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Broadband RF current detector

George Cabrera

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FLORIDA INTERNATIONAL UNIVERSITY

Miami, Florida

BROADBAND RF CURRENT DETECTOR

A thesis submitted in partial satisfaction of the
requirements for the degree of

MASTER OF SCIENCE

IN

ELECTRICAL ENGINEERING

by

George Cabrera

1994

To: Dr. Gordon Hopkins
Dean, School of Engineering & Design

This thesis, written by George Cabrera, and entitled Broadband RF Current Detector, having been approved in respect to style and intellectual content, is referred to you for judgement.

We have read this thesis and recommend that it be approved.

Dr. Malcolm Heimer

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Date of Defense: November 28, 1994

The Thesis of George Cabrera is approved.

Dr. Gordon Hopkins
Dean, School of Engineering & Design

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Florida International University, 1994

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ABSTRACT OF THE THESIS

BROADBAND RF CURRENT DETECTOR

George Cabrera

Florida International University, 1994

The problems to be solved in this thesis were 1) development of a broadband RF preamplifier to be used with non-ferrous current probes so that the amplified signal exceeds the errors due to cable pickup, no detection is needed in this application, and 2) development of a self-contained device that amplifies and detects the output from a non-ferrous current probe, providing a digital readout of the current. These instruments have been completed and are being tested for use by the National Institutes of Occupational Safety and Health (NIOSH). The self-contained current meter operates at frequencies up to 600 MHz, and detects currents as low as 8 mA . At these current magnitudes, the probe (pick-up coil) will output a voltage of 500 μ V (-53 dBm on 50 Ω) which will have to be raised above 0 dBm. The final circuit uses a RF mixer as a variable attenuator in order to increase the dynamic range, two Monolithic Microwave Integrated Circuits (MMIC) for preamplification, a final broadband amplifier to raise the output compression point, a

Schottky diode detector, a sample and hold circuit, and a liquid crystal digital panel meter.

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1. Introduction

1.1 Electromagnetic Hazards and the Need for Dosimeters.

There is a growing interest, by the general public (Foster, 1992; Jauchem, 1993) as well as by engineers and other professionals (Albanese *et al.*, 1994; Gochfeld, 1990; Goodman *et al.*, 1993; Guy, 1992; Repacholi, 1990; Tenforde, 1992; Wilkening *et al.*, 1990; Yost, 1992), in the possible hazards from exposure to electromagnetic radiation. Popular books such as "The Zapping of America" (Brodeur, 1977), "Currents of Death" (Brodeur, 1989), and "The Great Power-line Cover-Up" (Brodeur, 1993) appear to have increased the public concern. Furthermore, while guidelines limiting human exposure have been established by numerous governmental agencies (e.g. Canada, 1979; Commonwealth of Massachusetts, 1983; World Health Organization, 1981) and private groups (e.g. ACGIH, 1988; ANSI C95.1, 1982; IEEE, 1992; IRPA, 1984; NRPB, 1986), the specialists have not yet reached a full consensus and their disagreement is evident from the differences in these various standards.

While the exact effects of low-level exposure on the human body are yet unknown, it is certain that the absorption of electromagnetic energy can heat tissues (thermal effects) and that intense fields can also cause shock and burn. Guidelines for human exposure which were made by the IEEE (IEEE, 1992) are based on the known hazards of heating (Cabanac, 1983; Mumford, 1969; Schwan *et al.*, 1980; Stolwijk, 1983), shock (Chatterjee *et al.*, 1986; Dalziel *et al.*, 1950; Dalziel *et al.*, 1969), and burn (Dalziel *et al.*, 1969; Rogers, 1981), and do not seriously consider the possible but as yet unquantified effects of low-level exposure (Adey *et al.*, 1982; Albert *et al.*, 1981; Anderson, 1993; Baranski *et al.*, 1976; Bawin *et al.*, 1977; Blackman *et al.*, 1985; Chou *et al.*, 1982; Cleary, 1993; Foster *et al.*, 1974; Frey, 1971; Frey *et al.*, 1983; Froehlich, 1975; Grundler *et al.*, 1977; Jensh, 1984; Lin, 1980; Lords *et al.*, 1973; Marg, 1991; Pressman, 1970; Sher, 1963; Taylor, 1981).

There are fundamental problems in determining the effects of exposure of the human body to low-level electromagnetic fields. For example, epidemiological studies in which the incidence of various diseases is correlated with possible environmental factors are complicated by the difficulty in quantifying the long-term exposure of

people to electromagnetic fields (Anderson, 1991; Garland *et al.*, 1990; Hamburger *et al.*, 1983; Hendee *et al.*, 1994; Kallen *et al.*, 1982; Lamarine *et al.*, 1992; Lancranjan *et al.*, 1975; Lester *et al.*, 1982; Loomis *et al.*, 1994; Milham, 1982; Muhm, 1992; Neutra, 1992; Schnorr *et al.*, 1991; Schreiber *et al.*, 1993; Shaw *et al.*, 1993; Tabrah *et al.*, 1991; Vena *et al.*, 1991; Walsh *et al.*, 1991). Furthermore, practical considerations make it difficult to conduct an animal study having a duration much greater than one month, and it is hard to justify extrapolating the available results of the relatively short-term animal studies to the lifetime exposure of humans to low-level electromagnetic fields (Abhold *et al.*, 1981; Adair *et al.*, 1985; Chou *et al.*, 1983; Guy *et al.*, 1980; Smialowicz *et al.*, 1981).

The guidelines for human exposure which were made by the IEEE Standards Coordinating Committee 28 (IEEE, 1992) suggest that the hazards due to heating of the human body should be quantified by determining the SAR (specific absorption rate, in Watts per kilogram of body mass), (Justesen, 1975). These guidelines further state that the hazards due to shock and burn should be evaluated by determining the current flowing through the human body. An upper limit of 0.4 W/kg was chosen as the working basis for setting limits of the MPE (maximum

permitted exposure, values of electric field, magnetic field, and equivalent power density to which a human may be exposed). Thus, an upper limit for the SAR is specified, but since there is no noninvasive means for determining the SAR (Tell, 1990) the MPE may be overly restrictive in some cases and may provide insufficient protection in others (Gandhi *et al.*, 1985). The limit of 0.4 W/kg and the tables of MPE in the guidelines have been used to compare the results of numerical simulations (Durney, 1980; Hagmann *et al.*, 1979A; Hagmann *et al.*, 1979B), invasive thermal measurements with animals (Krupp, 1983), and thermal measurements with figurines made of tissue-simulating materials (Gandhi, 1975; Guy, 1971; Guy *et al.*, 1976; Olsen *et al.*, 1989).

The IEEE guidelines for human exposure (IEEE, 1992) also specify upper limits for the foot current (measured between the feet and ground) and the contact current (measured between the hand and a conducting object), but no limits are given for the currents elsewhere in the human body because the technology for noninvasive measurements of body current were not considered by the IEEE Standards Coordinating Committee. Separate guidelines are given for controlled environments (where people are aware of the potential for exposure or passage is

transitory) and uncontrolled environments (where people may have no knowledge or control of their exposure). For uncontrolled environments the MPE for induced and contact RF currents at frequencies from 3 to 100 kHz is specified as $450 f$ mA measured through either foot or through either hand, and as a total of $900 f$ mA through both feet, where f is the frequency in MHz. At frequencies from 100 kHz to 100 MHz the MPE is specified as 45 mA through either foot or either hand, and as 90 mA through both feet. For controlled environments the MPE at frequencies from 3 to 100 kHz is specified as $1,000 f$ mA measured through either foot or through either hand, and as $2,000 f$ mA through both feet. At frequencies from 100 kHz to 100 MHz the MPE is specified as 100 mA through either foot or either hand, and as 200 mA through both feet. All current values are given as rms.

A summary of typical values for the current induced in human subjects may help to put the MPE values in perspective. For example, an adult human exposed to a vertically polarized plane wave at 40 MHz has a foot current as large as 13 mA per V/m in the incident field (Gandhi *et al.*, 1986). Measured values of the foot currents in workers using RF sealers are as large as 330 mA (Gandhi *et al.*, 1986). Personnel exposed to a 100

kV/m Electromagnetic Pulse simulator (EMP, simulating the intense fields generated by a nuclear detonation) from a vertical dipole source have a peak current as high as 500 A in the legs (Guy, 1989). Despite the large value, due to the short duration of the EMP it has been observed that a peak current of 10 A in the arm may cause no sensation, and 100 A produces only a slight tingling sensation in the hand (Gronhaug, 1986).

There is a need for a means of characterizing the exposure of humans to electromagnetic fields in various settings. That is, a device like the dosimeters that are generally used with ionizing radiation. At present there are field probes that may be used to measure the strength of electric and magnetic fields at radio frequencies (Aslan, 1971; Bassen; 1983; Green, 1975; Herman *et al.*, 1980; Rudge *et al.*, 1970). Thus, the fields incident on the human body may be measured. However, the transfer function relating these incident fields to the fields induced in the human body is quite complex, and depends on the frequency, polarization, and distance of the source, as well as the height, build and posture of the human subject (Durney *et al.*, 1986; Gandhi *et al.*, 1985; Hagmann *et al.*, 1979A; Hagmann *et al.*, 1979B). The MPE in the IEEE guidelines are estimated limits for electric field, magnetic field, and equivalent power

density such that the SAR at will not exceed 0.4 W/kg at any point in the human body. However, due to the extreme variability of the transfer function, the incident fields are not an accurate predictor of a potential hazard and so it appears that the MPE values may be overly restrictive in some cases and may provide insufficient protection in others (Gandhi *et al.*, 1985).

While it appears that a dosimeter is required in order to determine human exposure to electromagnetic fields, it is known that even the invasive measurement of SAR is difficult because conventional thermometry is incompatible with strong RF fields. For example, RF currents induced in a temperature probe can cause an erratic response by the readout electronics, and can also cause ohmic heating of the probe, thus changing the SAR and causing burns if the probe is implanted in tissue. Several types of temperature probes have been designed that are compatible with RF fields. These include 1) a thermocouple with high impedance leads (Bowman, 1976), 2) color changes in a liquid crystal (Johnson *et al.*, 1975), 3) a fiber-optic probe based on the temperature dependence of the attenuation of light in a semiconductor (Christensen, 1977; Rozzell *et al.*, 1974; Windhorn *et al.*,

1979), and 4) fiber-optic "fluoroptic" probes based on the temperature dependence of the emission of light by phosphors (Giallorenzi *et al.*, 1986; Kolodner *et al.*, 1983; Wickersheim, 1990). These RF-compatible temperature probes have been implanted in animal models for determining the SAR, and in human subjects exposed to electromagnetic fields for therapy, but it would not be practical to use invasive thermometry for routine dosimetry with humans.

1.2 Previous use of RF Current Measurements for Dosimetry.

It has been suggested that measurements of the RF current at a point of contact with the human body, either between the feet and a ground plane (Chatterjee *et al.*, 1986; Gandhi *et al.*, 1986; Hill *et al.*, 1985) or at another point of contact such as the hand (Chatterjee *et al.*, 1986; Stuchly *et al.*, 1991), could be used as the basis for human dosimetry. Instruments for dosimetry that are based on this principle have been patented (Gandhi, 1987), and are presently sold by two different manufacturers (Holaday Industries Inc., Eden Prairie, MN; Loral Microwave-Narda, Hauppauge, NY). It was already mentioned in this thesis that the IEEE guidelines for human exposure (IEEE, 1992) specify upper limits for

these currents as well as for the SAR. These limits are below measurements of the threshold currents for perception (Chatterjee *et al.*, 1986), and are considerably less than the values causing painful shocks (Chatterjee *et al.*, 1986; Dalziel *et al.*, 1950; Dalziel *et al.*, 1969), and burns (Dalziel *et al.*, 1969; Rogers, 1981). Furthermore, the value of the SAR near the point where the current is measured could be estimated by calculating the RF power using the measured current with known values for the dielectric properties of human tissues (Stoy *et al.*, 1982). Indeed, there is some evidence that for frequencies below 50 MHz the greatest value of the SAR may occur in the lower leg, near where the foot current would be measured (Gandhi *et al.*, 1985).

Measurements were made using the stand-on current probes produced by both Holaday Industries Inc. and Loral Microwave-Narda with human subjects exposed to RF fields in an anechoic chamber at Brooks Air Force Base, Texas (Hagmann, 1994). These tests showed that 1) both of these probes act as antennas, responding to RF fields with erroneous current readings when there is no human subject, 2) both probes are unreliable when the human subjects stand on a wood floor instead of on a large metal ground plane, and 3) while the probes are

rated for use at frequencies up to 100 MHz, the readings do not correlate with true values of the current except when the frequency is at or below 60 MHz and the human subjects stand on a large metal ground plane. The observed failure of these current probes at frequencies above 60 MHz makes them unsuitable for use at commercial FM facilities. Furthermore, the requirement that the subject must be standing on a large metal ground plane makes the stand-on current probes inappropriate for use in most common industrial settings where the floor may be made of wood or concrete. Numerical simulations (Hagmann *et al.*, 1979A; Hagmann *et al.*, 1979B) have shown that for frequencies less than 200 MHz, when a human subject is standing on a ground plane, the total foot current generally exceeds that at any location within the human body. However, when the subject is not in contact with the ground plane, the current at the base of the feet is null. These results may be understood by using antenna theory to model the human as a monopole antenna at the ground plane, and as a dipole above the ground plane, respectively. Thus, the stand-on current probes could not be used with human subjects on wooden scaffolds, climbing antennas, or in other situations in which the subject would not be in contact with ground. All of the current probes that measure only the current at a point of contact with the human body

fail to indicate values of the current at other points within the body which may be more significant regarding potential hazards (e.g. organs such as the heart, etc.) and this is especially objectionable when the subject is not in contact with ground.

1.3 Previous use of Ferrous Current Probes for Dosimetry.

Ferrous current probes have been used to determine the current induced in the limbs of human subjects exposed to both CW (Blackwell, 1990) and pulsed EM sources (Gronhaug, 1986; Gronhaug, 1988), and they have suggested that these probes may be used for dosimetry. Ferrous current probes suitable for measurements over a wide range of frequencies (20 Hz to 1 GHz) are available commercially (e.g. Eaton Corporation, Los Angeles, CA; EG&G Inc., Albuquerque, NM; Pearson Electronics Inc., Palo Alto, CA; TEGAM Inc., Geneva, OH). These instruments (Miller 1986; Ricketts *et al.*, 1976), which may be considered to be an extension of the clamp-on ammeters in common usage at power frequencies, are toroidal transformers which must be placed so that the circular opening (the window or aperture of the probe) is around the object in which the current is to be measured. The current passing through

the aperture acts as the primary circuit for the transformer, and a small secondary winding serves as the output port which is connected to an oscilloscope or other readout device (see Appendix I). Thus, unlike the stand-on and other types of contact current probes, the ferrous current probes may be used to measure the current at various points within the human body, and are also usable for measurements in a person that is not in contact with ground.

While ferrous current probes may be used at frequencies up to 1 GHz, the maximum usable size for the probe window varies inversely with frequency. For example, commercial probes that are usable to 100 MHz have a maximum aperture diameter of 12.7 cm which could accommodate the human ankle or forearm. Probes that are usable to 1 GHz have a maximum aperture diameter of 3.8 cm (TEGAM Inc., Geneva, OH) which could only accommodate a finger. The size limitation, which is fundamental to ferrous current probes, is due to the requirement that the length of the ferrous toroid must be much less than a wavelength. A further difficulty with ferrous current probes is that they perturb the circuit in which they are used by introducing an insertion impedance. It has been shown by both calculations and measurements

(Hagmann *et al.*, 1993) that these perturbations may cause significant errors in measuring the RF currents induced in the human body. The reactive component of the insertion impedance is unusually large when ferrous current probes are used in dosimetry because a large fraction of the aperture is filled, rather than a single wire. Furthermore, while others have suggested that ferrous current probes may be used for human dosimetry (Blackwell, 1990; Gronhaug, 1986; Gronhaug, 1988), the weight of these probes (about 3.6 kg for an ankle-sized probe) would make it impractical to wear them during daily activities.

1.4 Previous use of Non-Ferrous Current Probes for Dosimetry.

Non-ferrous toroidal current probes have also been used to measure RF currents (Baum *et al.*, 1978; Hagmann *et al.*, 1993). As with the ferrous probes, non-ferrous toroidal current probes also act as transformers in which the current-carrying element which is placed in the aperture acts as the primary circuit (see Appendix II). Thus, unlike the stand-on and other types of contact current probes, both ferrous and non-ferrous current probes may be used to measure the current at various

points within the human body, and are also usable for measurements in a person that is not in contact with ground.

Non-ferrous current probes have a number of advantages relative to the ferrous current probes. 1) The maximum aperture diameter for a non-ferrous probe may be several times greater than that for a ferrous current probe designed to be used at the same frequency (Hagmann *et al.*, 1993). This difference is attributed to the increased velocity of propagation on the toroid due to the lower value of relative permeability. 2) It has been shown by both calculations and measurements (Hagmann *et al.*, 1993) that errors due to perturbation of the circuit by insertion impedance are much less for non-ferrous probes. This difference may be understood in that the added inductance is decreased due to the lower value of relative permeability. 3) The non-ferrous current probes are inherently more accurate because the sensitivity is proportional to the permeability of free space, which is a fundamental constant, instead of the permeability of the ferrous core. 4) The non-ferrous current probes are much simpler to construct so they should be less expensive, whereas the ferrous current probes typically cost from \$2,000-\$3,000. Finally, the non-ferrous current probes have much lower weight, which is essential in order

for a dosimeter to be worn and carried. The disadvantages of the non-ferrous current probes are that 1) the sensitivity is lower due to the decreased value of mutual inductance, and 2) the winding of the secondary (output) coil must be evenly distributed over the full length of the toroid (see Appendix II) so this coil must be resistively loaded (Typical value of 30 k Ω) in order to limit the effects of coil resonances on the flatness of the frequency response. The loading could also be accomplished by adding a shunt conductance to the coil, but a series resistance is preferred in order to further reduce perturbations by insertion impedance.

Professors Hagmann and Babij developed non-ferrous toroidal current probes at Florida International University during the period from 1986-1991, and these instruments were the first inventions to be patented at the University (Hagmann *et al.*, 1990A; Hagmann *et al.*, 1990B). A typical probe designed to measure currents induced in the human thigh is contained within a toroidal aluminum shield that has an inner diameter (aperture diameter) of 22 cm, an outer diameter of 30 cm, and a height of 5 cm. Inside the shield there is a coil with 200 turns of resistive line, having a total resistance of 33.2 k Ω , that is wound evenly over the full

length of a Plexiglas toroidal core having a diameter of 1.9 cm and a radius of curvature of 13.0 cm.

The equations in Appendix II may be used to determine the open-circuit (unloaded) response of the non-ferrous current probes at low frequencies. Being careful to allow for the thickness of the resistive line when calculating the area of the winding, the mutual inductance between the coil winding and a current passing through the aperture would be approximately 100 nH, and the magnitude of the transfer impedance (the ratio of the output from the probe, the potential measured across the resistive coil, to the current) would be approximately 0.63 m Ω /MHz. We have measured a mutual impedance of 300 pH $\pm$ 12 % in the frequency range of 1 to 200 MHz, with no substantial roll-off in sensitivity until the frequency exceeds 250-300 MHz. This corresponds to a transfer impedance (the ratio of the output from the probe, the potential measured across the resistive coil, to the current) of 1.90 m Ω /MHz $\pm$ 12 % over the frequency range of 1 to 200 MHz. The measured values are within 1 % of the theoretical values when corrected for voltage division by the output load (one 50 Ω load on each terminal of the winding) relative to the winding resistance. The observed proportionality of the

transfer impedance to the frequency, required by Faraday's law of induction, may be corrected by adding an integrator to provide a flat response from 1 to 200 MHz.

By comparison, the model 94606-2 current probe (Eaton Corporation, Los Angeles, CA) which appears to be the commercially available ferrous current probe having the largest aperture size (12.8 cm diameter, accommodating only a human ankle). Ferrous current probes typically contain an RC integrating circuit in order to provide a frequency response that is relatively flat. This probe has a transfer impedance of $1.0 \Omega \pm 20 \%$ in the frequency range from 300 kHz to 70 MHz, with appreciable oscillations at frequencies over 100 MHz and appreciable roll-off at frequencies below 100 kHz. These values confirm, as already noted in this thesis, that the non-ferrous current probes have the advantage that they may be several times larger in diameter than a ferrous current probe designed to be used at the same frequency. However, these values also confirm that the non-ferrous probes have a lower sensitivity due to decreased mutual impedance due to the lower value of relative permeability.

Non-ferrous probes constructed at Florida International University, which were similar to the current probe already described which was designed to measure currents induced in the human thigh, were used successfully to measure the currents induced in the arms and legs of human subjects exposed to the fields of Electromagnetic Pulse (EMP) simulators. Measurements were made at the Air Force Los Alamos Laboratory Electromagnetic Pulse Calibration and Simulation (ALECS) facility, and the Navy EMPRESS I installation (Hagmann, 1993). At each of these test facilities a RF/optical transducer was connected directly to each non-ferrous current probe, and a fiber optic data link was used to connect to the A/D converters in the main-frame computer used for data analysis. The current was measured as a function of time with a usable bandwidth exceeding the 200 MHz limit of the current probe.

In all of the EMP tests the non-ferrous current probes were found to have negligible readings when no current-carrying element was present within the aperture, even in field strengths of 50 kV/m. Furthermore, when non-ferrous current probes were placed on vertical pipe, acting as an antenna, the current readings agreed with the ferrous probes that are conventionally used (Miller, 1986; Ricketts *et al.*, 1976) in EMP testing. By

contrast, tests made at the EMP facilities with stand-on current probes showed that they had large null readings (erroneous outputs due to interference when there was no conductor on the platform), and the indicated currents were not consistent with ferrous probes when the stand-on and ferrous probes were placed on the same antenna (Hagmann, 1993).

1.5 Difficulties in using Non-Ferrous Current Probes.

Tests of the non-ferrous probes have not been as successful when a fiber-optic data link was not available. For example, open field tests were made using the non-ferrous probes with a man-sized phantom at the Naval Aerospace Medical Laboratory (NAMRL) in Pensacola, Florida. In these tests the phantom was positioned to stand on a metal ground plane a distance of 1 m from a 10 m vertical monopole antenna, and a power of 1 KW at 29.9 MHz was fed to the antenna. Good agreement was found between thermally-determined values of SAR in the phantom and values of SAR calculated using the measured currents with the known dielectric properties of the phantom. However, usable current readings were only obtained after a full day of adjusting the shielding and

grounding of cables connecting the non-ferrous current probes to an oscilloscope used for readout.

Hyperthermia, the therapeutic use of heat, has shown promise as an adjuvant to radiotherapy in the treatment of cancer (Charny *et al.*, 1987; Gottlieb *et al.*, 1990). However, when RF heating is used for hyperthermia, there are often aberrant effects in which excessive heating in unintended regions of the body may pose a danger to the patient (Hagmann *et al.*, 1986). The non-ferrous current probes were tested with human subjects during hyperthermia treatments at the University of Utah Medical Center. However, a fiber-optic data link was not available for these tests so it was necessary to use coaxial cables to connect the probes to an oscilloscope located outside the screen room that was used for patient treatment. In this application it was not possible to eliminate cable pickup, or to separate the cable pickup from the signal generated by the current probe, so reliable current readings could not be obtained in the clinical tests.

Cable pickup is a common problem with ferrous current probes as well as other devices when they are used in an RF field, but the errors caused by cable pickup are more serious with the non-ferrous current probes. This is because, due to the relatively low sensitivity of the non-

ferrous current probes, the output signal is quite weak (often only a few millivolts) and so that the signal may have a smaller magnitude than that of the error signal that is added due to cable pickup. Various techniques that are commonly used to reduce cable pickup (Morrison, 1986) were tried, such as the changing the orientation of the cable, using double-shielded cable, ferrite-loaded cable, absorber-covered cable, shielded balanced line, or cable continuously connected to a ground plane. However, each of these techniques been found to provide inadequate protection when the non-ferrous probes are used in RF fields.

1.6 Statement of the Problem to be Solved in this Thesis.

The problem to be solved in this thesis is that of developing a means for reducing or eliminating the effects of cable pickup so that non-ferrous current probes may be used as dosimeters in order to determine the current induced in humans exposed to electromagnetic fields. The two approaches that were used in order to accomplish this task are as follows:

A. Development of a preamplifier to be attached to a non-ferrous current probe in order to provide sufficient amplification so that the amplified signal exceeds the errors added to the signal due to cable

pickup. The input signal to this circuit could be as low as -53 dBm, and should be amplified 20 dB, to deliver an output voltage in the mV range. A broadband balun transformer is needed followed by an small-signal amplifier.

B. Development of a self-contained device that amplifies and detects the output from a non-ferrous current probe and provides an indication of the current with a readout device such as a digital meter. Again, a broadband balun transformer is used, but in this case followed by two stages of small-signal amplification, and a medium power amplifier as the output stage. This output stage, which requires linearity at levels around $+10$ dBm, is finally designed using the PHILIPS RF transistor BFQ-34T. This output signal is detected using a SCHOTKY diode, and a DC circuitry, consisting of a DC amplifier, peak detector and a sample and hold, conditions the DC signal for the LCD read-out.

The details for the solution of these tasks are described in the following chapters of this thesis.

2. Development of a Preamplifier for Non-Ferrous Probes.

A preamplifier was developed that may be attached directly to a non-ferrous current probe in order to provide sufficient amplification so that the amplified signal exceeds the errors added to the signal due to cable pickup. The non-ferrous current probe used with this preamplifier has a transfer impedance of $1.90 \text{ m}\Omega/\text{MHz} \pm 12\%$ over a frequency range of 50 to 200 MHz. That is, at a frequency of 50 MHz, a current of 5 mA will cause the probe to have an output of approximately 500 μV across a 50Ω load, corresponding to a power of -53 dBm. The design objective for the preamplifier was to bring this power to -33 dBm, or 5 mV across 50Ω , so a gain of 20 dB is required. The requirements for the preamplifier are 1) an overall power gain $> 20 \text{ dB}$, 2) a usable bandwidth of 50 to 200 MHz, 3) compatibility with the balanced 50Ω output of the current probe, and 4) low power consumption.

The schematic for the preamplifier is shown in Fig. 1. A balun 1:1 transmission line transformer is used to couple the non-ferrous current probe to the circuit. The balun is followed by one stage of MMIC (Microwave Monolithic Integrated Circuit) amplification using a MAR-6

Fig. 1

PREAMPLIFIER USED IN BROOKS AFB ON AUGUST 1994

(Mini-Circuits, Brooklyn, New York). The balun, which had the same design as the balun used later in the self-contained current meter, is described in section 3.2 of this thesis.

The characteristics of the MAR-6 MMIC are 1) power gain of 20 dB, 2) output compression point of 0 dBm, 3) bandwidth of DC to 2 GHz, and 4) a noise figure of 3 dB (at 500 MHz and 25 °C). The internal structure of the MAR series of MMIC amplifiers are Darlington-connected transistor pairs, each with resistive feedback and resistive biasing. The internal resistive networks prematch both the input and output to 50 Ω , so no separate matching sections are needed. It was only necessary to add bias circuitry, input and output DC-blocking capacitors, and embedded in microstriplines as close as 50 Ω as possible. The circuit is powered with a 9 Volt battery, was shielded in a box and attached to the non-ferrous current probe.

The measured sensitivity of the non-ferrous current probe with the preamplifier are shown in Fig. 2. At frequencies of 60 MHz and higher the effective transfer impedance of the probe/preamplifier combination exceeds 1 Ω , which is typical of ferrous current probes, and consistent

Fig. 2 Transfer impedance for the preamplifier of the non-ferrous probe, used in Brooks AFB, August 1994

with the observation that the cable pickup could be neglected (as it is with ferrous current probes) during field tests.

The probe/preamplifier combination was used successfully in tests with human subjects exposed to RF fields in an anechoic chamber at Brooks Air Force Base, Texas (Hagmann, 1994). Ferrous current probes, as well as other non-ferrous current probes and stand-on current probes, were compared in these measurements. A length of approximately 8 meters of coaxial cable was required to connect each ferrous and non-ferrous current probe, as well as electric and magnetic field sensors, to RF feedthroughs mounted on the wall of the chamber. Thus, readings could be made using an external oscilloscope and other devices. Measurements were made at frequencies of 60, 100, 150 and 200 MHz, both with and without a ground plane. The probe/preamplifier/cable/oscilloscope combination typically gave a signal with no human subject (a signal caused by cable pickup) that was approximately 5 to 10 % of that when currents in the human body were measured. Thus, the amplification was great enough that the signal could be measured. In general, there was fairly good agreement between measurements with the ferrous and non-ferrous current probes. For example, current

readings taken on the ankle with the Eaton model 94606-2 ferrous current probe described in section 1.4 of this thesis were generally within 20 % of the values obtained using the probe/preamplifier/cable/oscilloscope combination.

As was previously mentioned in section 1.2 of this thesis, measurements were also made using the stand-on current probes produced by both Holaday Industries Inc. and Loral Microwave-Narda with human subjects exposed to RF fields in an anechoic chamber (Hagmann, 1994). These studies showed that both probes often give erroneous current readings when there is no human subject, and are unreliable unless the human subject is standing on a large metal ground plane, and the frequency is 60 MHz or less. For example, at 100 MHz the Holaday probe gave readings that were typically 4 to 5 times those with the ferrous and non-ferrous current probes.

3. Development of a Self-Contained Current Meter.

3.1 Introduction

A self-contained instrument that amplifies and detects the output from a current probe, and provides an indication of the current with a readout device (e.g. a digital meter) would have several advantages relative to the probe/preamplifier combination. With the latter system, it is necessary to use a cable to connect the probe/preamplifier combination to an oscilloscope or other readout device, and the cable causes errors by perturbing the RF fields in the region of measurement. Either the readout device must be modified so that it is shielded (Morrison, 1986), or else the cable must be long enough such that the readout device may be located where the fields are weak and will not interfere with the readout. There are many applications in which there is no null field region in which the readout apparatus may be located. The self-contained device has the additional advantage in that it would be immune to cable pickup since there would be no cable.

Blackwell was the first to describe a self-contained current meter that may be worn on the human body to measure the RF current induced in a mobile human subject (Blackwell, 1990). His circuit is based on a ferrous current probe that is large enough to be placed around the ankle, and can read currents in the range of 8 to 100 mA at frequencies from

100 kHz to 80 MHz. The output from the ferrous current probe is fed to a current transformer, a RF peak detector using a Schottky diode, an amplifier which uses a method described earlier (Greibenkemper, 1987) to compensate for nonlinearity of the detector, zero-setting circuitry, and finally a digital panel meter.

The first self-contained current meter that was constructed for this thesis was designed for measurements at frequencies in the AM broadcast band. While we prefer to use non-ferrous current probes, a ferrous current probe was required for this application in order to have adequate sensitivity at the lower frequencies (0.5 to 1.6 MHz). The current meter was based on an Eaton model 94606-2 ferrous current probe having a measured transfer impedance of $1.0 \Omega \pm 20 \%$ in the frequency range from 300 kHz to 70 MHz. A battery-powered amplifier and detection circuit with an analog panel meter were housed in a small shielded box connected directly to the ferrous current probe.

The amplifier for this self-contained current meter was designed to provide much greater sensitivity than the device that was described by Blackwell (Blackwell, 1990) because high sensitivity is often necessary to

determine the actual values of the currents when there are complaints due to human exposure. The meter has three ranges, the most sensitive requiring 160 μA for full scale deflection of the analog panel meter, which permits detecting a current of 16 μA at 1/10 of full scale. This instrument provides 500 times the sensitivity of the circuit by Blackwell (8 mA).

The schematics for the self-contained amplifier and detection circuit is shown in Fig. 3. The circuit consists of two stage of amplification, the first amplifier consists of a 2N3563 transistor biased in common emitter configuration. The voltage gain of this stage is

$$A_V \approx -1 \cdot \left(\frac{g_m}{1 + g_m R_E} \right) \cdot R_O \approx \frac{R_O}{R_E} \approx -2$$

The second stage was designed around a Darlington transistor to match the relatively high output impedance of Q2 and the low impedance of the detector. The voltage gain of this stage is 16, for an overall voltage gain of 336.

Because the frequency of operation is low, a germanium diode was selected as the detector. Finally, a DC amplifier with a gain of 5 isolates the detector load from the meter. Measurements of the sensitivity of the circuit at 1.0 MHz are shown in Figs.4 and 5. As shown in the schematic, there are three scales, with different sensitivities. Scale 1 is the most

sensitive, and Scale 3 is the least sensitive. The calibrations in Figs. 4 and 5 are for scales 2 and 3, respectively. Scale 1 was not fully calibrated because its extremely high sensitivity was not needed for this application.

Fig. 3 Schematic for the low-frequency self-contained current meter

Fig. 4 Scale 2 response of the Current Probe used
in the AM Radio Station

Fig. 5 Scale 3 response of the Current Probe used
in the AM Radio Station

The battery-powered amplifier and detection circuit shown in Fig. 3 differs from that used by Blackwell (Blackwell, 1990) in that he used the detector diode in its non-linear region (square law), and corrected for the nonlinearity in the DC path, by using a second diode in the feedback. The circuit which was constructed for this thesis avoids this complication because, due to amplification before the diode, the diode is operated in the linear region, so that compensation is not required.

Dr. David Conover and Dr. Gene Moss of NIOSH requested that this low frequency meter be used to measure the currents induced in personnel at an AM radio station where the employees had complained about possible hazards. The measurements were made and a description of the results is contained in Appendix III of this thesis.

In accordance with the statement of the problem to be solved in this thesis, it was necessary to develop a self-contained device, similar to that made by Blackwell (Blackwell, 1990), but based on a non-ferrous current probe. From section 1.4 of this thesis, the improvements to be expected from using a non-ferrous probe are that 1) the probe size could be increased for measurements at the thigh, waist, etc., or else

frequencies as high as 450 MHz could be used in measurements at the ankle, 2) perturbations due to insertion impedance would be decreased, 3) the accuracy would be increased, 4) the cost would be decreased, and 5) the weight would be decreased.

The challenge in designing a self-contained current meter based on a non-ferrous current probe is considerably greater than that when a ferrous probe is used. This difference is due to the following factors; 1) the non-ferrous current probe has much lower sensitivity so that much greater amplification is required, 2) the coil winding acts as a balanced source so that a balun is required to couple it to unbalanced amplifier and detection circuits, and 3) the coil has a high series resistance (typically 30 k Ω) so that the amplifier must be matched to a high source impedance. The following sections of this thesis address each of these factors and describe the methods used to produce a final working instrument that is suitable for use in dosimetry.

3.2 Balun: Balanced-To-Unbalanced Transformer.

Transmission lines may be characterized as being either balanced (symmetrical, such as two parallel-conductor line), or unbalanced (coaxial having both inside and outside parts) (Morrison, 1986). Whenever balanced and unbalanced devices are connected together it is necessary to use a BALUN (Balanced to Unbalanced transformer) in order to prevent currents from flowing on the outside of the unbalanced device (Balanis, 1982; Kraus, 1988; Shuhao, 1987). For example, when coaxial cable is connected to a dipole antenna the pattern of the antenna is changed, shocks may be received when touching the outside of the coaxial cable during transmission, and excessive interference may occur during reception. The coil of a non-ferrous current probe acts as a balanced source, so it is necessary to use a BALUN to couple this coil to the unbalanced amplifier and detection circuits.

A variety of different designs have been used for baluns (Shuhao, 1987). Most balun designs require that a section of transmission line be a fixed fraction of a wavelength in length, so they are narrowband devices

(Woodward, 1953). The bandwidth can be considerably increased by employing ferrite cores (Motorola, 1991; Ruthroff, 1959; Weeks, 1968).

It was necessary to choose a balun transformer with a ferrite core in the self-contained current probe because of the requirement for a large bandwidth. Furthermore, a balun with transmission line for the winding in place of separate primary and secondary windings (as in a conventional transformer), was used in order to obtain greater bandwidth by reducing the stray inductance and winding capacitance. Twisted pair was used in place of a coax for the winding because this permits a smaller size, and is appropriate because the signal is small in amplitude and the frequency is generally below 500 MHz (Hilbers, 1970). A transformation ratio (output impedance to input impedance) of unity was satisfactory so it was possible to simplify the construction by using a single winding. This winding was made of a transmission line formed by twisting a pair of CuEm (copper enameled) wires of 10 mils diameter, in order to have a characteristic impedance very close to $50\ \Omega$ (Hilbers, 1970). The core used for the balun is ferrite material number 61 (FAIR-RITE Products Corp., Wallkill, New York) which has a relative permeability (μ_r) of 125.

The inductance of the balun should be large because it determines the amount of reflection at the low frequency end of the band. The inductance may be determined by using the following equation:

$$L = \mu_o \cdot \mu_r \cdot n^2 \cdot \frac{A}{l} \quad (1)$$

where L is the inductance in Henrys, $\mu_o = 4\pi 10^{-7}$ Henrys/meter, n is the number of turns, A is the average ferrite cross section in square meters, and l is the average length of the lines of force in meters.

If the inductance is too large, then the performance at the high end of the band is degraded. A good practical value for the inductance is given by

$$L = \frac{4 \cdot R}{\omega_{\min}} \quad (2)$$

where R is the midband input resistance in Ω (for our case = 50 Ω), and $\omega_{\min}$ is the minimum frequency in radians per second. Applying this empirical equation (Hilbers, 1970), the recommended value for the inductance is 0.636 μH . Thus, solving for n in Eq. (1);

$$n = \sqrt{\frac{L \cdot l}{\mu_o \cdot \mu_r \cdot A}} \quad (3)$$

The ferrite bead of 61-type material used in this balun, has a cross section of $42 \cdot 10^{-7}$ square meters, and a mean perimeter of 0.00718 meters.

Thus, from Eq. (3),

$$\frac{1}{A} = 1712.37 \text{m}^{-1} \quad (4)$$

Substituting into Eq. (3), we obtain the following value for n:

$$n = \sqrt{\frac{0.636 \cdot 10^{-6} \cdot 1712.37}{4 \cdot \pi \cdot 10^{-7} \cdot 125}} = 2.63 \quad (5)$$

so that 3 turns were used for the winding.

The balun was tested in two different configurations. In the first, which is termed the "transmission line" configuration, the two ends of the twisted pair are used for the primary and secondary connections, respectively. In the second configuration, which is termed "conventional winding", the two ends of one wire are used for the primary connection and the two ends of the other wire are used for the secondary connection. In the extreme low frequency limit (DC), the first configuration provides zero insertion loss and the second has infinite insertion loss. The insertion loss measured for the balun in the two configurations is shown in Figs. 6 and 7. The transmission line configuration was found to provide much less insertion loss, so this configuration was used in the project.

Fig. 6 Insertion loss of a ferrite balun, using the twisted pair as transmission line.

Fig. 7 Insertion loss of a ferrite balun, using the twisted pair as a conventional winding.

3.3 Detection Circuit.

Detectors are essentially low sensitivity receivers which function by rectifying an RF signal by means of a non-linear resistive element - a diode. Generally, detectors can be classified into two distinct types: the small-signal type, also known as square-law detectors; and the large-signal type, also known as linear or peak detectors (Hewlett Packard, 1981).

The small-signal operation of a detector is dependent on the slope and curvature of the IV (current voltage) characteristics of the diode in the neighborhood of the bias point. The output of the detector is proportional to the power input to the diode, that is, the output voltage (or current) is proportional to the square of the input voltage (or current), hence the term "square law". The large signal operation of a detector is dependent on the slope of the IV characteristic in the linear portion, so the diode functions essentially as a switch. In large signal detection, the diode conducts over a portion of the input cycle and the output current of the diode follows the peaks of the input signal waveform with a linear relationship between the output current and the input voltage.

The square law dynamic range of a detector is defined as the difference between the power input for a 1 dB deviation from the ideal square law response (compression point) and the power input corresponding to the tangential signal sensitivity (TSS) (Hewlett Packard, 1975). The tangential signal sensitivity (TSS) is a measurement of the ability of the diode to distinguish a small signal from noise. The name relates to a type of radar display with the bottom of the signal pulse tangent to the top of the noise level.

A schottky barrier diode has a metal-semiconductor barrier that is formed by depositing a metal layer on a semiconductor. The Schottky diode is more rugged than point contact diodes, and it has better response than p-n junction diodes at high frequencies because the response of p-n junction diodes is limited by minority carriers that are not present in the Schottky diode. Normal operating conditions for the Schottky detector call for a large load resistance (100 K Ω) and a small bias current (20 μ A) (Hewlett Packard, 1982). These normal conditions assure the minimum value of TSS input level (maximum sensitivity), but not the maximum value of compression level.

The compression level can be raised by reducing the value of the load resistance, R_L (Hewlett Packard, 1975). However, the sensitivity degrades by the factor:

$$\frac{R_L}{R_L + R_v} \quad (6)$$

Where R_v is the diode's resistance. This degradation in TSS exceeds the improvement in compression, so there is no improvement in square law dynamic range. Another technique for raising the compression level is to increase the bias current. This also degrades the sensitivity, but the improvement in compression exceeds this degradation so square law dynamic range is increased.

Over a wide range of input power level, P , the output voltage, V , from a diode follows the equation

$$V = K \cdot (\sqrt{P})^\alpha \quad (7)$$

where K and α are constants (Hewlett Packard, 1981). At low levels of power, from -60 dBm and up to -20 dBm, the value of α is 2. This is the square law region. When a DC bias current is used (usually a few

microamperes), then the diode impedance is independent of power level.

At higher power levels, the diode impedance changes with power. At these levels, the value of α drops from the original 2, passing through 1 (linear rectification), and going as low as 0.8.

Conventional Schottky detector diodes are tested and specified for use with 20 μ A of DC bias. The bias current reduces the junction resistance so that most of the detected voltage appears across the load resistance. In some applications the diode is used to monitor power rather than to detect a low level signal. In this case, the signal level may be high enough to reduce the junction resistance sufficiently without the use of a DC bias.

A detector diode may be considered as a voltage source in series with the diode resistance. The output voltage is taken from a load resistance in series with the diode. This circuit is a voltage divider. The detected voltage is divided between the diode and the load. The useful output voltage is given by

$$V_o = \frac{V_{oc} \cdot R_L}{R_L + R_D} = \frac{V_{oc}}{1 + \frac{R_D}{R_L}} \quad (8)$$

For best performance the ratio of diode resistance to load resistance should be small (Hewlett Packard, 1982). However, the use of load resistances greater than 100K Ω would significantly increase the response time of the overall circuit.

The junction resistance may be calculated from the diode equation (Hewlett Packard, 1982):

$$I = I_s \cdot (e^{\frac{V}{0.027}} - 1) \quad (9)$$

where V is the voltage across the junction and I_s a constant. The inverse derivative (dV/dI) is the junction resistance, which is given by

$$R_j = \frac{0.027}{1 + I_s} \quad (10)$$

The constant I_s , called saturation current, is about 0.7×10^{-9} amperes for the HP 5082-2755 detector diode. With 20 μ A of DC bias the junction resistance is 1,350 Ω . However, at zero bias the resistance is about 40 M Ω .

The following conclusions regarding the use of biased and unbiased detectors are supported by the data which have been taken with Schottky diodes (Hewlett Packard, 1975):

1.- A biased Schottky diode with an input power between -60 to -20 dBm has a square law response given by

$$V_{\text{out}} = K_1 \cdot (\sqrt{P_{\text{in}}})^2 = K_2 \cdot (V_{\text{in}})^2 \quad (11)$$

With a load of 100 K Ω , the output voltage ranges from fractions of a millivolt to as high as 100 mV. In order to use this wide dynamic range, a high DC gain needed (greater than 200), which requires a low-drift DC amplifier and a very accurate bias circuit for the Schottky diode. Since we would like to have an output proportional to the input RF current, then a "square rooter" circuit (for which the output voltage is proportional to the square root of the input voltage) must be included in the DC section.

2.- An unbiased Schottky diode has poor sensitivity when the input power is less than -15 dBm because only the small fraction of the input voltage that exceeds the diode threshold voltage will appear at the output. However, with an input power from -15 to +10 dBm, the diode output is given by

$$V_{\text{out}} = K_3 \cdot \sqrt{P_{\text{in}}} = K_4 \cdot V_{\text{in}} \quad (12)$$

Thus, a "square rooter" circuit is not required. Furthermore, for this range of power, the output voltage across a 100 K Ω load is in the range from tenths of a Volt, to several Volts.

In conclusion, after working with circuits using biased and unbiased Schottky diodes, it was decided that greater accuracy and stability may be obtained by amplifying the signal and then detecting it with an unbiased Schottky diode in the linear region. The challenge in this approach comes from the fact that considerable amplification is required to bring a small input signal (typically -40 dBm) to levels as high as 0 dBm. Amplifiers with such a high gain tend to be unstable due to oscillations caused by stray feedback. The usual solution is to use tight RF shielding between sections, or else the more elaborate method of frequency conversion.

3.4 RF Preamplifier.

Three separate devices which were built in the research for this thesis have required the development of RF amplifiers. These include the preamplifier for use with non-ferrous probes described in Section 2 , the self-contained low-frequency current meter described in Section 3.1, and the self-contained high-frequency current meter described in the present section of this thesis. For this last circuit, an unusually high gain is required because it is necessary to use an unbiased Schottky diode in its linear region.

3.4.1 Broadband amplifier using the MWA-120

The balanced input from the high-frequency non-ferrous current probe, described in Section 3.7, was transformed using a balun similar as the one described in Section 3.2. A variable attenuator follows, to control the power level that reach the amplifier chain, and keep them in their linear region. This attenuator is made with a double-balanced mixer (SBL-1, Mini-Circuits), controlling the isolation between the RF and LO (local oscillator) ports. This control is realized by adjusting the current flowing

through the IF port. At zero current, the mixer shows an isolation of 40 dB (almost an open circuit), at 0.5 mA the isolation drops to 20 dB, and at current above 10 mA, the isolation goes down to only 2 dB (almost a short circuit). After this variable attenuator, two stages of amplifications were needed using two cascaded MARS-6 MMIC (Section 2), followed by a hybrid amplifier with a greater output capability. This amplifier is the MWA-120, with a frequency range up to 700 MHz, Power Gain of 14 dB, and a 1 dB compression point of +8.2 dBm. A circuit was built and tested (Fig. 8), with the results shown in Figs. 9, 10 and 11. The circuit performance fell short in dynamic range. To increase it, a discrete broadband amplifier was finally designed to replace the output hybrid amplifier.

Fig. 9 Scale 1 response at 465 MHz for the current probe using the MWA-120.

Fig. 10 Scale 2 response at 465 MHz for the current probe using the MWA-120

Fig. 11 Scale 3 response at 465 MHz for the current probe using the MWA-120

3.4.2 Broadband amplifier using the BFQ-34T

To increase the output power of the final amplifier, the BFQ-34T transistor was used in a broadband configuration.

3.4.2.1 Broadband Matching

The design of a broadband RF amplifier is a problem of the power-gain roll-off respect to frequency. In general, there are four types of broadband amplifier design techniques:

a) **Compensated matching techniques.** The compensated matching technique allows a small mismatch of the input and output matching networks to compensate for the variations of the forward transmission parameter of S_{21} with frequency. This method can be achieved by using the Smith Chart. By means of a CAD program, the design process should be much easier.

b) **Network synthesis techniques.** The intrinsic input and output impedances of a transistor are measurable and available. To match its

equivalent R-L-C circuits, a well-known network synthesis method, such as Butterworth or Chebyshev response, can be used analytically.

The disadvantage of this method is to use lumped elements for matching purposes.

c) Feedback network techniques. Feedback network is often used in a broad-band amplifier to provide a flat gain response and to reduce the input and output VSWR.

d) Balanced-amplifier techniques. A balanced amplifier is also commonly used to obtain a broadband amplifier with a flat gain and good input and output VSWR.

In this project, the first method for broadmatching will be used and explained.

3.4.2.2 Broadband matching using compensated techniques.

- Selection of the active device.

The amplifier design specifications are:

Gain = 12 dB $\pm$ 0.5 dB

$P_{1dB} = +15\text{dBm}$

Bandwidth= 100 MHz to 500 MHz

The transistor selected is the BFQ-34T, from Philips. It is a NPN transistor in a plastic SOT-37, intended for wideband applications. The device features high output power capabilities.

Its quick reference data is as follows:

Collector-base voltage (open emitter)	VCBO max	25V
Collector-emitter voltage (open base)	VCEO max	18 V
Collector current (d.c.)	Ic max	150 mA
Total power dissipation (up to Tamb=45°C)	Ptot max	1 Watt
D.C. current gain	hFE min	25

Ic=100mA, Vce=10V

Transition frequency at f=500 MHz	fT typ.	3.7 GHz
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Ic=100mA; Vce=10V

Maximum Power Gain at f=300 MHz	Gum typ.	19.5 dB
---------------------------------	----------	---------

Ic=100mA; Vce=10V

Output power at 1 dB gain compression	PL1 typ.	+24 dBm
---------------------------------------	----------	---------

Ic=100mA; Vce=10V; f=300MHz

Third Order Intercept Point	ITO	+43 dBm
-----------------------------	-----	---------

Ic=100mA; Vce=10V; f=300MHz

The manufacturer offers a wide range when describing this transistors with its S-Parameter. In frequency goes from 40 MHz to 2.0 GHz, and is subdivided according to the bias condition, which range from 10 mA to as high as 100mA, at $V_{ce}=10V$.

Because our application demands portability, our detector will be battery-powered, and we will always fall short in our concern for consumption.

With this in mind, we targetted the specified compression point with the minimum biased offered, in this case 10 mA at $V_{ce}=10V$.

The S-Parameters (TOUCHSTONE format) for the BFQ-34T at 10 mA are offered below:

```
! BFQ34T10.S2P
```

```
! BFQ-34T
```

```
! VCE=10V; IC=10mA
```

```
# GHz S MA R 50
```

```
! S-PARAMETER DATA
```

```
0.04  0.79  -45.4  20.9  153  0.03  68.1  0.89  -21.2
```

```
0.1    0.66  -98.3  14.2  122.8  0.04  49.5  0.64  -41.2
```

```
0.2    0.57  -137.7  8.3   103   0.06  46.5  0.43  -49.9
```

0.5	0.58	-178.3	3.6	76.2	0.08	57.1	0.33	-64.5
0.8	0.57	162	2.3	61.2	0.12	67.7	0.36	-80.5
1.0	0.59	150.0	1.9	51.5	0.15	70.1	0.38	-90.4
1.2	0.63	138.5	1.5	45.4	0.18	72.9	0.38	-100.6
1.5	0.61	127.8	1.3	35.7	0.25	72.3	0.43	-115.6
2.0	0.66	101.8	1.0	25.8	0.36	68.1	0.48	-143.1

Where the first column stands for frequency in GHz, the 2nd and 3rd for the module and angle of S_{11} , fourth and fifth for S_{21} , sixth and seventh for S_{12} (neglectable in our case), and finally the last two columns represent the module and angle of S_{22} .

3.4.2.3 Stability considerations.

The stability of an amplifier is a very important consideration in a RF circuit design. The stability can be determined by the S-parameters, the source and load impedances.

Oscillations are possible in a two-port network if either the input or output port, or both, have negative resistance or when either $|\Gamma_{in}| > 1$ or $|\Gamma_{out}| > 1$.

There are two traditional expressions used when speaking of stability; conditional and unconditional stability:

_Conditional stability.

A network is conditionally stable if the real part of the input impedance Z_{in} and the output impedance Z_{out} is greater than zero for some positive real source and load impedances at a specific frequency.

_ Unconditional stability.

A network is unconditionally stable if the real part of the input impedance Z_{in} and the output impedance Z_{out} is greater than zero for all positive real source and load impedances at a specific frequency.

It is important to note that these two conditions apply only at one specific frequency. The conditions we will now discuss will have to be investigated at many frequencies to ensure broadband stability.

By definition the reflection coefficient at any point "i" in a transmission line with a characteristic impedance equal to Z_0 is defined as

$$\Gamma_i = \frac{V_{\text{reflected}}}{V_{\text{incident}}} = \frac{Z_i - Z_0}{Z_i + Z_0} \quad (13)$$

Then, by application of this definition, having positive real source and load impedances implies:

$$\Gamma_S \text{ and } \Gamma_L \leq 1 \quad (14)$$

When we want to conjugately match the network to the load and source to achieve maximum transducer gain, the Rollett stability factor or simply the K factor will give us an indication of unconditional stability if it happens to be greater than 1.

$$K = \frac{1 + |S_{11} \cdot S_{22} - S_{12} \cdot S_{21}|^2 - |S_{11}|^2 - |S_{22}|^2}{2 \cdot |S_{12}| \cdot |S_{21}|} > 1 \quad (15)$$

In our design, we will evaluate at three different points, these are: 100MHz, 200 MHz and 500 MHz. Using the S-parameters given by the manufacturer, and evaluating we obtain:

At 100 MHz: K=0.296

At 200 MHz: K=0.576

At 500 MHz: K=0.987

These results tell us that the device is potentially unstable at these three frequencies (and probably at frequencies in between), and will most likely oscillate with certain combinations of source and load impedance. We then must be extremely careful in choosing source and load impedances for the transistor. It does not mean that the transistor can not be used for our application, it merely indicates that the transistor will be more difficult to use.

When a transistor is potentially unstable, the source and load reflection coefficient will have to be confined to safe or stable regions. To determine these regions, we will calculate the constant gain circle for the source and load reflection coefficients, needed to comply with the gain specified. In this sense, we will resume the stability analysis once we have determined the region of our source and load.

3.4.2.4 Constant-Gain Circles.

Let us now find the input and output matching network such that provide a constant overall gain between 100 MHz and 500 MHz. The $|S_{12}|$ term of our transistor in this frequency range is not greater than 0.08, then a significant simplification is made if we neglect this term and assume the transistor to be unilateral. It is obvious that for $\Gamma_s = S_{11}^*$ and $\Gamma_L = S_{22}^*$, the power gain G_s or G_L is equal to a maximum, respectively. It is also clear that for $|\Gamma_s| = 0$ or $|\Gamma_L| = 1$, the power gain G_s or G_L has a value of zero. For any arbitrary value of G_s or G_L , between these extremes of zero and G_{smax} or G_{Lmax} , solutions for Γ_s or Γ_L lie on a circle.

For $0 < G_s < G_{smax} \text{ (dB)}$

$$g_s = \frac{1 - |\Gamma_s|^2}{|1 - S_{11} \cdot \Gamma_s|^2} \quad (16)$$

It is convenient to plot these circles on a Smith Chart. The circles have their centers located somewhere on the vector drawn from the center of the Smith Chart to the point S_{11}^* or S_{22}^* . The distance from the center of the Smith chart to the center of the constant-gain circle is given by

$$d_s = \frac{g_{ns} \cdot |S_{11}|^2}{1 - |S_{11}|^2 \cdot (1 - g_{ns})} \quad (17)$$

The radius of the constant-gain circle is expressed by

$$r_s = \frac{\sqrt{1 - g_{ns}} \cdot (1 - |S_{11}|^2)}{1 - |S_{11}|^2 \cdot (1 - g_{ns})} \quad (18)$$

where g_n is the normalized value for the gain circle g_s or g_L , respectively.

That is,

$$g_{ns} = \frac{g_s}{g_{s \max}} = g_s \cdot (1 - |S_{11}|^2) \quad (19)$$

where $0 < g_{ns} < 1$.

These circles are interpreted as follows: any given matching network, with its reflection coefficient located right on a constant-gain circle, will contribute a gain equal to the gain of the circle. Since the expression for the output gain term has the same form as that for G_s , a similar set of constant gain circles can be drawn for this term. In our case we obtain;

For 100 MHz:

$$G_{U \max} = \frac{1}{1 - |S_{11}|^2} \cdot |S_{21}|^2 \cdot \frac{1}{1 - |S_{22}|^2} \quad (20)$$

substituting the values:

$$G_{\max} = 27.83 \text{ dB}$$

We also know that the intrinsic gain of the transistor equals:

$$G_{\text{device}} = 10 \cdot \log |S_{21}|^2$$

Substituting,

$$G_{\text{device}} = 23.05 \text{ dB}$$

In essence, an RF amplifier consists of an input matching section, an active device (feed-back used in applications below 1 GHz) and an output matching network. The term G_{device} tells us the gain of the active device when its input and output ports (base and collector in our case) are directly connected to 50 Ω systems, without any matching at all. However,

by implementing input and output matching networks, we can control the overall gain of the amplifier, to fit our needs. In other words, the gain can be made less, equal or even greater than the intrinsic gain of the device.

Then we have,

$$G_{\text{max}} - G_{\text{device}} = 27.83 - 23.05 = 4.78 \text{ dB}$$

If instead of dropping the transistor right on 50Ω terminations, we match its input and output in such a way that the base sees S_{11}^* and the collector sees S_{22}^* , then the overall gain would have increased in 4.78 dB above the intrinsic gain of 23.05. However, we are not interested in such a high gain, our goal is to achieve a flat response of 12 dB, then the input and output network will have to act as attenuator. As the frequency increases, their task will have to change, they will not longer be attenuators, they will have to contribute gain to the amplifier, again by doing a better matching at the input and output ports.

As our goal is 12 dB:

$$12\text{dB} - G_{\text{device}} = 12 - 23.05 = -11.05 \text{ dB}$$

The matching network will have to decrease by 11 dB the intrinsic gain of the device at this frequency, this attenuation has to be shared by the input and output circuitry. I selected an attenuation of 6 dB (gain = -6 dB) at the input, and a gain = -5 dB at the output, these are the gain values that we will have to use to get the constant-gain circles.

For the input constant-gain circle at 100 MHz:

$$g_{SdB} = -6 \text{ dB}$$

$$g_s = 10^{\left(\frac{g_{SdB}}{10}\right)} = 0.251$$

We also know that,

$$g_{ns} = \frac{g_s}{g_{s \max}} = g_s \cdot (1 - |S_{11}|^2) = 0.141 \quad (21)$$

and substituting in the equations for the constant-gain circles we obtain:

$$d_s = \frac{g_{ns} \cdot |S_{11}|^2}{1 - |S_{11}|^2 \cdot (1 - g_{ns})} = 2.71 \quad (22)$$

and

$$r_s = \frac{\sqrt{1 - g_{ns}} \cdot (1 - |S_{11}|^2)}{1 - |S_{11}|^2 \cdot (1 - g_{ns})} = 2.24 \quad (23)$$

Where d_s is the vector that goes from the center of the Smith Chart to the center of the circle, with an angle equal to the conjugate of S_{11} , and r_s is the radius of the circle.

Calculating now the constant-gain circle for the output network:

As we said, the gain of this section should be negative, to indicate attenuation:

$$g_{LdB} = -5 \text{ dB}$$

Converting from gain in dB to unitary gain:

$$g_L = 10^{\left(\frac{g_{LdB}}{10}\right)} = 0.316$$

And normalizing this gain respect its maximum possible gain:

$$g_{nL} = \frac{g_L}{g_{L \max}} = g_L \cdot (1 - |S_{22}|^2) = 0.186 \quad (24)$$

Applying the formulas for the constant-gain circle;

$$d_L = \frac{g_{nL} \cdot |S_{22}|^2}{1 - |S_{22}|^2 \cdot (1 - g_{nL})} = 0.179 \quad (25)$$

$$r_L = \frac{\sqrt{1 - g_{nL}} \cdot (1 - |S_{22}|^2)}{1 - |S_{22}|^2 \cdot (1 - g_{nL})} = 0.798 \quad (26)$$

At 200 MHz we have:

And its intrinsic gain at this frequency is:

$$G_{\text{device}} = 10 \cdot \log|S_{21}|^2$$

$$G_{\text{device}} = 10 \cdot \log(3.6)^2$$

$$G_{\text{device}} = 11.12 \text{ dB}$$

Let us remember that we need 12 dB of gain, then;

$$12\text{dB} - G_{\text{device}} = 12 - 11.12 = 0.88 \text{ dB}$$

The gain of our matching network should be around 1 dB, being positive, instead of an attenuation we do need gain. This could be confusing, because the network is conformed of passive (ideally lossless) elements, and is demanded to "amplify". This "amplification" is actually achieved by improving the natural input or output reflection coefficients, offered by the manufacturer in the S-paramaters set of data. The input network was selected to have 1 dB of gain, and 0 dB for the output network,

$$g_{\text{SdB}} = 1\text{dB}$$

Referred to the unitary gain;

$$g_s = 10^{(\frac{g_{\text{SdB}}}{10})} = 1.259$$

We also know that,

$$g_{\text{ns}} = \frac{g_s}{g_{s \text{ max}}} = g_s \cdot (1 - |S_{11}|^2) = 0.835 \quad (33)$$

and substituting in the equations for the constant-gain circles we obtain:

$$d_s = \frac{g_{ns} \cdot |S_{11}|^2}{1 - |S_{11}|^2 \cdot (1 - g_{ns})} = 0.513 \quad (34)$$

and

$$r_s = \frac{\sqrt{1 - g_{ns}} \cdot (1 - |S_{11}|^2)}{1 - |S_{11}|^2 \cdot (1 - g_{ns})} = 0.285 \quad (35)$$

Referring to the constant-gain equations for the load, it should have a gain of zero dB,

$$g_{LdB} = 0 \text{ dB}$$

Converting from gain in dB to unitary gain:

$$g_L = 10^{\left(\frac{g_{LdB}}{10}\right)} = 1.0$$

And normalizing this gain respect its maximum possible gain:

$$g_{nL} = \frac{g_L}{g_{L \max}} = g_L \cdot (1 - |S_{22}|^2) = 0.891 \quad (36)$$

Applying the formulas for the constant-gain circle;

$$d_L = \frac{g_{nL} \cdot |S_{22}|^2}{1 - |S_{22}|^2 \cdot (1 - g_{nL})} = 0.298 \quad (37)$$

$$r_L = \frac{\sqrt{1 - g_{nL}} \cdot (1 - |S_{22}|^2)}{1 - |S_{22}|^2 \cdot (1 - g_{nL})} = 0.298 \quad (38)$$

At this point, we have already determined the regions in which our source and load reflection coefficients should be. At the beginning of this chapter, the K factor was determined, giving values below one for the whole frequency range. But will this K factor be something determinant, or on the other hand, might be giving a false indication of instability for the locus of reflection coefficients previously calculated? To answer to this question I will use an alternative K factor, which is presented in a technical paper by A.J. Slobodnik, Jr. and R.T. Webster, in Microwave Journal, February 1994.

3.4.2.5 Alternative Stability Factor for Amplifier Design.

This alternative K factor, K_A , provides a definitive indication of stability and permits the design engineer to concentrate stability circle investigations on those frequency ranges where further insight is needed.

Fig. 11.1 Constant-Gain and Stability circles.

Fig. 11.1 shows a Smith Chart and source plane stability circle as an example, including the source reflection coefficient with uncertainty radius δ_s and other parameters of interest. For simplicity purposes, it is assumed temporarily that the stable region is outside of the stability circle. The vector to the center point of the stability circle is given by

$$C_s = \frac{S_{22} \cdot \Delta^* - S_{11}^*}{|\Delta|^2 - |S_{11}|^2} \quad (39)$$

And the radius of the stability circle by

$$R_s = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{11}|^2} \right| \quad (40)$$

where

$$\Delta = S_{11} \cdot S_{22} - S_{12} \cdot S_{21} \quad (41)$$

For this circuit to be stable, the source reflection coefficient Γ_s , with any associated uncertainty δ_s , must lie outside the circle. From straightforward geometric considerations, it can be seen that this condition is met if the magnitude of the difference vector between C_s and Γ_s is greater than R_s plus δ_s .

This is,

$$C_s - \Gamma_s > R_s + \delta_s \quad (42)$$

With this in mind, the input alternative stability factor can be defined as

$$K_{AS} = \left(\frac{|C_s - \Gamma_s|}{R_s + \delta_s} \right)^{P_s} \quad (43)$$

The exponent P_s is introduced to account for the general case where either the outside or the inside of the stability circle corresponds to the stable region. P_s is determined by computing

$$\Gamma_{OUT} = S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{LP}}{1 - S_{11} \cdot \Gamma_{LP}} \quad (44)$$

where

Γ_{LP} = a test point chosen inside the stability circle.

If $|\Gamma_{OUT}| > 1$, then following definition of a stability circle, the outside of the circle is the stable region. $P_s = +1$ is then used. Otherwise, if $|\Gamma_{OUT}| < 1$, the inside of the circle is the stable region and $P_s = -1$ is used. The output alternative stability factor K_{AL} is determined in a similar manner. A definitive indication of stability is then provided as for $K_{AS} > 1$ and $K_{AL} > 1$, the network is stable under the given source and load conditions, and for $K_{AS} < 1$ or $K_{AL} < 1$ the network is unstable. These conditions correspond as much as possible to those of the traditional stability factor, preserving the intuitive analysis of a given design.

3.4.2.6 Application of the Alternative K factor in our design.

We already know the region in which our source and load impedance will move in order to achieve the gain targetted. The procedure that follows is divided in:

- 1.- Determine the source and load stability circles.
- 2.- Apply the alternative K factor, and see which frequency is conditionally stable. Determine whether is because of the input or output, or both.
- 3.- Plot in the Smith Chart the stability circles for those frequencies which are prone to oscillation (alternative K factor less than 1). In the same chart, draw the constant-gain circles and see the region in which they overlap. This zone is unstable, and we will try to stay out of it, when matching our device.

3.4.2.7 Determining the source stability circles.

At f=100 MHz:

The center of the source stability circle is defined by the vector $\vec{C}_s$:

$$\vec{C}_s = \frac{S_{22} \cdot \Delta^* - S_{11}^*}{|\Delta|^2 - |S_{11}|^2} = -1.413 + j2.304 \quad (45)$$

where

$$|\vec{C}_s| = 2.703$$

$$\vec{C}_s(\text{angle}) = 121.51^\circ$$

And the radius of the stability circle is given by:

$$R_s = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{11}|^2} \right| = 2.233 \quad (46)$$

At f=200 MHz:

The center of the source stability circle is located at:

$$\bar{C}_s = \frac{S_{22} \cdot \Delta^* - S_{11}^*}{|\Delta|^2 - |S_{11}|^2} = -2.257 + j1.588 \quad (47)$$

where

$$|\bar{C}_s| = 2.76$$

$$\bar{C}_s(\text{angle}) = 144.87^\circ$$

And the radius of the stability circle is given by:

$$R_s = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{11}|^2} \right| = 2.06 \quad (48)$$

At f=500 MHz:

The center of the source stability circle is located at:

$$\bar{C}_s = \frac{S_{22} \cdot \Delta^* - S_{11}^*}{|\Delta|^2 - |S_{11}|^2} = -1.881 + j0.138 \quad (49)$$

where

$$|\bar{C}_s| = 1.886$$

$$\vec{C}_s(\text{angle}) = 175.81^\circ$$

And the radius of the stability circle is given by:

$$R_s = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{11}|^2} \right| = 0.892 \quad (50)$$

3.4.2.8 Determining the load stability circles.

At f=100 MHz:

The center of the load stability circle is defined by the vector $\vec{C}_L$:

$$\vec{C}_L = \frac{S_{11} \cdot \Delta^* - S_{22}^*}{|\Delta|^2 - |S_{22}|^2} = 1.207 + j2.683 \quad (51)$$

where

$$|\vec{C}_L| = 2.942$$

$$\vec{C}_L(\text{angle}) = 65.77^\circ$$

And the radius of the stability circle is given by:

$$R_L = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{22}|^2} \right| = 2.487 \quad (52)$$

At f=200 MHz:

The center of the load stability circle at 200 MHz is located at:

$$\bar{C}_L = \frac{S_{11} \cdot \Delta^* - S_{22}^*}{|\Delta|^2 - |S_{22}|^2} = 2.66 + j4.85 \quad (53)$$

where

$$|\bar{C}_L| = 5.533$$

$$\bar{C}_L(\text{angle}) = 61.22^\circ$$

And the radius of the stability circle is given by:

$$R_L = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{22}|^2} \right| = 4.897 \quad (54)$$

At f=500 MHz:

The center of the load stability at 500 MHz is given by:

$$\bar{C}_L = \frac{S_{11} \cdot \Delta^* - S_{22}^*}{|\Delta|^2 - |S_{22}|^2} = 2.154 + j3.387 \quad (55)$$

where

$$|\bar{C}_L| = 4.014$$

$$\bar{C}_L(\text{angle}) = 57.54^\circ$$

And the radius of the stability circle is given by:

$$R_L = \left| \frac{S_{12} \cdot S_{21}}{|\Delta|^2 - |S_{22}|^2} \right| = 3.024 \quad (56)$$

Applying now the alternative stability factor:

Let us remember that the K factor has already been calculated for the three frequencies under study, and has always been less than 1, an indication of conditional stability. With the following procedure we will try to discard those frequencies in which our source and load reflection coefficients do not conflict with the unstable zones, and we will only have to plot those which overlap. We should first find a testing point within the stability circle, and determine if it causes the input or output reflection coefficient to be greater than 1. If such is the case, then the region confined within the stability circle is said to be unstable. To find this point, I will consider the fact that at these three frequencies:

$$|\vec{C}_S| - R_S < 1$$

and

$$|\vec{C}_L| - R_L < 1$$

Which means that both the source and load stability circle at these three frequencies extends, at least a portion, within the normal Smith Chart.

Because of this, then it is simple to find a point which will belong to both the stability circle and the normal Smith Chart.

For the Source, this point is given by:

$$\Gamma_{SP} = \frac{\bar{C}_s}{|\bar{C}_s|} \quad (57)$$

And is shown in Fig. 11.2:

Fig. 11.2 Test point within the Stability circle.

The same consideration will be applied for the test point in the load.

Finally, going for the calculation for the source:

At f=100 MHz:

The test point is given by:

$$\Gamma_{SP} = \frac{\vec{C}_s}{|\vec{C}_s|} = -0.523 + j0.853$$

And this source reflection coefficient will cause the output reflection coefficient have a magnitude of:

$$|\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{SP}}{1 - S_{11} \cdot \Gamma_{SP}} \right| = 1.84 \quad (58)$$

As $|\Gamma_{OUT}| > 1$, then the inside of the Stability Circle defines the locus of unstable points. Following the convention for the alternative K factor;

$$P_S = 1$$

We now make $\Gamma_{SP} = \vec{d}_S$ and $\delta_S = r_S$, where $\vec{d}_S$ is the vector that goes from the center of the Smith Chart, with an angle equal to the conjugate of S_{11} , to the center of the source constant-gain circle, and r_S is its radius. These two values have already been determined. We are now ready to apply the formula for K_{AS} , at 100 MHz:

$$K_{AS} = \left(\frac{|C_S - \Gamma_{SP}|}{R_S + \delta_S} \right)^{P_S} = 0.896 \quad (59)$$

For f=200 MHz:

The test point is given by:

$$\Gamma_{SP} = \frac{\bar{C}_s}{|\bar{C}_s|} = -0.818 + j0.575$$

And this source reflection coefficient will cause the output reflection coefficient have a magnitude of:

$$|\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{SP}}{1 - S_{11} \cdot \Gamma_{SP}} \right| = 1.559 \quad (60)$$

As $|\Gamma_{OUT}| > 1$, then,

$$P_S = 1$$

Making $\Gamma_{SP} = \bar{d}_S$ and $\delta_S = r_S$,

Then at 200 MHz:

$$K_{AS} = \left(\frac{|C_S - \Gamma_{SP}|}{R_S + \delta_S} \right)^{P_S} = 0.972 \quad (61)$$

At 500 MHz:

The test point at 500 MHz is given by:

$$\Gamma_{SP} = \frac{\bar{C}_s}{|\bar{C}_s|} = -0.997 + j0.073$$

And this source reflection coefficient will cause the output reflection coefficient have a magnitude of:

$$|\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{SP}}{1 - S_{11} \cdot \Gamma_{SP}} \right| = 1.01 \quad (62)$$

As $|\Gamma_{OUT}| > 1$, then,

$$P_S = 1$$

Making $\Gamma_{SP} = \bar{d}_S$ and $\delta_S = r_S$,

Then at 500 MHz:

$$K_{AS} = \left(\frac{|C_S - \Gamma_{SP}|}{R_S + \delta_S} \right)^{P_S} = 1.13 \quad (63)$$

Calculation for the load K_{AL} :

For f=100 MHz:

The test point at 100 MHz is given by:

$$\Gamma_{LP} = \frac{\bar{C}_L}{|\bar{C}_L|} = 0.41 + j0.912$$

And this source reflection coefficient will cause the output reflection coefficient have a magnitude of:

$$|\Gamma_{IN}| = \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{LP}}{1 - S_{22} \cdot \Gamma_{LP}} \right| = 1.801 \quad (64)$$

As $|\Gamma_{IN}| > 1$, then,

$$P_S = 1$$

Making $\Gamma_{LP} = \vec{d}_L$ and $\delta_L = r_L$, where $\vec{d}_L$ and r_L are the position vector and radius of the load constant-gain circle respectively.

Then at 100 MHz:

$$K_{AL} = \left(\frac{|C_L - \Gamma_{LP}|}{R_L + \delta_L} \right)^{P_S} = 0.866 \quad (65)$$

For f=200 MHz:

The test point at 200 MHz is given by:

$$\Gamma_{LP} = \frac{\vec{C}_L}{|\vec{C}_L|} = 0.481 + j0.876$$

And this source reflection coefficient will cause the output reflection coefficient have a magnitude of:

$$|\Gamma_{IN}| = \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{LP}}{1 - S_{22} \cdot \Gamma_{LP}} \right| = 1.421 \quad (66)$$

As $\Gamma_{IN} > 1$, then,

$$P_S = 1$$

Making $\Gamma_{LP} = \vec{d}_L$ and $\delta_L = r_L$,

Where $\vec{d}_L$ and r_L are the position vector and radius of the load constant-gain circle respectively.

Then at 200 MHz:

$$K_{AL} = \left(\frac{|C_L - \Gamma_{LP}|}{R_L + \delta_L} \right)^{P_S} = 0.952 \quad (67)$$

And finally for f=500 MHz:

The test point at 500 MHz is given by:

$$\Gamma_{LP} = \frac{\vec{C}_L}{|\vec{C}_L|} = 0.537 + j0.844$$

And this source reflection coefficient will cause the output reflection coefficient have a magnitude of:

$$|\Gamma_{IN}| = \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_{LP}}{1 - S_{22} \cdot \Gamma_{LP}} \right| = 1.006 \quad (68)$$

As $\Gamma_{IN} > 1$, then,

$$P_S = 1$$

Making $\Gamma_{LP} = \vec{d}_L$ and $\delta_L = r_L$, where $\vec{d}_L$ and r_L are the position vector and radius of the load constant-gain circle respectively.

Then at 500 MHz:

$$K_{AL} = \left(\frac{|C_L - \Gamma_{LP}|}{R_L + \delta_L} \right)^{P_S} = 1.019 \quad (69)$$

These parameters are presented in Table 1.

CONSTANT GAIN CIRCLES										STABILITY CIRCLES							
SOURCE					LOAD					SOURCE				LOAD			
f(MHz)	K factor	Gumax (dB)	S21 (dB)	gs (dB)	ds	rs	gL (dB)	dL	rL	Cs	Cs (angle)	Rs	Kas	CL	CL (angle)	RL	Kal
100	0.296	27.83	23.05	-6	0.15	0.84	-5	0.179	0.798	2.71	121.5	2.24	0.896	2.95	65.8	2.49	0.866
200	0.576	20.97	18.38	-3	0.25	0.7	-3	0.197	0.704	2.76	144.8	2.06	0.972	5.53	61.2	4.89	0.952
500	0.987	13.4	11.12	-1	0.51	0.29	0	0.298	0.298	1.886	175.8	0.892	1.13	4.01	57.5	3.02	1.019

Table 1 Design parameters for the BFQ-34T

By examining Table 1, the alternative K factor is greater than one only at 500 MHz, for both the source and load networks. Then, we will not need to plot the stability circles for 500 MHz, because they do not overlap at this frequency with the source and load constant-gain circles. However, the stability circles will have to be plotted for 100 and 200 MHz, at the same time the matching is performed, to avoid the unstable regions at these frequencies.

3.4.2.9 Impedance Matching

Because of the ease with which series and shunt components can be added in ladder-type arrangements on the Smith Chart, while easily keeping track of the impedance as seen at the input terminals of the structure, the chart seems to be an excellent candidate for an impedance-matching tool. The idea here is simple. Given a load impedance and given the impedance that the source would like to see, simply plot the load impedance and, then begin adding series and shunt elements on the chart until the desired impedance is achieved.

On the chart, the impedance coordinates provide a visual indication of what occurs when a series reactance is added to an

impedance, and the admittance coordinates will tell us what happens when a shunt element is added to an *admittance*. The following equations will have to be used when working with the Smith Chart;

For a series-C component:

$$C = \frac{1}{\omega \cdot X \cdot N}$$

For a series-L component:

$$L = \frac{X \cdot N}{\omega}$$

For a shunt-C component:

$$C = \frac{B}{\omega \cdot N}$$

For a shunt-L component:

$$L = \frac{N}{\omega \cdot B}$$

Where B is the susceptance, ω is the frequency in radians per second and N normalizes the reference impedance of the chart (equal to 50 when normalizing to 50 Ω).

We have two ports (base and collector) to be matched at three different frequencies. In these cases, the use of the classical Smith chart is possible, but extremely tedious and eventually might lead us to errors, because of the constant renormalizations needed when changing

frequencies, and it gets even harder when transmission lines of different characteristic impedances are used, because impedance renormalizations are needed too. Because of all this, I resorted to an impedance matching program freely distributed by MOTOROLA, its name is MIMP. This program provides a simple environment for entering and analyzing impedance matching circuitry. Its main features are:

- a) The Smith chart can be instantly renormalized to any characteristic impedance. All impedances (with interconnecting arcs) are automatically recalculated and redisplayed.
- b) There is an option for overlaying constant Return Loss circles for any complex source impedance.
- c) Multiple transmission line transformations (each with different characteristic impedances) are displayed simultaneously and in exact graphical relationships to each other.
- d) A tabular impedance display is provided to view the impedance at any node.
- e) Constant 'Q' arcs can be added to the Smith Chart.
- f) Real time changes in the impedance transformation are displayed while individual circuit elements are tuned. This utility is provided to perform manual circuit optimization.

Fig. 11.3 Input and output networks.

By means of this program I found the following components values to comply with the restrictions imposed by the constant-gain and stability circles at the three frequencies studied. The schematic is shown in Fig. 11.3.

Their values are presented below:

$$L1=10 \text{ nH}$$

$$L2=6 \text{ nH}$$

$$L3=15 \text{ nH}$$

$$L4=14.6 \text{ nH}$$

$$C1=25 \text{ pF}$$

$$C2=36 \text{ pF}$$

$$C3=53 \text{ pF}$$

$$C4=10 \text{ pF}$$

$$C5=20 \text{ pF}$$

$$C6=7 \text{ pF}$$

$$C7=47 \text{ pF}$$

The input capacitor C_{IN} , is included in the circuit with the only purpose to block the DC, and its value is large enough (1000 pF) to represent a small reactance (much smaller than 50Ω), even at the lowest frequency of operation (100 MHz). The input matching transformations, at each frequency, and plotted along with the constant-gain and stability circles are presented in Figs. 12, 13 and 14.

Fig. 12 Input Matching at 100 MHz, showing the Stability and Constant-Gain circles.

Fig. 13 Input Matching at 200 MHz, showing the Stability and Constant-Gain circles.

Fig. 14 Input Matching at 500 MHz showing the Constant-Gain Circles. The Stability circle is not plotted because K_{AS} is greater than 1.

At this stage of the design, I had two options to follow; work-bench testing, or computer optimization. Before going to the bench, I went a step further in the circuit analysis by using the linear simulation program TOUCHSTONE, Ver. 1.7. The program listing is given below, as presented in its original version, without optimization;

```
!BROADBAND AMPLIFIER DESIGN
!DEVICE USED: BFQ-34T, BIASED AT 10V, 10mA.
!GAIN=12dB, BANDWIDTH=100MHz TO 500MHz
```

```
DIM
  FREQ MHZ
  RES OH
  IND NH
  CAP PF
  LNG MIL
  TIME PS
  COND /OH
  ANG DEG
```

```
VAR
  L1# 1 10 30
  L2# 1 6 30
  L3# 1 15 30
  L4# 1 14.6 30
  C1# 1 25 50
  C2# 1 36 50
  C3# 1 53 100
  C4# 1 10 50
  C5# 1 20 50
  C6# 1 7 50
  C7# 1 47 100
```

```
EQN
```

CKT

CAP 1 0 C^C1

IND 1 2 L^L1

CAP 2 0 C^C2

IND 2 3 L^L2

CAP 3 0 C^C3

S2PA 3 4 0 C:\AA\BFQ34T10.S2P

CAP 4 0 C^C4

IND 4 5 L^L3

CAP 5 0 C^C5

IND 5 6 L^L4

CAP 6 0 C^C6

CAP 6 7 C^C7

DEF2P 1 7 BFQ34AMP

TERM

Z0=50

PROC

OUT

BFQ34AMP DB[S21] GR1

FREQ

SWEEP 40 500 10

GRID

RANGE 100 500 25

GR1 0 20 2

OPT

RANGE 100 500

BFQ34AMP DB[S21]=12

The plot of S_{21} against frequency is given in Fig. 15, in which we see that it moves around 12 dB, but with great deviation for frequencies in-between. This ripple in S_{21} will be smoothed once we apply the optimization routine offered by TOUCHSTONE.

DBIS211 20.00
BFQ34AMP

Fig. 15 Response of S21 by TOUCHSTONE without optimization.

3.4.2.10 Optimization of the original design.

After running the optimization routine in TOUCHSTONE, I found the following values for the matching networks:

$$L1=21.12 \text{ nH}$$

$$L2=7.31 \text{ nH}$$

$$L3=24.31 \text{ nH}$$

$$L4=14.42 \text{ nH}$$

$$C1=12.36 \text{ pF}$$

$$C2=23.96 \text{ pF}$$

$$C3=69.30 \text{ pF}$$

$$C4=3.78 \text{ pF}$$

$$C5=1.53 \text{ pF}$$

$$C6=1.04 \text{ pF}$$

$$C7=12.15 \text{ pF}$$

These values were later tested on the Smith chart, and clearly fit outside the Stability circles. The optimized program listing is given below:

!BROADBAND AMPLIFIER DESIGN
!DEVICE USED: BFQ-34T, BIASED AT 10V, 10mA.
!GAIN=12dB, BANDWIDTH=100MHz TO 500MHz

DIM
FREQ MHZ
RES OH
IND NH
CAP PF
LNG MIL
TIME PS
COND /OH
ANG DEG

VAR
L1# 1 21.12640 30
L2# 1 7.31554 30
L3# 1 24.31199 30
L4# 1 14.42339 30
C1# 1 12.36280 50
C2# 1 23.96571 50
C3# 1 69.30901 100
C4# 1 3.78284 50
C5# 1 1.53108 50
C6# 1 1.04698 50
C7# 1 12.15977 100

EQN

CKT
CAP 1 0 C^C1
IND 1 2 L^L1
CAP 2 0 C^C2
IND 2 3 L^L2
CAP 3 0 C^C3
S2PA 3 4 0 C:\AA\BFQ34T10.S2P
CAP 4 0 C^C4
IND 4 5 L^L3
CAP 5 0 C^C5

```

IND 5 6 L^L4
CAP 6 0 C^C6
CAP 6 7 C^C7
DEF2P 1 7 BFQ34AMP
TERM
Z0=50
PROC

OUT
BFQ34AMP DB[S21] GR1

FREQ

SWEEP 40 500 10

GRID
RANGE 100 500 25
GR1 0 20 2
OPT
RANGE 100 500
BFQ34AMP DB[S21]=12

```

The optimized plot of S_{21} against frequency is given in Fig. 16:

Fig. 15 Response of S21 by TOUCHSTONE after optimization.

3.4.2.11 Work-bench test of the amplifier.

A circuit was implemented with the optimized values offered by TOUCHSTONE. The circuit worked from the beginning, with no oscillation present. However, a peaking at 120 MHz and a roll-off at frequencies above 400 MHz was evident, so I had to trim some values to achieve the flatness needed.

The final capacitor values are presented below. The values for the inductors were measured in the laboratory using a Q meter type 190-A from Boonton Radio Corporation.

$$L1=29 \text{ nH}$$

$$L2=10 \text{ nH}$$

$$L3=20 \text{ nH}$$

$$L4=20 \text{ nH}$$

$$C1= 10 \text{ pF}$$

$$C2= 22 \text{ pF}$$

$$C3= 47 \text{ pF}$$

$$C4= 4.3 \text{ pF}$$

$$C5= 5 \text{ pF}$$

$$C6= 2.0 \text{ pF}$$

$C7 = 12 \text{ pF}$

Looking for higher compression point, the bias was increased to 30 mA, and the circuit amplified linearly up to +14 dBm. After sweeping the circuit, the curve of the measured S_{21} is given in Fig 17.

Fig. 17 Measured response of the amplifier using the BFQ-34T

3.5 DC Amplifier and Readout Circuitry.

A DC amplifier with a gain of 20 was built to amplify the DC voltage rectified by the Schottky diode. As a requirement, a peak detector with a large time constant follows to hold the maximum value read in an interval. The time constant is limited only by the current leakage of the tantalum capacitor C16, the reverse current of D2, and the high but finite input impedance of U3a. By means of the switch S1, the hold state is activated or not. As the operational amplifier, the TL074 was selected due to its JFET (high impedance) inputs.

3.6 Description of the Completed Circuit.

The whole circuit is now completely defined; a ferrite balun in a transmission line configuration is at the input, followed by a DC-controlled attenuator (SBL-1), two stages of small-signal amplification using MARS-6, one stage of low-power RF amplification using the BFQ-34T transistor (replaced the MWA-120), a Schottky diode (5082-2811) as a detector, a DC amplifier with a peak and hold detector, and finally the Liquid Crystal

Readout. The schematic of the final circuit is contained in Fig. 18. The calibrations for the final circuit appear in Figs. 19 to 25.

The response using the MWA-120 hybrid amplifier at 465 is shown in Figs. 19, 20 and 21. The response at 155 MHz appears in Fig. 23, while finally a measurement was done looking for its maximum frequency, found to be at 600 MHz, and Fig. 22 depicts it.

The response using the BFG-34T transistor is shown in Figs. 24 and 25. It demonstrates its higher output capability and slight less gain. No measurement was done at 600 MHz because it is considerably greater than its cut-off frequency.

Fig. 19 Scale 1 response at 465 MHz for the current probe using the MWA-120.

Fig. 20 Scale 2 response at 465 MHz for the current probe using the MWA-120.

Fig. 21 Scale 3 response at 465 MHz for the current probe using the MWA-120

Fig. 22 Scale 1 response at 600 MHz for the current probe using the MWA-120

Fig. 23 Scale 1 response at 155 MHz for the current probe using the MWA-120

Fig. 24 Scale 1 response at 465 MHz for the current probe using the BFQ-34T

Fig. 25 Scale 2 response at 465 MHz for the current probe using the BFQ-34T

3.7 Mechanical Design.

The current probe is built in two halves that are hinged so it may be opened and closed like a clamp-on amp meter. At the hinge points where the two halves meet the coil winding is extended using a flexible piece of wire. At the other points the two windings are interconnected using F-connectors. In Fig. 26 appears the RF current meter.

The resistive winding was made using $100\ \Omega$ 1/8 Watt metallic thin film resistors soldered in series. Metallic thin film resistors were chosen for maximum reliability at high frequencies, and resistors with the minimum available wattage were chosen to minimize the size. In previous non-ferrous current probes the resistive winding has been made using either 1) discrete resistors soldered in series, 2) carbon loaded plastic, or 3) Chromel resistance wire (Hagmann *et al.*, 1993).

Fig. 26 Picture of the high frequency clamp-on current probe.

3.8 Calibration Fixtures.

Calibration is essential for ferrous current probes. However, measurements with non-ferrous current probes (Hagmann *et al.*, 1993) have shown that the transfer impedance is typically within 4 % of the value that is calculated from theory. The errors are only appreciable at frequencies near the upper limit for which the non-ferrous probes are rated. However, a calibration fixture is useful for verifying the sensitivity of the non-ferrous probes, and is also needed in order to determine the response of these probes for frequencies near the upper limit where the low-frequency analysis based on Faraday's law of induction is not appropriate.

The standard practice in testing commercial ferrous current probes (Radmacher, 1990) is to place a probe inside of a shielded coaxial test fixture which is designed so that a known current may be passed through the probe aperture. These test fixtures, which are coaxial TEM cells, consist of a metal box in which the probe is placed, and a center conductor that passes through the probe aperture. The radius of the center conductor is

chosen (ideally 0.434 times the mean radius of the box) such that the test fixture has a characteristic impedance of approximately $50\ \Omega$ for a low VSWR. A signal generator and a $50\ \Omega$ dummy load are connected to the fixture, and the voltage across the load is measured in order to determine the current. The arrangement for this is shown in Fig. 27.

Fig. 27 Instrumentation set-up

Due to the larger size of some of the non-ferrous current probes, it is not possible for box-type fixtures to simultaneously provide 1) a large enough center conductor for an impedance of $50\ \Omega$, and 2) a large enough box so that less than one-third of the cross-section is filled by the probe to limit field perturbations. For this reason, a new type of test fixture was developed (Hagmann *et al.*, 1993) in which a center conductor and the inner surface of the probe shield constitute a coaxial TEM cell. Each of these fixtures has two flat metal plates with a coaxial connector at the center, and a metal cylinder between the pins of the two connectors. The plates are placed on the top and base of the probe so that the cylinder and the shield of the probe form a coaxial transmission line. The radius of the metal cylinder is chosen so that the probe-containing fixture has a characteristic impedance of $50\ \Omega$. The VSWR is typically less than 1.1 over the usable frequency range for the probe. This design also permits testing at higher frequencies because the higher-order modes which disrupt TEM performance have greater cutoff frequencies due to the reduction in size by not using a box.

An improved version of the latter test fixture design was used in the measurements for this thesis. In this device, the two plates are made

concave in order to shorten the length of the center conductor. Thus, due to the decreased length of the test fixture, it is not necessary to have an impedance match, so the diameter of the center conductor is not critical. Again, the VSWR is typically less than 1.1 over the usable frequency range for the probe.

The precision required in assembling the test fixture decreases at lower frequencies. For example, the need for closing the ends of the test fixture by metal plates decreases, and we have found that the plates are not required at all for frequencies on the order of 1 MHz or less. Others have also used both open and closed test fixtures at different frequencies (Blackwell, 1990).

4. Conclusions.

The research done for this thesis has resulted in solution of the problem for the thesis as it was stated in section 1.6. The problem to be solved was that of developing a means for reducing or eliminating the effects of cable pickup so that non-ferrous current probes may be used as dosimeters in order to determine the current induced in humans exposed to electromagnetic fields. The first approach that was taken in order to

accomplish this task was the development of a preamplifier to be attached to a non-ferrous current probe in order to provide sufficient amplification so that the amplified signal exceeds the errors added due to cable pickup. The second approach was the development of a self-contained device that amplifies and detects the output from a non-ferrous current probe and provides an indication of the current with a readout device such as a digital meter. Both of these approaches have been followed successfully.

The IEEE guidelines for human exposure (IEEE, 1992) specify upper limits for the RF currents induced in humans. Current values are only specified at the point of contact between the feet and ground, and at the point of contact between the hands and various objects. Furthermore, the current values are only specified at frequencies up to 100 MHz. It appears that the reason that there is no specification limiting the current values at various points within the human body (instead of only at points of contact) and at frequencies above 100 MHz, is due to a lack of suitable instrumentation for making such measurements. The work done for this thesis has resulted in several new instruments that extend the ability to measure RF currents in human dosimetry and other applications.

The specific accomplishments of the research for this thesis are as follows:

1. A preamplifier was developed that may be attached directly to a non-ferrous current probe in order to provide sufficient amplification so that the amplified signal exceeds the errors added to the signal due to cable pickup. This instrument was tested successfully in measurements with human subjects exposed to RF fields in an anechoic chamber at Brooks Air Force Base. A description of the instrument and the measurements is contained in section 2.
2. A self-contained current meter for use at frequencies from 300 kHz to 70 MHz was constructed by making a battery-powered amplifier and detection circuit with an analog panel meter that is connected directly to a ferrous current probe. This instrument was tested successfully in measurements with human subjects at an AM radio station. A description of the instrument and the measurements is contained in section 3.1 and Appendix III.
3. A self-contained current meter for use at frequencies up to 600 MHz was constructed which can be used as a clamp-on meter and detect currents as low as 8 mA. The final circuit uses a RF mixer as a variable attenuator in order to increase the dynamic range, two Monolithic

Microwave Integrated Circuits (MMIC) for preamplification, a final broadband amplifier to raise the output compression point, a Schottky diode detector, a sample and hold circuit, and a liquid crystal digital panel meter. A description of this instrument is contained in section 3 of this thesis.

References

R. H. Abhold, M. J. Ortner, M. J. Galvin and D. I. McRee (1981), "Studies on Acute In Vivo Exposure of Rats to 2450 MHz Microwave Radiation: II. Effects on Thyroid and Adrenal Axes Hormones," Radiation Research, vol. 88, pp. 448-455.

ACGIH (1988), Threshold Limits and Biological Exposure Indices for 1988-1989, American Conference of Governmental Industrial Hygienists, Cincinnati, Ohio.

E. R. Adair, D. E. Spiers, R. O. Rawson, B. W. Adams, D. K. Shelton, P. J. Pivrotto and G. M. Akel (1985), "Thermoregulatory Consequences of Long-Term Microwave Exposure at Controlled Ambient Temperatures," Bioelectromagnetics, vol. 6, pp. 339-363.

W. R. Adey, S. M. Bawin and A. F. Lawrence (1982), "Effects of Weak Amplitude-Modulated Microwave Fields on Calcium Efflux from Awake Cat Cerebral Cortex," Bioelectromagnetics, vol. 3, pp. 295-307.

R. Albanese, J. Blaschak, R. Medina and J. Penn (1994), "Ultrashort Electromagnetic Signals: Biophysical Questions, Safety Issues, and Medical Opportunities," Aviation Space & Environmental Medicine, vol. 65, Suppl. 5, pp. A116-120.

E. N. Albert and J. M. Kerns (1981), "Reversible Microwave Effects on the Blood-Brain Barrier," Brain Research, vol. 230, pp. 153-164.

L. E. Anderson (1991), "ELF: Exposure Levels, Bioeffects, and Epidemiology," Health Physics, vol. 61, pp. 41-46.

L. E. Anderson (1993), "Biological Effects of Extremely Low-Frequency Electromagnetic Fields: In Vivo Studies," American Industrial Hygiene Association Journal, vol. 54, pp. 186-196.

ANSI C95.1 (1982), Safety Levels with respect to Human Exposure to Radiofrequency Electromagnetic Fields, 300 kHz to 100 GHz, American Standard C95.1, American National Standards Institute, New York.

E. Aslan (1971), "Electromagnetic Leakage Survey Meter," Journal of Microwave Power, vol. 6, pp. 169-177.

C. A. Balanis (1982), Antenna Theory, New York: Harper & Row, pp. 365-368.

S. Baranski and P. Czerski (1976), Biological Effects of Microwaves, Stroudsburg, PA: Dowden, Hutchinson and Ross.

C. E. Baum, E. L. Breen, J. C. Giles, J. O'Neill and G. D. Sower (1978), "Sensors for Electromagnetic Pulse Measurements Both Inside and Away from Nuclear Source Regions," IEEE Transactions on Antennas and Propagation, vol. 26, pp. 22-35.

S. M. Bawin and W. R. Adey (1977), "Sensitivity of Calcium Binding in Cerebral Tissue to Weak Environmental Electric Fields Oscillating at Low Frequency," Proceedings of the National Academy of Science (USA), vol. 73, pp. 1999-2003.

H. I. Bassen (1983), "Electric Field Probes--A Review," IEEE Transactions on Antennas and Propagation, vol. 31, pp. 710-718.

C. F. Blackman, S. G. Benane, D. E. House and W. T. Joines (1985), "Effects of ELF (1-120 Hz) and Modulated (50 Hz) RF Fields on the Efflux of Calcium Ions from Brain Tissue," Bioelectromagnetics, vol. 6, pp. 1-11.

R. P. Blackwell (1990), "The personal current meter--A novel ankle-worn device for the measurement of RF body current in a mobile subject," Journal of Radiological Protection (UK), vol. 10, pp. 109-114.

R. R. Bowman (1976), "A probe for measuring temperature in radio-frequency-heated material," IEEE Transactions on Microwave Theory and Techniques, vol. 24, pp. 43-45.

P. Brodeur (1977), The Zapping of America: Microwaves, Their Deadly Risk, and the Cover-Up, New York, Norton.

P. Brodeur (1989), Currents of Death: Power Lines, Computer Terminals, and the Attempt to Cover Up Their Threat to Your Health, New York, Simon and Schuster.

P. Brodeur (1993), The Great Power-Line Cover-Up: How the Utilities and the Government are Trying to Hide the Cancer Hazard Posed by Electromagnetic Fields, Boston, Little, Brown and Co.

M. Cabanac (1983), "Thermoregulatory Behavioral Responses, pp. 307-357, in E. R. Adair editor, Microwaves and Thermoregulation, New York: Academic Press.

Canada (1979), Recommended Safety Procedures for the Installation and use of Radiofrequency and Microwave Devices in the Frequency Range 10 MHz-300 GHz, Publication 79-EHD-30, Environmental Health Directorate, Health and Welfare Canada.

C. K. Charny, M. J. Hagmann and R. L. Levin (1987), "A Whole Body Thermal Model of Man During Hyperthermia," IEEE Transactions on Biomedical Engineering, vol. 34, pp. 375-387.

I. Chatterjee, D. Wu and O. P. Gandhi (1986), "Human Body Impedance and Threshold Currents for Perception and Pain for Contact Hazard Analysis in the VLF-MF Band," IEEE Transactions on Biomedical Engineering, vol. 33, pp. 486-494.

C. K. Chou, A. W. Guy and R. Galambos (1982), "Auditory Perception of Radio-Frequency Electromagnetic Fields," Journal of the Acoustical Society of America, vol. 71, pp. 1321-1334.

C. K. Chou, A. W. Guy, L. E. Borneman, L. L. Kunz and P. Kramar (1983), "Chronic Exposure of Rabbits to 0.5 and 5 mW/cm² 2450 MHz CW Microwave Radiation," Bioelectromagnetics, vol. 4, pp. 63-77.

D. A. Christensen (1977), "A New Nonperturbing Temperature Probe using Semiconductor Band Edge Shift," Journal of Bioengineering, vol. 1, pp. 541-546.

S. E. Cleary (1993), "A Review of In Vitro Studies: Low-Frequency Electromagnetic Fields," American Industrial Hygiene Association Journal, vol. 54, pp. 178-185.

Commonwealth of Massachusetts (1983), Regulations Governing Fixed Facilities Which Generate Electromagnetic Fields in the Frequency Range of 300 kHz to 100 GHz and Microwave Ovens, 105 CMR 122.000, Massachusetts Register Issue No. 379.

C. F. Dalziel and T. H. Mansfield (1950), "Effect of Frequency on Perception Currents," Transactions of the American Institute of Electrical Engineers, vol. 69, Part II, pp. 1162-1168.

C. F. Dalziel and W. R. Lee (1969), "Lethal Electric Currents," IEEE Spectrum, vol. 6, pp. 44-50.

C. H. Durney (1980), "Electromagnetic Dosimetry for Models of Humans and Animals: A Review of Theoretical and Numerical Techniques," Proceedings of the IEEE, vol. 68, pp. 33-40.

C. H. Durney, H. Massoudi and M. F. Iskander (1986) Radiofrequency Radiation Dosimetry Handbook, USAFSAM-TR-85-73, USAF School of Aerospace Medicine, Brooks Air Force Base, Texas.

K. R. Foster and E. D. Finch (1974), "Microwave Hearing: Evidence for Thermoacoustic Auditory Stimulation by Pulsed Microwaves," Science, vol. 185, pp. 256-258.

K. R. Foster (1992), "Health Effects of Low-Level Electromagnetic Fields: Phantom or Not-So-Phantom Risk?," Health Physics, vol. 62, pp. 429-435.

A. H. Frey (1971), "Biological Function as Influenced by Low-Power Modulated RF Energy," IEEE Transactions on Microwave Theory and Techniques, vol. 19, pp. 153-164.

A. H. Frey and L. S. Wesler (1983), "Dopamine Receptors and Microwave Energy Exposure," Journal of Bioelectricity, vol. 2, pp. 145-157.

H. Froehlich (1975), "The Extraordinary Dielectric Properties of Biological Materials and the Action of Enzymes," Proceedings of the National Academy of Science (USA), vol. 72, pp. 4211-4215.

O. P. Gandhi (1975), "Conditions of Strongest Electromagnetic Power Deposition in Man and Animals," IEEE Transactions on Microwave Theory and Techniques, vol. 23, pp. 1021-1029.

O. P. Gandhi, I. Chatterjee, D. Wu, and Y.-G. Gu (1985), "Likelihood of High Rates of energy Deposition in the Human Legs at the ANSI Recommended 3-30 MHz RF Safety Levels," Proceedings of the IEEE, vol. 73, pp. 1145-1147.

O. P. Gandhi, J.-Y. Chen and A. Riazzi (1986), "Currents induced in a human being for plane-wave exposure conditions 0-50 MHz and for RF sealers," IEEE Transactions on Biomedical Engineering, vol. 33, pp. 757-767.

O. P. Gandhi (1987), RF Personnel Dosimeter and Dosimetry Method for use therewith, U. S. Patent Number 4,672,309, issued June 9, 1987.

F. C. Garland, E. Shaw, E. D. Gorham, . F. Garland, M. R. White and P. J. Sinsheimer (1990), "Incidence of Leukemia in Occupations with Potential Electromagnetic Field Exposure in United States Navy Personnel," American Journal of Epidemiology, vol. 132, pp. 293-303.

T. G. Giallorenzi, J. A. Bucaro, A. Dandridge and J. H. Cole (1986), "Optical-Fiber Sensors Challenge the Competition," IEEE Spectrum, vol. 23, pp. 44-49.

M. Gochfeld (1990), "Microelectronics, Radiation and Superconductivity," Environmental Health Perspectives, vol. 86, pp. 285-289.

R. Goodman, Y. Chizmadzhev and A. Shirley-Henderson (1993), "Electromagnetic Fields and Cells," Journal of Cellular Biochemistry, vol. 51, pp. 436-441.

C. F. Gottlieb, F. L. Moffat, M. J. Hagmann, T. M. Babij, A. A. Abitbol, A. A. Lewin, P. V. Houdek and J. G. Schwade (1990), "Preclinical Evaluation of Submillimeter Diameter Microwave Interstitial Hyperthermia Applicators," Journal of Microwave Power and Electromagnetic Energy, vol. 25, pp. 149-160.

J. Grebenkemper (1987), "The Tandem Match--An Accurate Directional Wattmeter," QST (Journal of the American Radio Relay League), vol. 71, pp. 18-26.

F. M. Green (1975), Development of Electric and Magnetic Near-Field Probes, NBS Technical Note number 658.

W. Grobner and N. Hofreiter (1966), Integraltafel Zweiter Teil Bestimmte Integrale, Berlin: Springer-Verlag, p. 331.

K.-L. Gronhaug (1986), Measurements and Calculations of Current Induced in a Human Body by EMP Illumination, FFI/NOTAT-86/4008, Norwegian Defence Research Establishment.

K.-L. Gronhaug (1988), Measurements of EMP Induced Currents in the Human Body, FFI/NOTAT-88/4038, Norwegian Defence Research Establishment.

W. Grundler, F. Keilmann and H. Frohlich (1977), "Resonant Growth Rate Response of Yeast Cells Irradiated by Weak Microwaves," Physics Letters, vol. 62A, pp. 463-466.

A. W. Guy (1971), "Analysis of Electromagnetic Fields Induced in Biological Tissues by Thermographic Studies on Equivalent Phantom Models," IEEE Transactions on Microwave Theory and Techniques, vol. 19, pp. 205-214.

A. W. Guy, M. D. Webb and C. C. Sorensen (1976), "Determination of Power Absorption in Man Exposed to High-Frequency Electromagnetic Fields by Thermographic Measurements on Scale Models," IEEE Transactions on Biomedical Engineering, vol. 23, pp. 361-371.

A. W. Guy, P. O. Kramar, C. A. Harris and C.-K. Chou (1980), "Long-Term 2450-MHz CW Microwave Irradiation of Rabbits: Methodology and

Evaluation of Ocular and Physiologic Effects," Journal of Microwave Power, vol. 15, pp. 37-44.

A. W. Guy (1989), Analysis of Time Domain Induced Current and Total Absorbed Energy in Humans Exposed to EMP Electric Fields, Final report, prepared for ERC Facilities Service Corporation, Fairfax, VA.

A. W. Guy (1992), "Electromagnetic Fields and Health: Some Thoughts about the Past and Future," Bioelectromagnetics, vol. 13, pp. 601-604.

M. J. Hagmann, O. P. Gandhi and C. H. Durney (1979A), "Numerical Calculation of Electromagnetic Energy Deposition for a Realistic Model of Man," IEEE Transactions Microwave Theory and Techniques, vol. 27, pp. 804-809.

M. J. Hagmann and O. P. Gandhi (1979B), "Numerical calculation of electromagnetic energy deposition in models of man with grounding and reflector effects," Radio Science, vol. 14, pp. 23-29.

M. J. Hagmann and R. L. Levin (1986), "Aberrant Heating: A Problem in Regional R-F Hyperthermia," IEEE Transactions on Biomedical Engineering, vol. 33, pp. 405-411.

M. J. Hagmann and T. M. Babij (1990A), High Frequency Ammeter, U. S. Patent No. 4,897,600, issued January 30, 1990.

M. J. Hagmann and T. M. Babij (1990B), Personal Dosimeter, U. S. Patent No. 4,913,153, issued April 3, 1990.

M. J. Hagmann (1993), "Human Dosimetry in EMP Exposures with Minimally-Perturbing Current Probes," in EMP Human Health Effects Review Panel Proceedings, published by the National Technical Information Service, U. S. Department of Commerce, pp. 59-78.

M. J. Hagmann and T. M. Babij (1993), "Noninvasive measurement of current in the human body for electromagnetic dosimetry," IEEE Transactions on Biomedical Engineering, vol. 40, pp. 418-423.

M. J. Hagmann (1994), personal communication.

S. Hamburger, J. N. Logue and P. M. Silverman (1983), "Occupational Exposure to Non-Ionizing Radiation and an Association with Heart Disease: An Exploratory Study," Journal of Chronic Diseases, vol. 36, pp. 791-802.

W. R. Hendee and J. C. Boteler (1994), "The Question of Health Effects from Exposure to Electromagnetic Fields," Health Physics, vol. 66, pp. 127-136.

Hewlett Packard (1975), "Dynamic Range Extension of Schottky Detectors," Application Note 956-5.

Hewlett Packard (1981), "Square Law and Linear Detection," Application Note 986.

Hewlett Packard (1982), "Is Bias Current Necessary," Application Note 987.

A. H. Hilbers (1970), "Design of H. F. Wideband Power Transformers," in Bipolar & MOS Transmitting Transistors, Application Reports, N. V. Phillips, Nijmegen, the Netherlands.

D. Hill and J. A. Walsh (1985), "Radiofrequency Current through the Feet of a Grounded Human," IEEE Transactions on Electromagnetic Compatibility, vol. 27, pp. 18-23.

IEEE (1992), IEEE Standard for Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields, 3 kHz to 300 GHz. New York, IEEE.

IRPA (1984), "Interim Guidelines on Limits of Exposure to Radiofrequency Electromagnetic Fields in the Frequency Range from 100 kHz to 300 GHz," Health Physics, vol. 46, pp. 975-978.

J. Jauchem (1993), "Alleged Health Effects of Electric or Magnetic Fields: Additional Misconceptions in the Literature," Journal of Microwave Power & Electromagnetic Energy, vol. 28, pp. 140-155.

R. P. Jensh (1984), "Studies of the Teratogenic Potential of Exposure of Rats to 6000 MHz Microwave Radiation--I. Morphologic Analysis at Term," Radiation Research, vol. 97, pp. 272-281.

C. C. Johnson and T. C. Rozzell (1975), "Liquid Crystal Fiberoptic RF Probes, Part I: Temperature Probe for Microwave Fields," Microwave Journal, vol. 18, pp. 55-57.

D. R. Justesen (1975), "Toward a Prescriptive Grammar for the Radiobiology of Non-Ionizing Radiations: Quantities, Definitions, and Units of Absorbed Electromagnetic Energy," Journal of Microwave Power, vol. 10, pp. 333-356.

B. Kallen, G. Malmquist and U. Moritz (1982), "Delivery Outcome Among Physiotherapists in Sweeden: Is Non-Ionizing Radiation a Fetal Hazard?," Archives of Environmental Health, vol. 37, pp. 81-85.

P. Kolodner and J. A. Tyson,(1983), "Remote Thermal Imaging with 0.7-m Spatial Resolution using Temperature-Dependent Fluorescent Thin Films," Applied Physics Letters, vol. 42, pp. 117-119.

J. D. Kraus (1988), Antennas, New York: McGraw-Hill, 2nd ed, pp. 741-743.

J. D. Kraus (1992), Electromagnetics, New York: McGraw-Hill, 4th ed., pp. 485-488.

E. Marg (1991), "Magnetostimulation of Vision: Direct Noninvasive Stimulation of the Retina and the Visual Brain," Optometry & Vision Science, vol. 68, pp. 427-440.

W. A. Merman and D. M. Witters (1980), Microwave Hazard Instruments: an Evaluation of the Narda 8100, Holaday HI-1500, and Simpson 380M, Technical Report FDA 80-8122, Food and Drug Administration, Bureau of Radiological Health.

S. Milham, Jr. (1982), "Mortality from Leukemia in Workers Exposed to Electrical and Magnetic Fields," New England Journal of Medicine, vol. 304, p. 249.

E. K. Miller (1986), Time-Domain Measurements in Electromagnetics, New York: Van Nostrand Reinhold, pp. 283-286.

R. Morrison (1986), Grounding and Shielding Techniques in Instrumentation, New York: John Wiley, 3rd ed.

Motorola (1991), "Three Balun designs for Push-Pull Amplifiers," Application Note ANA 1034 in Motorola RF Device Data, vol. II, pp. 7:255-260.

J. M. Muhm (1992), "Mortality Investigation of Workers in an Electromagnetic Pulse Test Program," Journal of Occupational Medicine, vol. 34, pp. 287-292.

W. W. Mumford (1969), "Heat Stress due to RF Radiation," Proceedings of the IEEE, vol. 57, pp. 171-178.

R. R. Neutra (1992), "How to Approach Complex Mixtures: Lessons from the Epidemiology of Electromagnetic Fields," Public Health Reviews, vol. 19, pp. 1-17.

NRPB (1986), Advice on the Protection of Workers and Members of the Public from the Possible Hazards of Electric and Magnetic Fields with Frequencies below 300 GHz: A Consultative Document, London, HMSO.

R. G. Olsen and T. A. Griner (1989), "Outdoor Measurement of SAR in a Full-Sized Human Model Exposed to 29.9 MHz in the Near Field," Bioelectromagnetics, vol. 10, pp. 161-171.

A. S. Pressman (1970), Electromagnetic Fields and Life, New York: Plenum Press.

D. Radmacher 1990, Eaton Corporation, Electronic Instrumentation Division, Los Angeles, CA, Personal Communication.

M. H. Repacholi (1990), "Cancer from Exposure to 50/60 Hz Electric and Magnetic Fields--A Major Scientific Debate," Australasian Physical & Engineering Sciences in Medicine, vol. 13, pp. 4-17.

L. W. Ricketts, J. E. Bridges and J. Miletta (1976), EMP Radiation and Protective Techniques, New York: John Wiley, p. 231.

R. J. Rogers (1981), "Radiofrequency Burn Hazards in the MF/HF Band," pp. 76-89 in J. C. Mitchell, ed., Proceedings of a Workshop on the Protection of Personnel Against Radiofrequency Electromagnetic Radiation, Report USAFSAM-TR-81-28, USAF School of Aerospace Medicine, Aerospace Medical Division, Brooks Air Force Base, Texas.

T. Rozzell, C. C. Johnson, C. H. Durney, J. L. Lords and R. G. Olsen (1974), "A Nonperturbing Temperature Sensor for Measurements in Electromagnetic Fields, Journal of Microwave Power, vol. 9, pp. 241-246.

A. W. Rudge and R. M. Knox (1970), Near Field Instrumentation, Technical Report BRH/DEP 70-16, Bureau of Radiological Health, U. S. Public Health Service (NTIS order number PB192748).

C. L. Ruthroff (1959), "Some Broad-Band Transformers," Proceedings of the IRE, vol. 47, pp. 1337-1342.

T. M. Schnorr, B. A. Grajewski, R. W. Hornung, M. J. Thun, G. M. Egeland, W. E. Murray, D. L. Conover and W. E. Halperin (1991), "Video Display Terminals and the Risk of Spontaneous Abortion," New England Journal of Medicine, vol. 324, pp. 727-733.

G. H. Schreiber, G. M. Swaen, J. M. Meijers, J. J. Slangen and F. Stumans (1993), "Cancer Mortality and Residence Near Electricity

Transmission Equipment: A Retrospective Cohort Study," International Journal of Epidemiology, vol. 22, pp. 9-15.

H. P. Schwan and K. R. Foster (1980), "RF-Field Interactions with Biological Systems: Electrical Properties and Biophysical Mechanisms," Proceedings of the IEEE, vol. 68, pp. 104-113.

G. M. Shaw and L. A. Croen (1993), "Human Adverse Reproductive Outcomes and Electromagnetic Field Exposures: Review of Epidemiologic Studies," Environmental Health Perspectives, vol. 101, Supplement 4, pp. 107-119.

L. L. Sher (1963), Mechanical Effects of AC Fields on Particles Dispersed in a Liquid: Biological Implications, Ph. D. Thesis, University of Pennsylvania, Philadelphia, PA.

H. Shuhao (1987), "The Balun Family," Microwave Journal, vol. 30, pp. 227-229.

R. J. Smialowicz, C. M. Weil, P. Marsh, M. M. Riddle, R. R. Rogers and B. F. Rehnberg (1981), "Biological Effects of Long-Term Exposure of Rats to 970 MHz Radiofrequency Radiation," Bioelectromagnetics, vol. 2, pp. 279-284.

J. A. J. Stolwijk (1983), "Evaluation of Thermoregulatory Response to Microwave Power Deposition," pp. 297-305 in E. R. Adair, editor, Microwaves and Thermoregulation, New York, Academic Press.

R. D. Stoy, K. R. Foster and H. P. Schwan (1982), "Dielectric properties of mammalian tissues from 0.1 to 100 MHz: A summary of recent data," Physical Medicine and Biology, vol. 27, pp. 501-513.

M. A. Stuchly, J. A. Kozlowski, S. Symons and D. W. Lecuyer (1991), "Measurements of Contact Currents in Radiofrequency Fields," Health Physics, vol. 60, pp. 547-557.

F. L. Tabrah and S. Batkin (1991), "Electromagnetic Fields: Biological and Clinical Aspects," Hawaii Medical Journal, vol. 50, pp. 113-118.

L. S. Taylor (1981), "The Mechanisms of Athermal Microwave Biological Effects," Bioelectromagnetics, vol. 2, pp. 259-267.

R. A. Tell (1990), "RF Hot Spot Fields: The Problem of Determining Compliance with the ANSI Radiofrequency Protection Guide," in Proceedings of the 44th Annual Broadcast Engineering Conference, National Association of Broadcasters, Atlanta GA.

T. S. Tenforde (1992), "Biological Interactions and Potential Health Effects of Extremely-Low-Frequency Magnetic Fields from Power Lines and

Other Common Sources," Annual Reviews of Public Health, vol. 13, pp. 173-196.

J. E. Vena, S. Graham, R. Hellmann, M. Swanson and J. Brasure (1991), "Use of Electric Blankets and Risk of Postmenopausal Breast Cancer," American Journal of Epidemiology, vol. 134, pp. 180-185.

M. L. Walsh, S. M. Harvey, R. A. Facey and R. R. Mallette (1991), "Hazard Assessment of Video Display Units," American Industrial Hygiene Association Journal, vol. 52, pp. 324-331.

W. L. Weeks (1968), Antenna Engineering, New York: McGraw-Hill, pp. 161-180.

K. R. Wickersheim (1990), "On-Line Measurements in Microwave Ovens," Tappi Journal, vol. 73, pp. 223-229.

G. M. Wilkening and C. H. Sutton (1990), "Health Effects of Nonionizing Radiation," Medical Clinics of North America, vol. 74, pp. 489-507.

T. H. Windhorn and C. H. Cain (1979), "Optically Active Binary Liquid Crystal Thermometry," IEEE Transactions on Biomedical Engineering, vol. 26, pp. 148-152.

O. M. Woodward, Jr. (1953), "Balance Measurements on Balun Transformers," Electronics, vol. 26, September 1953, pp. 188-191.

World Health Organization (1981), Environmental Health Criteria 16, Radiofrequency and Microwaves, ISBN: 9241540761, Geneva: World Health Organization.

M. G. Yost (1992), "Occupational Health Effects of Nonionizing Radiation," Occupational Medicine: State of the Art Reviews, vol. 7, pp. 543-566.

Appendix I

Analysis for Ferrous Current Probes

Figure A.1 shows a ferrous toroidal core having a length c , a cross-sectional area A , and a permeability μ , that surrounds a wire filament carrying a current I . A coil of N turns which is wound on the core has an output voltage of V . The magnetic flux ϕ must be a constant throughout the full length of the core. Therefore, since the cross-sectional area is constant, and the permeability is constant, the magnitude of the magnetic field intensity H must also be a constant throughout the full length of the core. Thus, Ampere's law requires that the magnetic field intensity $H = I/c$, regardless of the position of the wire, as long as the wire is contained within the aperture of the probe. Therefore, the magnetic flux $\phi = \mu AI/c$. At relatively low frequencies, when the length $c \ll \lambda$, Faraday's law of induction requires that the output voltage $V = (\mu NA/c) dI/dt$. Therefore, the ferrous current probe will act as a transformer with a mutual impedance equal to $\mu NA/c$. For a sinusoidal current with angular frequency ω radians/second, the transfer impedance defined by $Z_T \equiv V/I$ is given by

$$Z_T = \frac{j \cdot \omega \cdot \mu \cdot N \cdot A}{c}$$

Fig. A.1 Analysis of a ferrous probe.

Appendix II

Analysis for Non-Ferrous Current Probes

Figure A.2 shows a coil of N turns that is wound evenly over the full length of a non-ferrous toroidal core having a cross-sectional area A . The toroid surrounds a wire filament carrying a current I . The outer and inner surfaces of the toroid are represented by two circles, and an increment of winding has been shaded. In the triangle shown in Fig. A.2, the distance from the center of the toroid to the current filament is d , and the distances from the center of the increment of winding to the center of the toroid and the current filament are r and s , respectively.

It may be shown from trigonometry that

$$\cos\alpha = \sqrt{1 - \frac{d^2 \sin^2 \theta}{s^2}} \quad (1)$$

and also that

Fig. A.2 Analysis of a non-ferrous probe

$$s^2 = r^2 + d^2 - 2rd \cos \theta \quad (2)$$

From Eqs. (1) and (2),

$$s \cos \alpha = r - d \cos \theta \quad (3)$$

From Eqs. (2) and (3),

$$\frac{\cos \alpha}{s} = \frac{r - d \cos \theta}{r^2 - 2rd \cos \theta + d^2} \quad (4)$$

The value of the magnetic field intensity normal to the increment of winding is given by

$$H_n = \frac{I \cdot \cos \alpha}{2 \cdot \pi \cdot s} \quad (5)$$

From Eqs. (4) and (5),

$$H_n = \frac{I}{2\pi} \cdot \left(\frac{r - d \cos \theta}{r^2 - 2rd \cos \theta + d^2} \right) \quad (6)$$

For a coil with small cross-section $A \ll r^2$, at relatively low frequencies when the radius $r \ll \lambda$, Faraday's Law of induction may be used with Eq. (6) to obtain the following expression for the voltage induced on the increment of winding:

$$dV = \frac{\mu_0}{2\pi} \cdot \left(\frac{r - d \cos \theta}{r^2 - 2rd \cos \theta + d^2} \right) \cdot \left(\frac{dI}{dt} \right) \cdot A \cdot dn \quad (7)$$

where dn is the (fractional number of turns on the increment of winding.

Thus, the total potential induced in the coil is given by

$$V = \frac{\mu_0 NA}{(2\pi)^2} \cdot \left(\frac{dI}{dt} \right) \cdot \int_0^{2\pi} \left(\frac{r - d \cos \theta}{r^2 - 2rd \cos \theta + d^2} \right) d\theta \quad (8)$$

Rearranging terms in Eq. (8),

$$V = \frac{\mu_0 NA}{(2\pi)^2} \cdot \left(\frac{dI}{dt} \right) \cdot \int_0^{2\pi} \left(\frac{1 - \frac{d}{r} \cos \theta}{1 - \frac{2d \cos \theta}{r} + \frac{d^2}{r^2}} \right) \cdot \frac{d\theta}{2\pi} \quad (9)$$

The integral in (9) has the value of 1 for $d < r$, $1/2$ for $d = r$, and 0 for $d > r$ (Grobner *et al.*, 1966). Thus, there is no output when the current-carrying element is outside of the aperture, and when the current is located anywhere within the aperture, the output is given by

$$V = \frac{\mu_0 N A}{c} \bullet \frac{dI}{dt} \quad (10)$$

where $c = 2\pi r$ is the circumference of the coil (the total length of the winding). But Eq. (10) is exactly the expression derived in Appendix I for a current probe with a ferrous core, but with μ replaced by μ_0 .

Appendix III

Report of Measurements at an AM Radio Station

On July 25, 1994 Dr. David Conover and Dr. Gene Moss of NIOSH, Charles Langford of OSHA, and Dr. Mark J. Hagmann and George Cabrera of Florida International University visited an AM radio station that will be unnamed for legal reasons. The Station Manager, Chief Engineer, and a consultant from Jules Cohen & Associates in Washington, D. C. were also present during the measurements. This visit was precipitated by complaints of the personnel at that facility which alleged that they had various physiological effects which they associated with exposure to intense RF fields.

Measurements of body current were made using an Eaton model 94606-2 ferrous current probe having a measured transfer impedance of $1.0 \Omega \pm 20 \%$ in the frequency range from 300 kHz to 70 MHz. While we prefer to use our non-ferrous current probes, a ferrous current probe was required for this application in order to have adequate sensitivity at the sub-MHz frequency of the AM radio station. The current probe acts as a

transformer, so that when the toroid is placed around a current-carrying member such as the arm or leg, an output of 1 Volt is generated for each Ampere of current (transfer impedance of 1 Ω). Readout from the probe was accomplished using a battery-powered amplifier and detection circuit with a panel meter that is housed in a small shielded box.

The total measurement system was calibrated immediately after the measurements, at the frequency of the AM transmitter. A signal generator set for an output of -30 dBm and connected to a 50 Ω dummy load was used as the current source (0.1414 mA rms) for calibration. This source gave readings of 135, 35 and 9 units on the 0-150 panel meter using the three sensitivity settings for the circuit. Fixed coaxial attenuators were connected between the current probe and the readout box when necessary in order to increase the dynamic range for the measurements. The following data were reduced on the basis of this calibration.

Tall male subject with back near the wall, touching column near the center of the second floor of the station (apparently the worst case for the subject on this floor of the building), shoulder 5 mA, ankle 2 mA.

Female subject touching the same column, shoulder 3 mA, ankle 4 mA.

Tall male subject in maintenance area by window (apparently the worst case for the subject in this room), shoulder 0.3 mA.

Tall male subject in the garage, touching the vertical door track (apparently the worst case for the subject in this room), shoulder 3 mA.

Tall male subject outside the building, in the parking lot by a car (apparently the worst case for the subject immediately outside the building), shoulder 3 mA.

Tall male subject outside the building, touching the door in a fence surrounding the front antenna (apparently the worst case for the subject, as close as he can get to this antenna), shoulder 12 mA.

Tall male subject outside the building, touching a fence surrounding the rear antenna (apparently the worst case for the subject, as close as he can get to this antenna), shoulder 14 mA.

The maximum current measured at the station was 14 mA, and the maximum current measured in building, in any of the areas where the complaints had been made, was 5 mA. At no time did any current cause pain or other sensation. The IEEE guidelines for human exposure at the frequency of the station give a MPE (maximum permitted exposure) for the current through either foot or hand of 45 mA for uncontrolled environments, and 100 mA for controlled environments.

NETWORK CHARACTERIZATION

A two-port device (Fig. A.3) can be described by a number of parameter sets. The most common systems are the H, Y and Z-parameter sets. All of these network parameters relate **total voltages** and **total currents** at each of the two ports. Below are presented these input and output total voltages and currents, and how they are treated as dependent or independent variables, according to the parameter set used.

Fig.A.3 Characterization of a two-port network

H-Parameters: $V_1 = h_{11} \cdot I_1 + h_{12} \cdot V_2$

(1)

$$I_2 = h_{21} \cdot I_1 + h_{22} \cdot V_2$$

Y-Parameters: $I_1 = y_{11} \cdot V_1 + y_{12} \cdot V_2$

(2)

$$I_2 = y_{21} \cdot V_1 + y_{22} \cdot V_2$$

Z-Parameters: $V_1 = z_{11} \cdot I_1 + z_{12} \cdot I_2$

(3)

$$V_2 = z_{21} \cdot I_1 + z_{22} \cdot I_2$$

The only difference in the parameter sets is the choice of independent and dependent variables. The parameters are the constants used to relate these variables. However, in high frequency systems, voltage, current and power can be considered to be in the form of waves traveling in both directions along a transmission line. A portion of

the waves incident on the load will be reflected. It then becomes incident on the source, and in turn re-reflects from the source (if $Z_s \neq Z_0$), resulting in a standing wave on the line. Under this conditions, the value of the total voltage at any given point along the length of a transmission line is the sum of the incident and reflected waves at that point;

$$V_t = E_{inc} + E_{refl} \quad (4)$$

The total current on the line is the difference between the incident and reflected voltage waves divided by the characteristic impedance of the line;

$$I_t = \frac{E_{inc} - E_{refl}}{Z_0} \quad (5)$$

SCATTERING PARAMETERS

By using these traveling waves as our variables, a new high frequency 2-Port Network can be implemented, as in Fig. A.4;

Fig. A.4 Traveling Waves in a two-port network

By looking at E_{r2} , we see that it is made up of that portion of E_{i2} reflected from the output port of the network as well as that portion of E_{i1} that is transmitted through the network. Each of the other waves are similarly made up of a combination of two waves.

Let us now find a parameter set that relates these four traveling waves. This set will be named S-Parameter or Scattering Parameter set. We will use the H-Parameter set to derive our new relationship. While the derivation will be made for two-port networks, it is applicable for n-ports as well. Recalling that the H-parameters set are:

$$V_1 = h_{11} \cdot I_1 + h_{12} \cdot V_2$$

(6)

$$I_2 = h_{21} \cdot I_1 + h_{22} \cdot V_2$$

Let us now find the total voltages and currents in Fig. A.4, by means of their corresponding traveling waves:

$$V_1 = E_{i1} + E_r \qquad V_2 = E_{i2} + E_{r2} \qquad (7)$$

$$I_1 = \frac{E_{i1} - E_{r1}}{Z_0} \qquad I_2 = \frac{E_{i2} - E_{r2}}{Z_0} \qquad (8)$$

By substituting these expressions for total voltage and total current into The H-parameter set, and arranging the equations such that the incident traveling voltage waves are the independent variables; and the reflected traveling voltage waves are the dependent variables, we get;

$$E_{r1} = f_{11}(h) \cdot E_{i1} + f_{12}(h) \cdot E_{i2} \qquad (9)$$

$$E_{r2} = f_{21}(h) \cdot E_{i1} + f_{22}(h) \cdot E_{i2}$$

The functions f_{11} , f_{21} and f_{12} , f_{22} represent a new set of network parameters, relating traveling voltage waves rather than total voltages

and total currents. Again, these functions were obtained by using the H-Parameters, but they could have been derived from any other parameter set. Now, if we divide both sides of these equations by $\sqrt{Z_0}$, the relationship will not change. But we will obtain new variables, these are:

$$a_1 = \frac{E_{i1}}{\sqrt{Z_0}} \quad a_2 = \frac{E_{i2}}{\sqrt{Z_0}} \quad (10)$$

$$b_1 = \frac{E_{r1}}{\sqrt{Z_0}} \quad b_2 = \frac{E_{r2}}{\sqrt{Z_0}} \quad (11)$$

As the square of these new variables has the dimension of power, then ;

$$|a_1|^2 = \text{Incident power on port 1}$$

$$|b_1|^2 = \text{Reflected power from port 1}$$

The same meaning will have the variables of port 2. And these new waves can be called traveling power waves rather than traveling voltage waves. Finally, we present the S-Parameter set of equations:

$$b_1 = S_{11} \cdot a_1 + S_{12} \cdot a_2$$

(12)

$$b_2 = S_{21} \cdot a_1 + S_{22} \cdot a_2$$

and Fig. A.5 shows how they relate in a 2-Port network:

Fig. 3 Dependent and independent variables
when using the S-parameters

TRANSDUCER POWER GAIN

The transducer power gain is defined as the power delivered to the load divided by the power available from the source.

$$G_T = \frac{P_{del}}{P_{avs}} = \frac{|S_{21}|^2 \cdot (1 - |\Gamma_S|^2) \cdot (1 - |\Gamma_L|^2)}{|(1 - S_{11} \cdot \Gamma_S) \cdot (1 - S_{22} \cdot \Gamma_L) - S_{21} \cdot S_{12} \cdot \Gamma_L \cdot \Gamma_S|^2} \quad (13)$$

Let us consider the network to be unilateral ($S_{21} \approx 0$), the previous expression is simplified to:

$$G_{TU} = \frac{(1 - |\Gamma_S|^2)}{|1 - S_{11} \cdot \Gamma_S|^2} \cdot |S_{21}|^2 \cdot \frac{(1 - |\Gamma_L|^2)}{|1 - S_{22} \cdot \Gamma_L|^2} \quad (14)$$

Where the mid term is related to the transistor used. This term will remain as a constant as long as the transistor and its bias condition are not changed. The other two terms, however, are not only related to the remaining S-parameters of the two-port device, S_{11} and S_{22} , but also to the source and load reflection coefficients. By means of these two quantities we will be able to control the gain in the design of the amplifier.

Because of all this, the unilateral transducer power gain is made up of three distinct and independent terms. These are:

$$G_{TU} = G_S \cdot G_O \cdot G_L \quad (15)$$

or

$$G_{TU}(\text{dB}) = G_S(\text{dB}) + G_O(\text{dB}) + G_L(\text{dB}) \quad (16)$$

where:

$$G_S = \frac{(1 - |\Gamma_S|^2)}{|1 - S_{11} \cdot \Gamma_S|^2} \quad (17)$$

$$G_O = |S_{21}|^2 \quad (18)$$

$$G_L = \frac{(1 - |\Gamma_L|^2)}{|1 - S_{22} \cdot \Gamma_L|^2} \quad (19)$$

The G_S term affects the degree of mismatch between the characteristic impedance of the source Z_o (usually 50 or 75 ohms) and the input reflection coefficient of the two-port device. Even though the G_S block is made up of passive components, it can have a gain contribution greater than unity. This is true because an intrinsic mismatch loss exists

between Z_0 and S_{11} , and the impedance transforming elements can be employed to improve this match, thus decreasing the mismatch loss, and in this sense, can be thought of as providing gain. The G_o term is related only to the transistor used and its bias conditions. The G_L term has the same meaning as the G_s , but applied to the output.

This expression for G_{TU} will have a maximum when the source and load reflection coefficients are chosen such that:

$$\Gamma_S = S_{11}^* \text{ and } \Gamma_L = S_{22}^* \quad (20)$$

Under these optimum conditions, we obtain:

$$G_{U \max} = \frac{1}{1 - |S_{11}|^2} \cdot |S_{21}|^2 \cdot \frac{1}{1 - |S_{22}|^2} \quad (21)$$