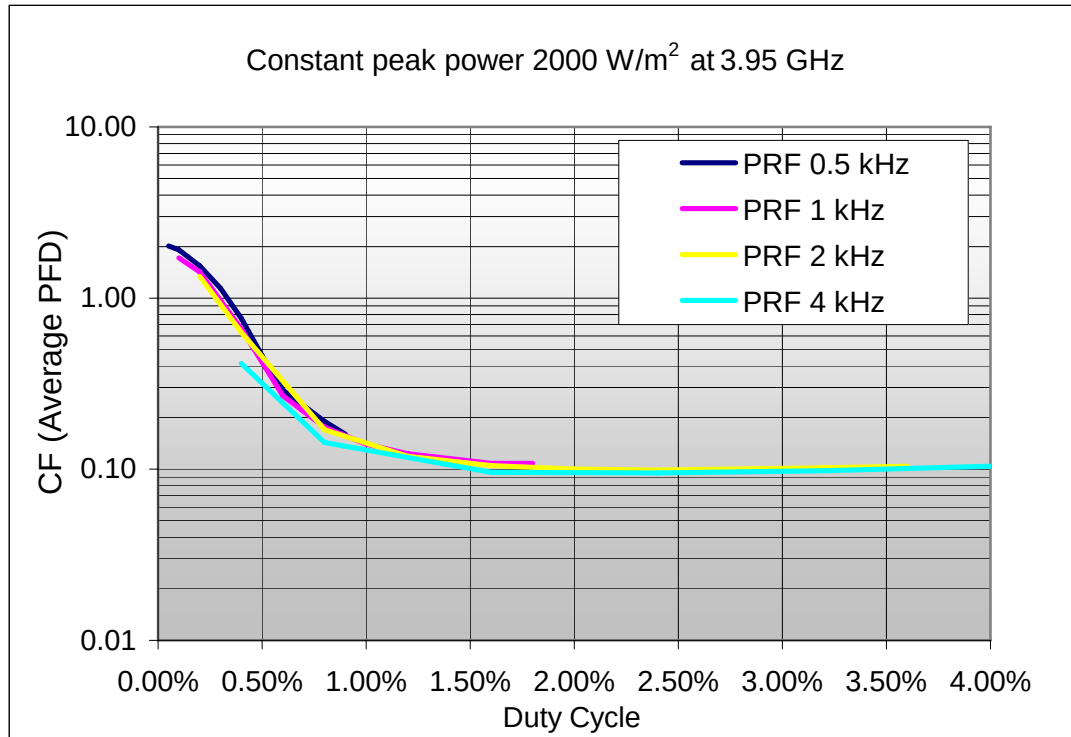


Figure 15

Average Correction Factors for Diode Probe Type 9.2 as a function of Duty Cycle for a constant peak Power Flux Density.

For Duty Cycles greater than 2% the probe sensitivity remains constant but reading 10 times higher than the true level. Eventually, as the duty cycle approaches 100% the probe correction factor must reach the CW value. To investigate this a CW amplifier was used to generate pulsed modulation with duty cycles up to 100% for a constant peak amplitude and PRF. The results are shown in Figure 16 which shows the probe correction factor relative to the average or power flux density. The RF frequency was chosen to suit the CW amplifier and is not the same as Figure 15. However, the graph shows a sharp reduction in probe sensitivity below 4% Duty Cycle and a gradual linear decrease in sensitivity from 4% to 100% Duty Cycle. To understand the latter decrease in sensitivity it must be noted that the sensitivity is relative to the average power flux density.

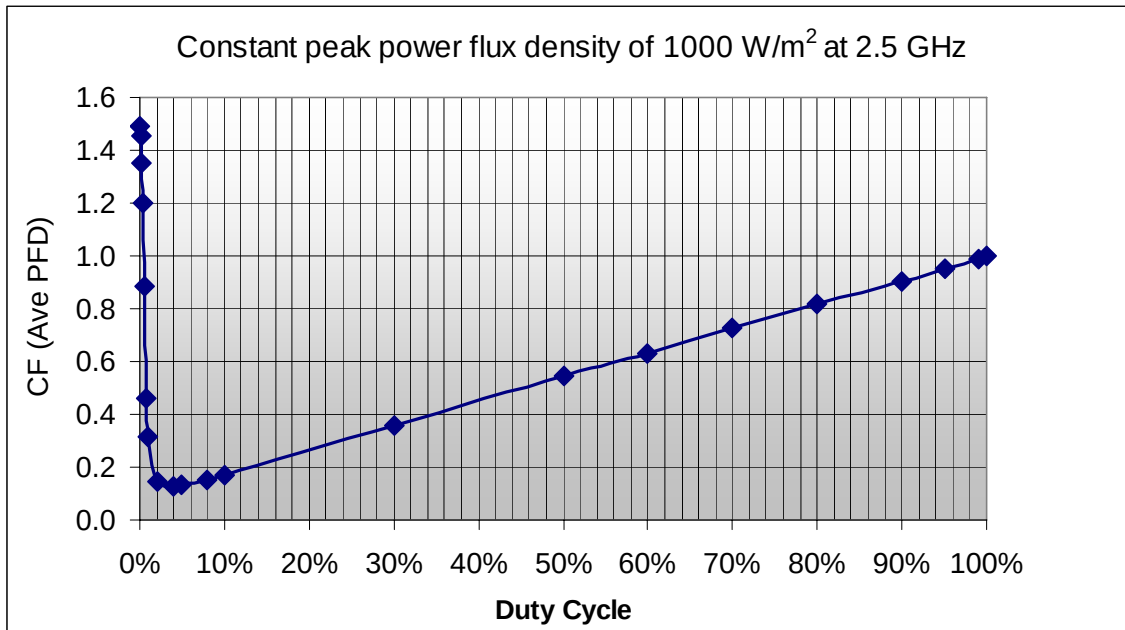


Figure 16

Average Correction Factors for Diode Probe Type 9.2 as a function of Duty Cycle up to 100% for a constant peak Power Flux Density.

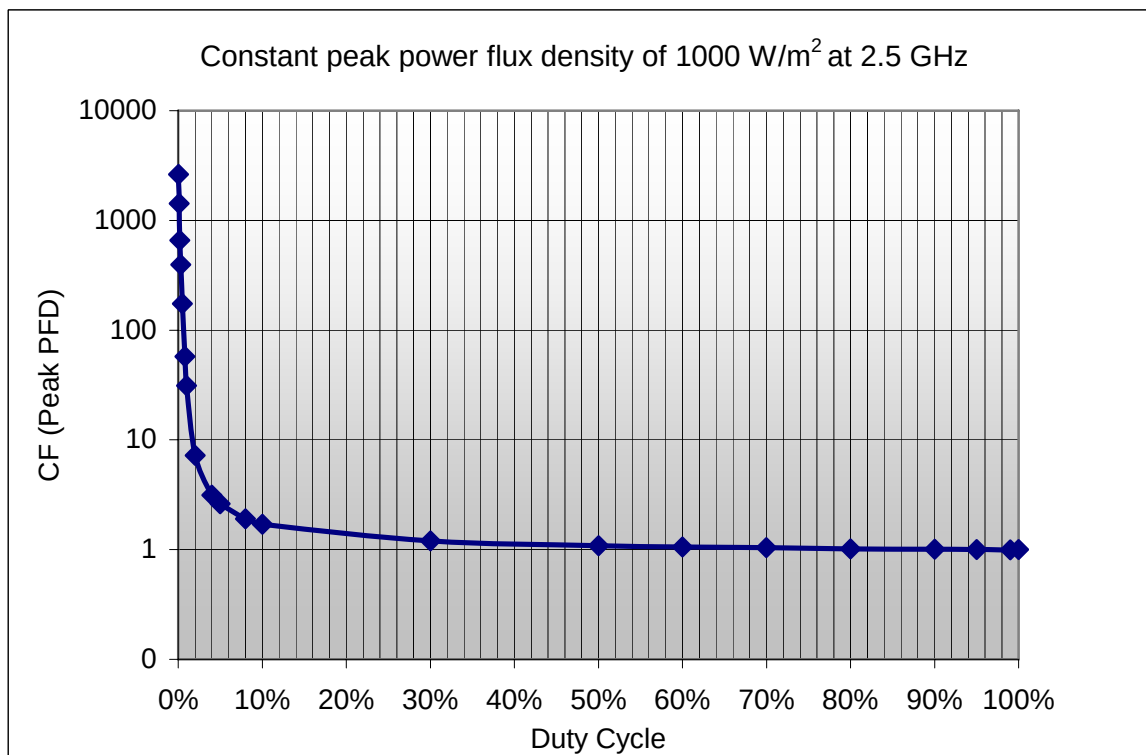


Figure 17

Peak Correction Factors for Diode Probe Type 9.2 as a function of Duty Cycle up to 100% for a constant peak Power Flux Density.

One of the reasons to measure field strength from pulsed radiators is to determine the field level at the peak of the pulse. It is useful therefore to use the same measured data from Figure 16 and plot the correction factor relative to the Peak field.

Figure 17 shows the constant peak power divided by the probe indication, plotted against the duty cycle, which is proportional to the power flux density averaged over the pulse period. The probe indicated field level increases rapidly up to a duty cycle of 4% and then more slowly as the average power increases from 4 to 30%. It then remains almost constant up to 100%. The pulse width for a duty cycle of 30% is 300 μs , which means that for a PRF of 1 kHz, the carrier is off for 700 μs . The detector is now operating as a peak detector with minimal droop over the 700 μs off period. Increasing the pulse width further produced very little extra output voltage from the detector. The apparent decrease in sensitivity from 30 to 100% Figure 16 is almost entirely due to the increase in the applied average power flux density as the pulse width increases.

In previous figures the probe response has been measured at constant applied peak levels. The effects of varying the applied peak level are shown in Figure 18. The CW values are included for comparison. The graphs show that the indicated value is linearly proportional to the applied peak level but the constant of proportionality, which is the probe correction factor, increases with the PRF until it reaches the CW value of 1. It is interesting to note that for low field levels the linearity is not as good as the higher levels and there is no evidence of diode non-linearity at the higher levels.

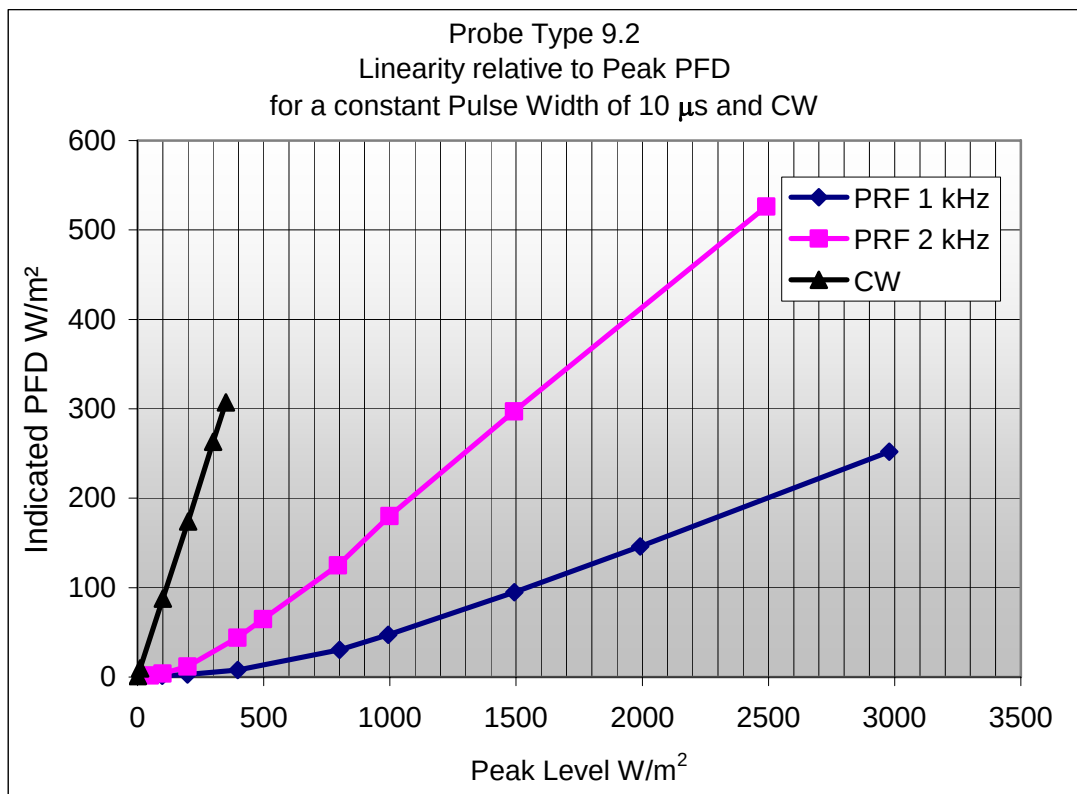


Figure 18
Linearity relative to the Peak PFD of Diode Probe Type 9.2
for varying Duty Cycles up to 100%

The linear response with the peak field results in non-linear behaviour when translated to the average field quantities. Figure 19 shows the how the probe correction factor for the average power flux density varies with the applied average field level. The graphs show that it is the Duty Cycle that determines the linearity response. Note that a probe with perfect linearity would have a constant correction factor.

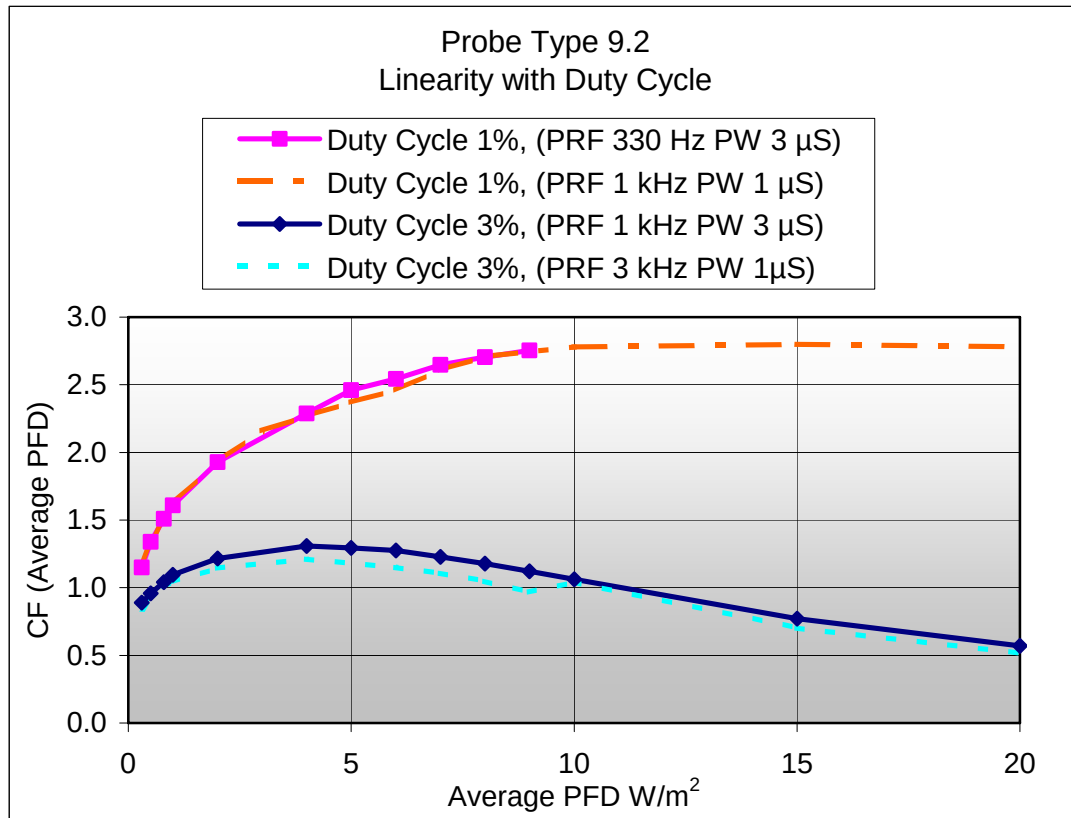


Figure 19

**Linearity relative to the Average PFD of Diode Probe Type 9.2
showing dependency on Duty Cycle**

The Type 9.2 Probe has a peak hold capability, which can capture a short burst of pulse modulation equivalent to the measurement of rotating radar. Knowing the speed of rotation and the 3 dB beam width, the number of pulses within this time window can be calculated for a particular radar and used in the calibration system. The response time of the detector will reduce the indicated field level according to Equation 5.

Figure 20 shows the probe correction factor for discontinuous pulses relative to that for continuous pulses, which is equivalent to the desensitisation factor. The Burst Period is the time interval between the bursts of pulses simulating the scanning period of a radar transmitter. A meter response time of 300 μ s has been used to calculate the desensitisation factor for t_{probe} in Equation 5. This was determined by experiment to obtain the closest fit to the measured data. The agreement between measured and calculated was between 0.04 and 1.75 dB.

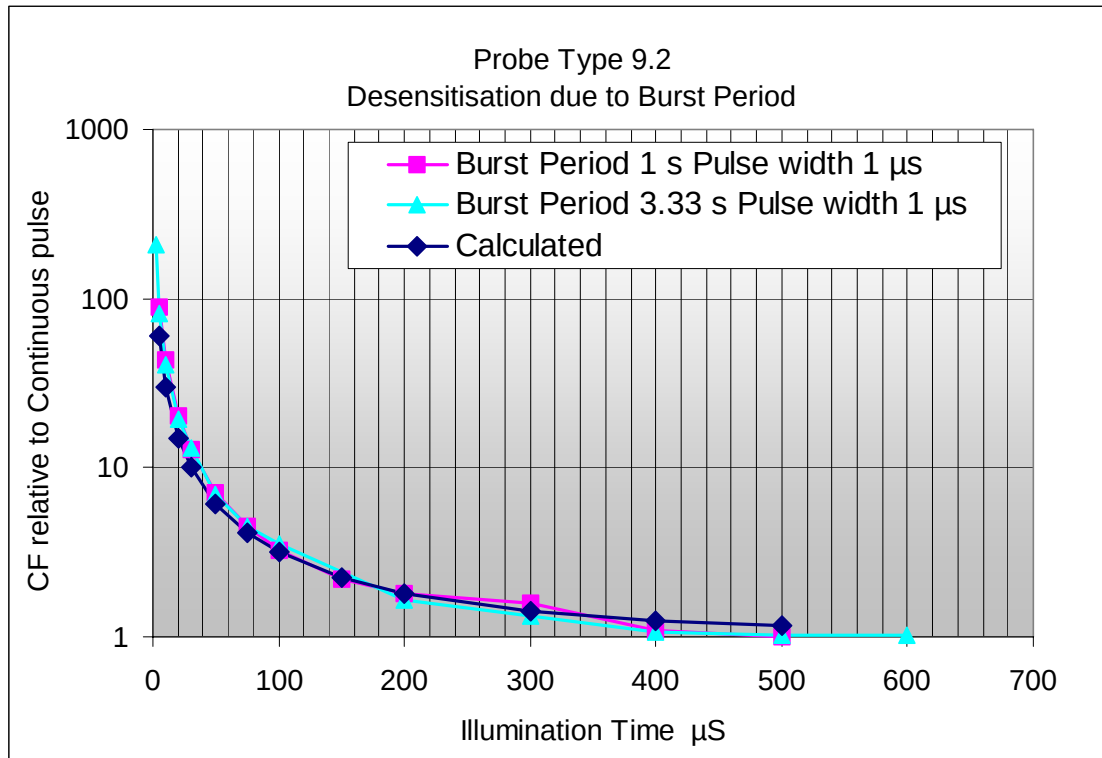
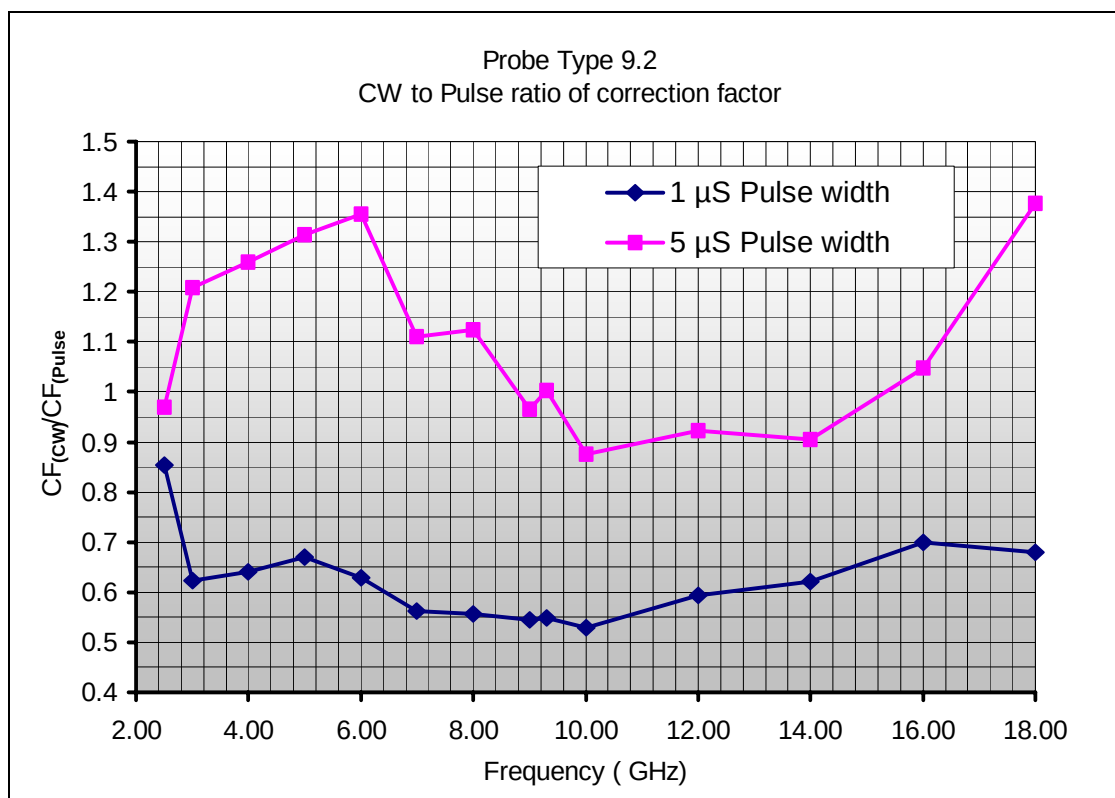


Figure 20

Measured and calculated Desensitisation Factors of Diode Probe Type 9.2 as a function of illumination time.

The effects of varying the carrier frequency have been investigated in Figure 21. The probe was calibrated at regular frequency intervals from 2.5 to 18 GHz for CW operation and for a 1 kHz PRF with pulse widths of 1 μs and 5 μs . The ratio of the CW and Pulse correction factors has been plotted against frequency. The results show that there is significant variation with frequency and therefore pulse characterisation must be done for each frequency of operation.

**Figure 21**

**Correction Factors of Diode Probe Type 9.2 for pulsed waveforms
relative to CW values as a function of Carrier Frequency.**

6.4 Narda SRM 3000 (Frequency Selective)

The probe has a triaxial antenna connected to a receiver display unit. The probe was measured for the X-axis only. A Narda representative operated the probe and a comparison was made between the NPL generating system and the Probe readings. The correct setting and choice of measurement modes has been described in the manufacturers application note 14

Due to time constraints only a limited number of measurements were made. These are presented in Table 3

Table 3

Results of measurements using Narda SRM 3000 (Frequency Selective)

Pulse Width μS	PRF kHz	NPL System		Probe Type SRM 3000		
				Time Analysis Mode		Spectrum Mode
		Peak V/m	Average V/m	Peak V/m	Average V/m	Peak V/m
10	1	100	10	106.6	9.6	107.8*
1	1	181	6	196	5.65	
CW	CW	50	50			51.3

* after correction using manufacturer's Desensitisation Factor.

The results show agreement to better than 0.7 dB for pulsed signals and 0.2 dB for CW. The SRM was also able to correctly measure the carrier frequency, the PRF and pulse width.

7 Conclusions

The NPL field strength and power flux density calibration system has been extended to allow the calibration of field probes using pulse modulated microwave frequencies from 2.5 to 18 GHz. The additional uncertainties due to the measurement of the peak and average power have been analysed and have increased the measurement uncertainty typically from 0.4 dB to 0.5 dB. The principles of operation of commercially available field probes have been described. No attempt has been made to model or predict their performance under pulsed conditions as has been previously been described by other workers in this field [15]. This is primarily because a precise knowledge of the components of each probe would be required and the safety aspects of probe calibration will always require a traceable calibration to be performed to verify the model and its input data.

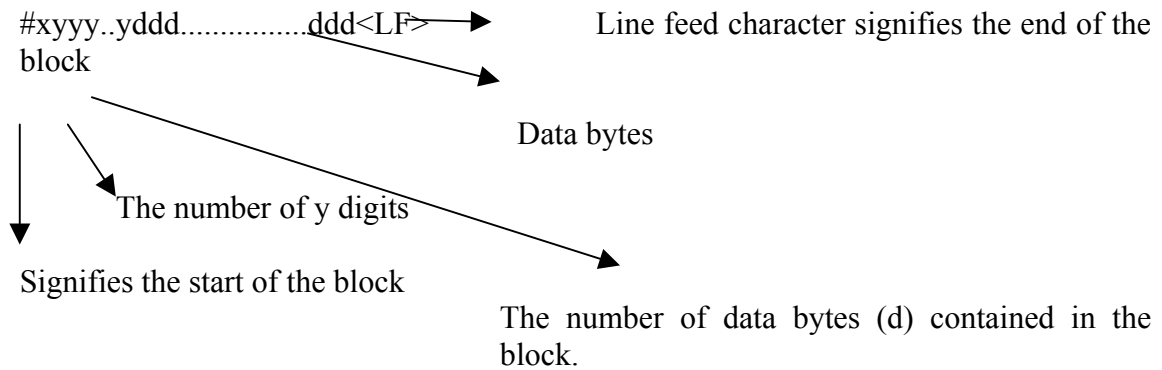
Four probes have therefore been tested with CW and pulse-modulated waveforms. Those probes employing thermocouple detection showed almost constant sensitivity when their readings were compared with the average power flux density but were not as sensitive as diode based probes and were not able to offer peak detection or measurement of individual sensor axes. Probes employing thermocouple detection are therefore the preferred choice when making measurements on stationary radar systems with RMS levels greater than 20 V/m.

The tests on a diode based probe showed that the sensitivity varied considerably according to the Duty Cycle. Sensitivity variations of +7 dB to -1.7 dB were observed as the duty cycle was varied from 0.05 to 4%. The variation in sensitivity with applied level varied by ± 2.5 dB. Using the peak hold function, it was possible to calibrate the probe for burst pulse operation. This would enable the diode probe to measure rotating radar systems. The correction for desensitisation was measured and could be up to 100, or 20 dB for illumination times of less than 5 μ s. The desensitisation factor could be calculated within limits of +0.7 to -1.7 dB using the response time of the probe and meter where this is known. Diode based probes are less expensive than thermocouple probes and are essential if measurements need to be made on rotating radars. They can be made to read with sufficient accuracy by careful calibration for specific radar parameters. These are: pulse width, PRF, illumination time and indicated level.

A frequency selective probe was tested and showed good agreement, within 0.7 dB with the NPL system. As well as Peak and Average Power, the probe was also able to measure the carrier frequency, PRF and Pulse Width. This type of probe is the most expensive but is the most versatile and comprehensive field measurement system.

8 Appendix 1

Data from the gated memory is returned in IEEE 488.2 arbitrary block program data format as follows:



Example: if there are 12435 data bytes, $y = 12435$ and $x = 5$
 Each data point in the memory is made up of four bytes to form a data block. The MS byte is transmitted first.

To efficiently decode the IEEE 488.2 arbitrary block format it was necessary to use pointers in Visual Basic to enable variable addresses to be identified and used to store binary data in specific memory locations. The following VB program defines these routines which are not directly available in the VB editing platform.

Option Explicit

```
Private Declare Sub CopyMemory Lib "kernel32" _
    Alias "RtlMoveMemory" (Destination As Any, _
    Source As Any, ByVal Length As Long)
Private Declare Function GetProcessHeap Lib "kernel32" () As Long
Private Declare Function HeapAlloc Lib "kernel32" (ByVal hHeap As Long, ByVal dwFlags As Long, ByVal dwBytes As Long) As Long
Private Declare Function HeapFree Lib "kernel32" _
    (ByVal hHeap As Long, ByVal dwFlags As Long, lpMem As Any) As Long
Private Declare Sub CopyMemoryWrite Lib "kernel32" Alias _
    "RtlMoveMemory" (ByVal Destination As Long, _
    Source As Any, ByVal Length As Long)
Private Declare Sub CopyMemoryRead Lib "kernel32" Alias _
    "RtlMoveMemory" (Destination As Any, _
    ByVal Source As Long, ByVal Length As Long)
```

Private Sub Form_Load()

```
Dim ptr As Long 'dim a pointer (memory address) which needs 2 bytes
Dim hHeap As Long
hHeap = GetProcessHeap()
ptr = HeapAlloc(hHeap, 0, 2) 'allocate 2 bytes for the pointer
Dim i(4) As Byte
    'The number is 123.5 which in 32bit FP can be represented by 4 by 8 bit bytes
```

```

i(1) = 66 ' MSB
i(2) = 247
i(3) = 0
i(4) = 0 'LSB

CopyMemoryWrite ptr, i(4), 1 ' i is a single 8 bit byte LSByte first
CopyMemoryWrite ptr + 1, i(3), 1 ' pointer address +1
CopyMemoryWrite ptr + 2, i(2), 1 ' pointer address +2
CopyMemoryWrite ptr + 3, i(1), 1 ' pointer address +3

Dim j As Single ' (32 bit floating point variable)
CopyMemoryRead j, ptr, 4
MsgBox "The address of ptr is " & CStr(ptr) & _
      vbCrLf & "and the value is " & CStr(j)
HeapFree GetProcessHeap(), 0, ptr
End Sub

```

This program takes a known number 123.5 and inputs the four 8 bit bytes. The Intel representation of numbers stores the LSB first, hence the LSB (Binary 66) is copied into memory location (ptr). Finally variable J is defined as a Single type, which corresponds to the 32 bit floating point format and the binary number from address (ptr) is copied to J. The message box displays the decoded number and the address of (Ptr).

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