con through fabrication service (presently not available to us) should produce a correctly functioning chip. The logic circuits of Fig. 2 are constructed using off-the-shelf IC's and are found to function correctly.

V. CONCLUSION

Compact, all-digital functional organization, special clocking and circuit techniques to obtain glitch-free operation, widely programmable built-in digital delay network, and the facility to cascade are the distinguishing features of the SACHIP. The design is considered a good ASIC candidate for implementation in silicon and to replace bulky, off-the-shelf signal annunciators on PCB's and to make smarter monitor/control panels for instrumentation.

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A Wide-Band FET Antenna and its Calibration

Toshimi Matsui

Abstract—**A small and very wide-band dipole antenna employing FET's with a bandwidth exceeding 200 Hz to 1000 MHz has been developed. The antenna factor in free space has been determined by using a TEM cell and the standard antenna method with error within ±3.2%.**

I. INTRODUCTION

For precision EMI measurements or waveform measurements of pulsed hazard radiation in free space, a small and very wide-band field sensor is required. A dipole antenna employing field effect transistors (FET) is adapted for these measurements in the HF to UHF range [1]. However, there were important problems in the calibration of the antenna at frequencies lower than 50 MHz.

A TEM cell is practical for antenna calibration equipment at low frequencies [2]. However, there is little reliability information in calibrated results by the method, because there were no experiments for precise comparison between the antenna factors measured with the TEM cell and those measured in free space. The TEM cell can only be used at very low frequencies, i.e., lower than its resonance frequency which is determined by its size. On the other hand, standard dipole antennas for these frequencies are usually very long, and could not fit into the cell.

The author has developed a very wide bandwidth short dipole antenna employing FET transistors, which can be used to determine the antenna factor by both methods at the same frequency. As the result, the characteristics of the TEM cell calibration system have been confirmed, and the antenna factor of the antenna has been calibrated at frequencies from 100 Hz to 1300 MHz.

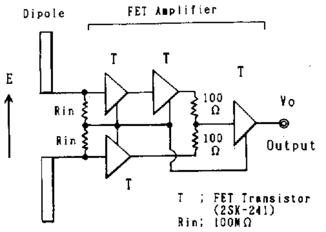

Fig. 1. Principle and construction of the FET antenna.

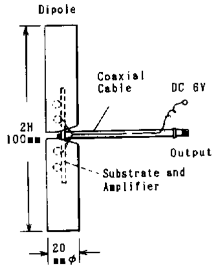

Fig. 2. Shape and size of the FET dipole antenna.

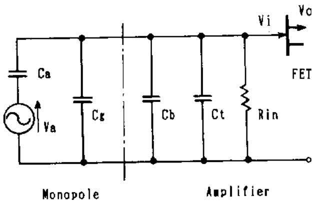

Fig. 3. Equivalent circuit at the input of the FET dipole antenna.

II. THEORY AND DETAILS OF THE FET ANTENNA

The theory and construction of the developed FET antenna, its shape and size, and the equivalent circuit for the input side of the antenna, etc., are shown in Figs. 1-3, respectively. The electrically short dipole antenna, which is much shorter than a wavelength, can be represented by an equivalent circuit consisting of a source V_a and a capacitance C_a as shown at the left hand side of Fig. 3 [3]. The capacitance C_a at frequencies much lower than the resonant frequency of the antenna is very small, on the order of 1 to 10 pF for practical short dipole antennas, and is independent of frequency. Thus if induced voltage is measured with a normal receiver of a 50-Ω impedance system, almost all the voltage is lowered by the capacitance, and its sensitivity and the flatness are decreased.

However, if the induced voltage is detected by an amplifier with a high impedance device such as an FET transistor, as shown in Figs. 1 and 3, the voltage is measured with little drop. The output voltage of the FET in Fig. 3 is given by

$$V_o = \frac{G_a V_a}{1 + \frac{C_{\text{in}}}{C_a} + j\frac{1}{\omega C_a R_{\text{in}}}} \tag{1}$$

where G_a, C_a, and R_{in} are the gain of the amplifier, the capacitance of the monopole (twice the capacitance of the dipole antenna), and the input resistance of the amplifier. C_{in} is a total of capacitances C_g, C_b, and C_t, which are the gap capacitance between an end-plane of the monopole and ground, the capacitance between the mono-

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pole and a substrate, and the input capacitance of the FET, respectivily. From (1), the low cutoff frequency and the gain in the middle frequency region are given by

$$F_{cl} = \frac{1}{2\pi R_{\text{in}}(C_a + C_{\text{in}})} \tag{2}$$

$$G_c = \frac{C_a}{C_a + C_{\text{in}}} \tag{3}$$

Capacitance C_a and C_{in} of the antenna were of the order of 3.5 pf and 10 pf, respectively. The gain of the FET amplifier at middle frequency was about 0.5 until about 300 MHz and then it inversely decreased at higher frequencies. The FET's and all elements are in the dipole to prevent unwanted pickup and to minimize the size.

III. MEASUREMENTS

For the antenna coefficient of the dipole antenna the antenna factor is generally used and is defined as [3]

$$AF = \frac{E}{V_o} \tag{4}$$

where E and V_o are the field strength to be measured and the output voltage of the antenna with its impedance matched to the standard coaxial line.

The antenna factors AF for 100 Hz to 100 MHz were measured with the TEM cell as shown in Fig. 4(a). When the voltage between a center plate and an upper plate of the TEM cell spaced by h is V_t, the field strength E is given by V_t/h. A 45 cm × 45 cm × 30 cm (height) TEM cell was used for the measurement and its theoretical resonance frequency is about 680 MHz.

The value of AF for 60 MHz to 1300 MHz was measured with the standard antenna method as shown in Fig. 4(b) in a 15 m × 10 m × 8 m (height) anechoic chamber. Three dipole antennas of length of 1, 0.3, and 0.1 m, which were calibrated by the three antenna method in the anechoic chamber with an error within 2.5% at 100 MHz or higher frequencies, were used as the standard antennas. The FET antenna has been compared with an error within ±1% relative to these antennas.

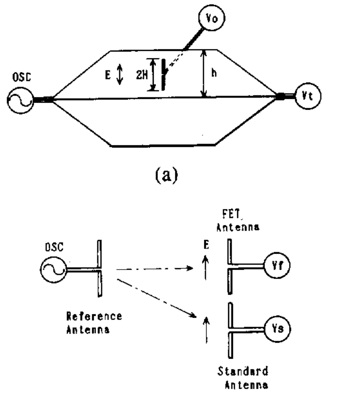
IV. MEASUREMENT RESULTS

The frequency dependence of the AF of the FET antenna measured by the three-antenna method and by the TEM cell method are shown by lines A and $B1$ in Fig. 5, in terms of inverse values of the antenna factor. Both lines show that the 3-dB cutoffs of the antenna are about 150 Hz and 1200 MHz, respectively. Curve A shows that the AF for about 60 to 200 MHz is very stable, and is lowered at higher frequencies. However, it rises again near 700 MHz and shows a peak near 1000 MHz due to the resonance of the antenna.

The antenna factor of line $B1$ at middle frequency from the low frequency cutoff to about 60 MHz is flat without detectable deviation, and decreases at higher frequencies. This decrease, however, is not due to the characteristics of the antenna, but is due to the effects of inserting the antenna and its lead wire into the TEM cell. From the bandwidth measured in free space and shown in curve A, if we could obtain the ideal TEM cell measurement, flatness up to about 300 MHz will be expected, as indicated by a line $B1'$. In this case, it is expected that lines A and $B1'$ will coincide with each other if both are measured with no errors. However, in practice there is about 17% difference between these values.

This difference is probably due to the effect of the size of the antenna compared to the size of the TEM cell. This was confirmed from the measurement of a length dependence of the induced voltage on a monopole antenna in the TEM cell, as shown in Fig. 6. Curves A and B are results for monopoles with diameters 20 mm and 1 mm. In the region where the height H is very short compared to the space h of the TEM cell (H/h is 0.3 or less) the induced voltage V_i is proportional with H as expected theoretically from (4). A straight line C is an asymptotic line to the curve in the region. However, when the height H is half of h or larger, the induced voltage increases faster than the straight line C, and the difference becomes larger when H is very close to h. This characteristic also depends on monopole thickness, as indicated by lines A and B.

At point of $H/h = 0.66$, the case of our experiment, voltage V_a on curve A is about 18% higher than the voltage V_c on line C. Curve $B2$ of Fig. 5 is a corrected line of curves $B1$ and $B1'$ with this effect. Curves A and $B2$ coincide with each other within about ±0.5%. In the case of the TEM cell measurements, however, error within about ±1.5% was estimated from the flatness of the field in



(a)

(b)

Fig. 4. Measurement of the antenna factor. (a) The TEM method. (b) The standard antenna method.

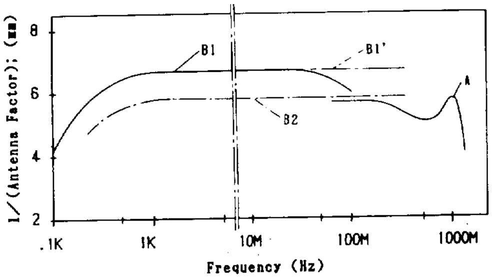


Fig. 5. Frequency dependence of measured equivalent length of the FET antenna.

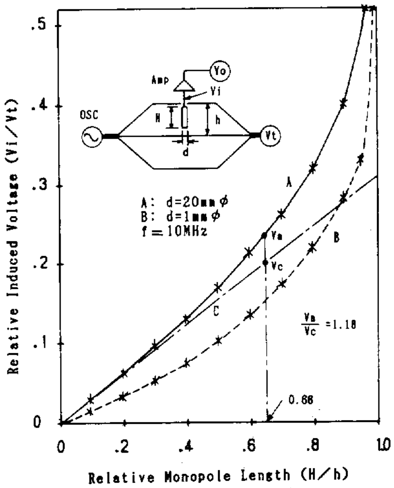


Fig. 6. Length dependence of the induced voltage to the monopole antennas in the TEM cell.

TABLE I
ERRORS IN ANTENNA CALIBRATION WITH THE TEM CELL

Individual Errors	Error (%)
Error in TEM measurement	±1.5
Error in comparison with standard antenna	±1.0
Error in standard antenna calibration	±2.6
Total (rss)	±3.16

the TEM cell, the mismatch between parts, and the error of power calibration error, etc. Therefore, we conclude that the antenna factor of the FET antenna in free space was calibrated within about ±3.2%. Each error in the measurement is listed in Table I.

V. CONCLUSION

A small very wide dipole antenna using FET's with bandwidth wider than for 200 Hz to 1000 MHz has been developed. The characteristics of the calibration system using a TEM cell were determined by using this wide band antenna. As the result, the antenna factor of the antenna in free space for this frequency range was determined with an error within about ±3.2 percent.

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Calorimetric Output Power Measurements on a CW 20 kW 7.16 GHz Microwave Transmitter

Raul M. Perez and Bruce L. Conroy

Abstract—**The calorimetric measurement of the output level of high power microwave transmitters has traditionally required knowing the thermal constants of the cooling medium to a high degree of precision. This paper describes a calorimetric measurement technique developed for NASA's Deep Space Network (DSN) transmitters that does not require data on the coolant's thermal parameters. Calibration of the measurement system is achieved by measuring the DC input power to the klystron and relating coolant temperature increases to this known power dissipation.**

I. INTRODUCTION

The operation of a high power microwave transmitter involves large power dissipations, mainly due to the efficiency of the transmitting tube (klystron) in the range of 30–40%. For this reason, these systems are water-cooled using a closed loop heat exchanger. In addition, when operating in a laboratory environment where radiation is not permissible, the tube output must also be dissipated in a water load. Since all the output power of the transmitter is dissipated in the cooling water, direct calorimetric measurement of the output power level is possible. To accomplish this, the rate of flow through the water load, the temperature rise of the cooling water, and the thermal constants of the coolant must be measured. In order to make the measurement of output power independent from the coolant's thermal parameters, a measurement system (see Fig. 1) was developed which uses the dc input power of the tube as the calorimetric calibration standard.

II. GENERAL METHOD

The relation used to compute a power dissipation as a function of coolant flow and temperature rise is [1]:

$$P = KF(T_o - T_i) \tag{1}$$

where,

P	Dissipated power in kW
K	Cooling constant of medium in (kW × Minutes)/(Gallons × Deg C)
T_o	Outlet temperature of coolant in degrees C
T_i	Inlet temperature of coolant in degrees C
F	Coolant flow in gallons per minute.

Both coolant flow and temperatures are readily measurable to a high degree of accuracy, using precision turbine flowmeters and resistance temperature detectors (RTD's). The cooling constant K is a function of the specific gravity and specific heat of the cooling medium, which are in turn functions of the temperature and composition of the coolant. As such, this cooling constant is not directly measurable in an electrical measurement environment. This difficulty is avoided by making use of the fact that, in operating a klystron amplifier, a condition of known dissipation exists when the high voltages beam supply is applied but the RF drive to the amplifier is not. In this case, all the beam supply power is dissipated in the collector of the klystron (about 50 kW for a 20 kW output power transmitter). Since the collector dissipation is accurately known from voltage and current measurements, (1) can be solved for K, obtaining:

$$K = \frac{P'}{F'(T'_o - T'_i)} \tag{2}$$

where,

P'	Collector dissipation with no RF output, in kW
F'	Coolant flow with no RF output, in gallons per minute
T'_o	Collector output temperature with no RF output, in degrees C
T'_i	Collector inlet temperature with no RF output, in degrees C.

Substituting the K computed from the collector dissipation into (1), and using the equalities:

$$T = T_o - T_i \tag{3}$$

$$T' = T'_o - T'_i, \tag{4}$$

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