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FOR SUSCEPTIBILITY TESTING

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The function of the Electromagnetic Radiation System (EMRS) is to generate electromagnetic energy so as to produce a constant field strength that can be automatically scanned as a function of frequency. The design objective is to cover the frequency range of 30 hertz to 40 gigahertz with field strength intensities up to 200 volts per meter. A stripline approach is described and proposed for use as the field generating device for the

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SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered) lower frequencies. The use of defocused parabolas are proposed for use at the higher frequencies. The final system configuration is described including specific equipment to be used along with interface and control circuitry necessary for proper operation.

TABLE OF CONTENTS

I. INTRODUCTION

A. The EMRS Design Plan Phase I Study of May 1977, outlined Phase II, the Hardware Phase, of the EMRS program. The Hardware Phase objective was to fabricate and test a representative model of the EMRS proposed in Phase I.

The Hardware phase was to include the detailed specification of the system configuration, selection and procurement of system components, and the design and development of any special antenna systems, interface circuits and control circuits.

After procurement and/or development of all components, test would be accomplished to verify that system performance and features satisfy the design goals and philosophy.

B. The basic EMRS design goal for the prototype hardware was to develop a system which could produce a 200 V/m field intensity using both stripline and refocused parabola antenna configurations.

Other goals included demonstrating the ability to program power output and frequency so that the systems could be used to test equipment to a tailored set of specifications. Flexibility to allow for future expansion of the EMRS was also a design goal. Future expansion might include: fully automatic computer control of field intensity, frequency, and sweep mode; expanded frequency coverage to extend up to 40 GHz; filling in of those frequency bands not covered in the prototype system.

C. With the above goals in mind, during the period of this report, the following has been accomplished:

1. The sweep oscillator system has been completely specified and ordered. Delivery is expected in February 1978.

2. Power amplifiers for all the demonstration frequency bands have been specified and ordered. The amplifier for the 30-60 MHz band has been

received and is undergoing testing. Delivery of the remaining three amplifiers is expected March 3, 1978.

3. Leveling loop and programmable attenuator components have been ordered and delivery is expected in March 1978.

4. Pulse modulation components have been ordered and are expected in March 1978.

5. The equipment rack for mounting EMRS rack mountable components has been ordered. The rack includes an integral cooling blower assembly.

6. Detailed tracking filter specifications have been completed and request for quotations from several filter manufacturers has been made.

7. Fabrication and testing of the stripline and refocused parabola antenna have been completed.

II. DEMONSTRATION EMRS HARDWARE.

A. Figure 1 shows a block diagram of the demonstration EMRS. Figure 1(a) shows the system for the frequency bands 30 to 60 MHz and 1 to 2 GHz. Figure 1(b) shows the system for the frequency bands 2 to 4 GHz and 12.4 to 18 GHz.

The system of Figure 1(a) consists of a sweep oscillator signal source (1), programmable attenuator (2), pulse modulator (3), power amplifier (4), tracking filter (5), antenna subsystem consisting of a power splitter (6), four section stripline (7), 50 ohm loads (8), 50 ohm attenuator (9), crystal detector (10), and external leveling filter/amplifier (11).

The system of Figure 1(b) is the same as Figure 1(a) except for the antenna subsystem which consists of a refocused parabola (6) and field probe (7).

B. Subsystem Functions.

1. The sweep oscillator (1) is the RF signal source for the demonstration EMRS. The signal source can be operated in a cw mode or several sweep

modes. Control is provided manually or by remote digital programming. The remote programming mode allows for computer control of oscillator frequency, frequency band and sweep mode.

2. The programmable attenuators (2) provide for remote digital control of the RF signal power. The attenuation is variable from 0 to 81 db in 1 db steps. Attenuation control signals are provided by eight parallel input lines.

3. The AM/pulse modulator (3) provides for external pulse modulation and fast amplitude modulation. The modulator is a PIN diode which is driven by a high current diode driver. A modulating frequency source (not part of EMRS) provides the modulating frequency signal.

4. The Power Amplifier (4) provides power amplification to obtain signal levels sufficiently high to generate the 200 V/m field intensity required by EMRS. For the frequency band 30 to 60 MHz, a solid state 100 watt amplifier is used. For frequencies above 1 GHz, traveling wave tube amplifiers are used.

5. The Tracking Filter (5) provides tunable narrow band filtering of the amplifier output signal. The filter is designed to track the center frequency of the sweep oscillator in either cw or sweep modes. The filter is mechanically tuned using a servo positioning unit. The sweep output ramp voltage is used as the filter driving signal.

6. The antenna subsystem (6), (7), (8), (9) of Figure 1(a), and (6) and (7) of Figure 1(b) transform the RF power from the tracking filter output into the field intensity required to meet the design goals of EMRS. Figure 1(a) shows a 4-section stripline with associated power splitter 50Ω loads and 50Ω attenuator. The equipment to be radiated forms a ground plane for the stripline and is subjected to E fileds normal to its surface. A detailed description of the stripline is given in Section III. F.

The refocused parabola shown in Figure 1(b) is designed to radiate a unit under test at approximately three meters distance. A field probe (7) monitors the RF field intensity.

7. RF power leveling circuitry consists of a crystal detector which produces a DC output voltage proportional to the RF power input. The DC level is amplified and filtered by the 1 Hz filter/amplifier. The amplifier gain and offset is programmable so that output power levels can be intentionally changed without affecting the leveling operation.

III. SYSTEM COMPONENTS.

A. RF Sweep Oscillator Subsystem-

The EMRS Design Plan dated May 1977, established these basic performance criteria for the sweep oscillator subsystem:

1. Four frequency bands covered:

   30 to 60 MHz

   1 to 2 GHz

   2 to 4 GHz

   12.4 to 18 GHz

2. Automatic and manual scanning of frequencies within each of the above bands.

3. Output power variable over a 53 db dynamic range.

4. Provision for computer control of all important sweep oscillator functions including frequency, sweep mode, and output power level.

5. Modulation capabilities to include:

   Amplitude modulation - 0 to 50% with 100 Hz to 10 MHz modulating frequencies.

   Frequency modulation - 0 to 10 MHz deviation with 100 Hz to 10 MHz modulating frequencies.

Pulse modulation - pulse width of 0.05 usec to 5 millisec. at pulse rates of 100 Hz to 10 MHz.

In addition to these basic requirements, the compatibility of the sweep oscillator with other system components must be considered with respect to overall system cost. For example, higher output power from the sweep oscillator could eliminate the need for and the expense of a preamplifier. Frequency accuracy and sweep output linearity are also important considerations. The EMRS system signal purity specifications require that all harmonics be 100 db below the power level of the center frequency. To achieve this, a narrowband tunable tracking filter is required. The signal that drives the tunable filter is obtained from the sweep oscillator sweep output. Hence, the correlation between the oscillator frequency and the sweep output voltage must be high so that the tracking filter passband is always centered on the fundamental frequency of the oscillator.

Other considerations included the flexibility of the sweep oscillator to adapt to future changes in EMRS requirement. Features, such as plug-in RF units, full compatibility with the IEEE-488/75 standard for computer control, and frequency coverage above that required in the EMRS Design Plan. The sweep oscillator selected for the demonstration EMRS is the Hewlett-Packard 8629C system. The selection of the HP sweep oscillator over similar equipment from other manufacturers was based on several technical factors mentioned above in addition to economic considerations.

In the 4th quarterly report*, it was mentioned that the main disadvantage of the HP8620C sweep oscillator system was that the frequency tolerance in the 30 to 60 MHz band was $\pm$ 20 MHz. This tolerance was obtained with the HP86320B RF module. Since the time of that report, several developments have

* ECOM-76-0332-4, "Electromagnetic Radiation System (EMRS) for Susceptibility Testing", April 1978.

occurred. Further research determined that if the HP86222B RF plug-in were used, frequency accuracy in the 30 to 60 MHz range could be improved to ± 200 kHz. This improvement is obtained using the crystal markers built into the 86222B.

Another factor favored using the HP 8620C system. Hewlett-Packard has just introduced a new series of RF plug-ins for the 8620C. The new plug-ins have much higher power output than any similar sweep oscillators. In particular, for the 2-4 GHz band, the 86235A RF plug-in output power is 40 milliwatts. The advantage of this is that a preamplifier would no longer be required as it was with the Systron-Donner system. In addition, the HP system is superior in frequency resolution when using digital frequency programming. The HP system can select 10,000 different frequencies across any band while the Systron Donner has only 1000. Another important factor was the greater FM modulation capabilities of the HP system. Details of this were given in the comparison table in the previous report.

B. Power Amplifier Subsystem.

1. 30 to 60 MHz Band.

The Electronic Navigation Industries, Inc. model 3100L was chosen for this frequency range. The 100 watt power bandwidth of the 3100L is 250 kHz to 105 MHz. 50 watts is available from 120 kHz to 120 MHz. The power gain is nominally 50 db.

The primary technical advantage of the ENI 3100L compared with similar equipment from other manufacturers is the low distortion at rated power and the very low noise figure. These are important in obtaining high signal purity at the antenna.

2. 1 to 2 GHz, 2 to 4 GHz and 12.4 to 18 GHz Bands.

For these bands, amplifiers manufactured by Logimetrics were chosen. They are:

Frequency	Model	Important Features		
		Gain	Power Out	VSWR Protection
1 - 2 GHz	A600L	40 db	200 W	Yes
2 - 4 GHz	A600S	43 db	200 W	Yes
12.4 - 18 GHz	A200U	30 db	10 W	Yes

The Logimetric amplifiers were chosen because the high gain, high output power and VSWR protection were not available from other manufacturers.

C. Programmable Power Level Circuitry.

To remotely program the field intensity generated by the stripline or refocused parabola, a two-stage programmable attenuator is placed in series with the output of the sweep oscillator. One stage consists of an attenuator variable from 0 to 70 db in 10 db steps. This is followed by a 0 to 11 db attenuator variable in 1 db steps. Control of the attenuators is via the IEEE-488 Bus. Manual control is also possible using toggle or rotary selector switches.

The programmable attenuators are the HP-8495H and the HP-8494H or equivalent.

D. Fast AM/Pulse Modulation Circuitry (Ref. Figure 2).

The fast AM/pulse modulation circuit consists of a PIN diode switch and switch driver. The PIN diode switch is an HP-33102A or equivalent.

For pulse modulation, the switch driver is a National semi-conductor DH0035 PIN diode driver. The DH0035 takes a TTL input pulse train and converts it to a high current output for driving the PIN diode. The turn-on and turn-off times for this configuration are 50 nanoseconds each. This allows a PRF greater than 1 MHz.

For fast AM modulation, the DH0035 is driven by a linear high current driver to be designed by AEL.

In the pulse mode of operation, the PIN diode acts as a short or open circuit in the transmission line. When heavily forward biased, the RF resistance of the diode is very low and a short circuit appears across the line. This relfects essentially all of the incident RF power back to the source. When unbiased or reversed biased, the diode appears as a very small capacitive reactance and allows the incident power to pass. Typically, a 40 db on/off ratio is obtained with a broadband PIN diode switch.

In the last AM mode of operation, the PIN diode would be biased at the midpoint of its dynamic range. This is accomplished by applying the proper DC bias level to the DH0035 driver. A modulating signal would then be applied to the driver and would swing above and below the quiescent level. The RF resistance of the PIN diode would vary approximately linearly with the changing bias signal.

E. Tracking Filter.

Requests for bids to manufacture the tunable tracking filters have been formally made to several vendors. When specifying the filters, the primary considerations were:

1. Power handling capability.
2. Tuning range.
3. Tracking accuracy.
4. Insertion loss.
5. Roll-off rate.
6. Tuning rate.

A detailed description and calculations are given in Appendix A. An example of the specifications provided to vendors is also shown.

Each filter will be mounted in the equipment rack next to its corresponding amplifier. The input connector will be on the front panel and the output connector will be mounted at the rear of the filter enclosure. This will put the filtered output near the screen room wall and allow for short cable lengths. Filter procurement should be completed during the next reporting period.

F. Antenna Subsystem.

Progress on the stripline and refocused parabola has proceeded as follows:

1. Stripline. Tests were performed to determine the relationship between RF susceptibility testing for a radiated field and a stripline generated field. Testing was completed and the final configuration has been determined. The significant results of this effort are the following:

a. An equivalence between RF susceptibility measurements using a field generating system where the electric fields are parallel to the surface of the unit-under-test (UUT) and where these fields are perpendicular to the surface of the UUT does exist.

b. Apertures in the surface of the UUT and protrusions into the stripline result in only slight changes in the impedance of the stripline.

c. Detailed calculation for the stripline are given in Appendix B.

A deliverable stripline package has been fabricated and will be delivered to the AEL Monmouth facility during the week of January 3, 1978. The power requirements as a function of field intensity to be generated has been calculated for the stripline and is shown in Figure 3. It should be noted that this is not frequency dependent if it is assumed the VSWR is the same at all frequencies.

2. Refocused parabola. Fabrication and testing has been completed on the parabolic reflectors for both the 2-4 GHz band and the 12.4-18 GHz band. Field probes for these bands have also been fabricated and tested.

Radiation patterns and gains for each reflector have been determined. The equation relating the antenna power input to the field strength produced at a distance R from the antenna is

$$P = \frac{E^2 R^2}{30 G_T}$$

where: E is the field intensity to be generated

R is the distance from the antenna to the point where the given field intensity is to exist.

G_T is the transmit antenna gain in power ratio

A plot of the power requirements as a function of frequency and field intensity to be generated is shown in Figures 4 and 5.

IV. SUMMARY

A. Action completed during this reporting period:

1. Final configuration of Demonstration EMRS was determined.

2. Sweep Oscillator RF power amplifiers, programmable attenuators, crystal detector, and equipment rack have been ordered.

3. The 30-60 MHz power amplifier has been received.

4. Request for bids to manufacture the tunable band pass filters have been made to several manufacturers.

5. Antenna fabrication and testing has been completed.

B. Action anticipated during next reporting period:

1. Receipt and testing of all ordered equipment.

2. Tunable band pass filter procurement.

3. Procurement of all remaining hardware.

4. Design and fabrication of control and interface circuitry.

5. Assembly of Demonstration EMRS.

6. Begin system testing.

SECTION V

APPENDIX A

TRACKING FILTER CONSIDERATIONS

1. The basic requirements of the tunable tracking filter are:

a. The filter pass band must be centered about the sweep oscillator frequency for all sweep modes.

b. The filter bandwidth must be wide enough to pass frequency components due to carrier modulation.

c. The stop band attenuation of harmonic and non-harmonic components must be sufficient to produce the required signal purity.

d. The power handling capabilities must be high enough to handle the required power levels.

2. To meet these requirements, the filter should be designed to the attached specification which is for the tracking filter in the 2 to 4 GHz band. The following calculations were used in arriving at the numbers in the specification.

3. Filter bandwidth. The highest modulating frequency anticipated in the demonstration EMRS is $\pm$ 10 MHz. For the 2 to 4 GHz band, this represents a required bandwidth of $\pm$ 0.5%.

At 2 GHz, the HP86235A has a frequency accuracy of $\pm$ 30 MHz or $\pm$ 1.5%. Frequency linearity over an octave band is $<\pm$ 0.5%.

Adding all of the above:

Modulation Bandwidth	$\pm$ 0.5%
Frequency Accuracy	$\pm$ 1.5%
Frequency Linearity	$\pm$ 0.5%
Total bandwidth required	$\pm$ 2.5%

Therefore, the bandwidth of the tracking filter should be $>\pm$ 2.5%. If $\pm$ 0.5%

is added for variations due to temperature, power variation, and RF power changes, then the required bandwidth is ± 3.0%.

Therefore, the filter bandwidth must be at least ± 3.0% plus the tracking accuracy of the filter; i.e.

Filter bandwidth = ± Filter tracking accuracy ± 3.0%.

4. Stop band attenuation. The 2nd, 3rd, and all higher order harmonics must be 100 db below the fundamental frequency. At rated power output, the 2nd harmonic is down by 6 db and the third harmonic is down by 17 db. Therefore, the filter attenuation requirements are:

2nd harmonic	94 db attenuation
3rd harmonic and higher	83 db attenuation

5. To achieve the above attenuation, the bandpass filter must have a roll-off rate greater than 94 db per octave.

APPLICATION		REVISIONS			
NEXT ASS'Y	USED ON	REV.	DESCRIPTION	DATE	APPROVED

1. RF Power Input: 200 watts at the center frequency.
   32 watts at the 2nd harmonic
   4 watts at the 3rd harmonic

2. Center frequency, variable from 2.1 GHz to 4 GHz.

3. Passband: 0.5 db bandwidth shall be greater than or equal to the following percent of the center frequency: $\pm$ filter tracking accuracy $\pm$ 3%.*

4. Passband attenuation, maximum: $\leq$ 0.5 db

5. Passband ripple, maximum: $\leq \pm$ 0.5 db

6. Passband INPUT SWR, maximum: $\leq$ 1.5:1

7. ROLL OFF RATE: 94 db/octave

8. Input and output impedance: 50 Ω nominally

9. Even harmonic attenuation: $\geq$ 94 db below the center frequency.

10. Odd harmonic attenuation: $\geq$ 83 db below the center frequency.

11. Filter shall work into a load SWR of $\leq$ 2.1:1

12. Filter Tuning: A servo positioning unit or its equivalent provided as an integral part of the filter to have the following characteristics:

    a. Tuning rate: variable from zero to a maximum of 30 sec/octave.

    b. Tuning imput: linearly variable from 0 to 10 VDC
       0 volts corresponding to a center frequency of 2.1 GHz $\pm$ A%.*
       10 volts corresponding to a center frequency of 4 GHz $\pm$ A%.

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AMERICAN ELECTRONIC LABORATORIES, INC.
COLMAR, PA.

SPECIFICATION CONTROL DRAWING
TUNABLE BAND PASS FILTER FOR THE FREQUENCY BAND 2.1 GHz TO 4 GHZ.

DWG SIZE	CODE IDENT	DWG NO.	REV.
A	93346	1690212	

SCALE | SHEET 1 OF 2 SHEETS

FORM NO. AAA

c. Tuning signal input impedance: $\geq 500 k\Omega$

d. Linearity (correlation between center frequency and the corresponding input voltage), $\pm$ B%.*

e. Tuning signal input connector: BNC, front panel mounted.

f. Tuning range: fc = 2.1 GHz to 4 GHz.

13. Input connector: Type N (v), front panel mounted.

Output connector: Type N (f) rear panel mounted.

14. Filter to be rack mountable in standard 19" rack.

15. Primary power: 115 VAC, 50 to 60 Hz.

* The tracking filter accuracy is the sum of A% and B% above. It is to be specified by the manufacturer such that the filter meets all other specifications.

SECTION V

APPENDIX B.

STRIPLINE CONSIDERATIONS

The effort during this reporting period has focused on determining the relationship between RF susceptibility testing for a radiated field and a stripline generated field. The significant difference between the two methods is the orientation of the electric fields relative to the test equipment surface. For a radiated field, the electric field lines are parallel to the surface of the equipment under test (EUT), whereas with the stripline technique, the electric field lines are perpendicular to, and terminate on, the surface of the EUT.

A theoretical analysis of this relationship was performed during the first phase of the program (see Appendix C), and the results are as follows:

1. The surface directed magnitude field produced by the stripline (H_b) is:

$$H_b = \frac{2\pi}{5} S_b \text{ (Oersteds)}$$

where S is the surface current density.

2. The surface directed magnetic field produced by the radiated field (H_i) is:

$$H_i = \frac{\pi}{5} S_i$$

Since the electric field is equal to the magnetic field in CGS units,

$$E_b = \frac{2\pi}{5} S_b$$

$$E_i = \frac{\pi}{5} S_i$$

If $E_i = E_b$, that is, the radiated field strength incident upon the EUT is equal to the field produced by the stripline perpendicular to the EUT, the relationship between the surface current densities produced is:

$$\frac{2\pi}{5} S_b = \frac{\pi}{5} S_i$$

and therefore, the ratio of the current densities induced on the surface of the EUT is:

$$\frac{S_b}{S_i} = \frac{1}{2}$$

The conclusion, therefore, is that the radiated field will induce twice the current density on the surface of the EUT than will be induced by the stripline technique. Since these currents are the mechanism by which energy is coupled into the equipment whose susceptibility is being measured, the implication is that to test a piece of equipment for its susceptibility to a radiated electric field of X volts per meter, the equivalent susceptibility test using the stripline technique would require generating a field strength 2X volts per meter.

The tests performed during this reporting period were designed to verify the theory. The nature of the test was as follows:

A box (19 x 15 x 6 inches) was constructed to be used to simulate a piece of communications equipment. A 6 x 0.3 inch slot was cut in the box. The relationship was to be tested from 1.0 to 2.0 GHz. Because the box was resonant at 1.75 GHz (which hampered the measurement) a layer of microwave absorber was added to the cavity to eliminate the resonance.

The gain of the slot in the box was measured from 1.0 to 2.0 GHz in 250 MHz increments. With this information, the power which would be received by this slot, if a 200 V/m radiated field were incident upon it, was calculated using the following equation:

$$P_R = \frac{E^2 G_R \lambda^2}{480 \pi^2}$$

where: E is the field strength (V/M)

G_R is the gain of the slot (power ratio)

λ is wavelength (meters)

The next step was to determine the power received by the slot if a 200 V/m field is generated using the stripline technique. This is determined in the following manner:

1. The box is placed over a hole cut in one of the ground planes of the stripline.
2. The power being fed into the stripline is measured.
3. The power being received by the slot is measured.
4. Knowing that the impedance of the line is fifty (50) ohms, the power required to produce a 200 V/m field between the center conductor and the test box surface can be determined using the following equation:

$$P = \frac{(E\ d)^2}{Z}$$

where:

E = field intensity (V/M)

d = spacing between center conductor and test box surface (meters)

Z = characteristic impedance of the stripline

$$P = \frac{(200)^2\ (.0381)^2}{50} \quad = 1.16 \text{ watts}$$

Because the input power divides equally to produce equivalent field intensities on each side of the center conductor, the power required is 2 (1.16) = 2.32 watts.

5. Taking the ratio of receive power to input power (determined above) and multiplying by 2.32 watts, the power received by the slot in the presence of a 200 V/m field can be determined for the stripline technique.

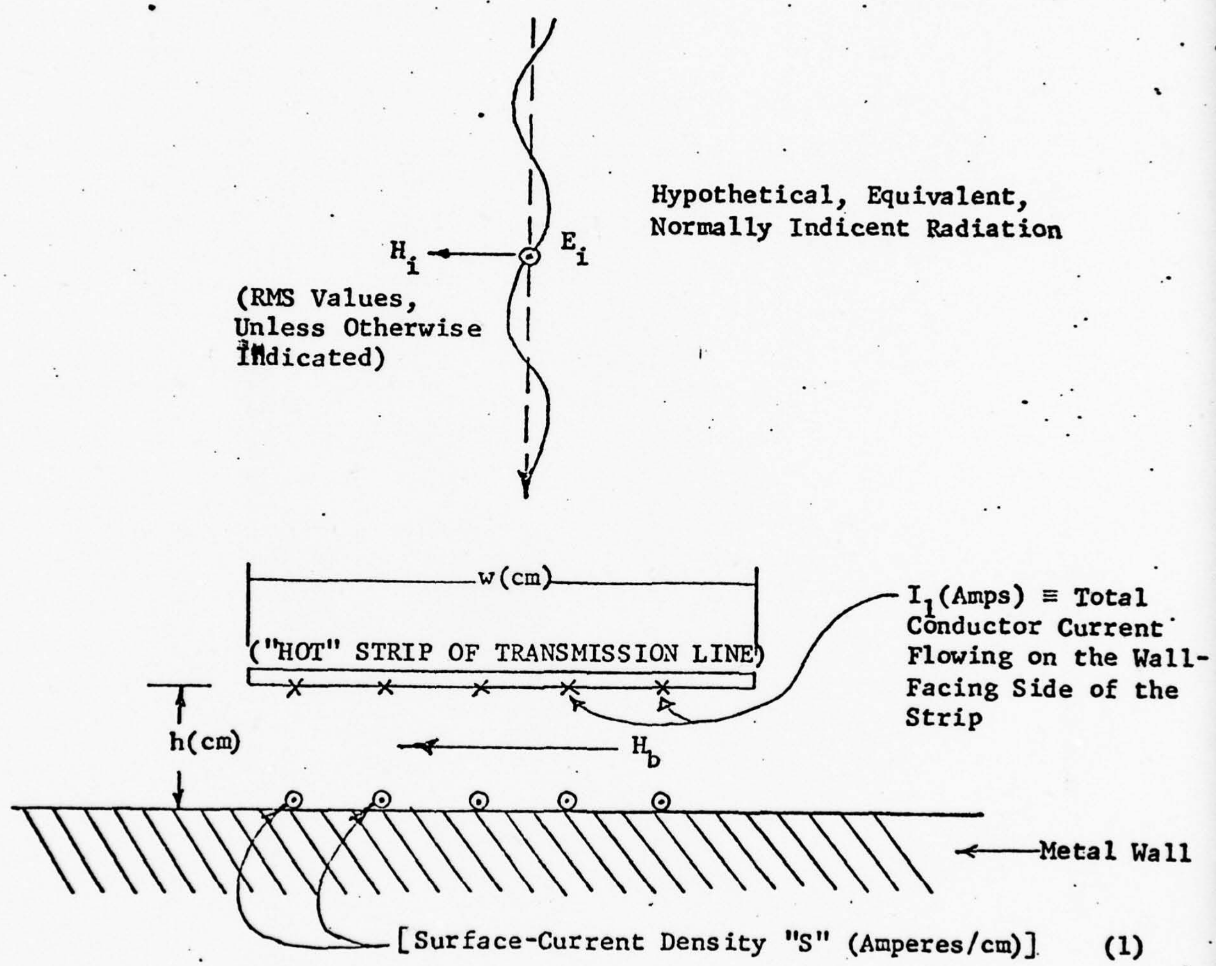
The ratio of the receive powers for the two techniques should be 4:1 which is based upon the theoretical analysis of Appendix C(square of the current density ratio).

The problem in determining this relationship is that small errors in the measurements will result in large differences in the apparent ratio. This was the problem encountered when the experimental verification was attempted. Typically, the measurements, for both the radiated and stripline technique, were not repeatable to greater than $\pm$ 0.5 db. However, from the data measured, the ratio appears to be on the order of 1.5.

The fact that the apparent ratio is 1.5 rather than 4 is not significant at this time. The exact equivalence between the two methods is not of primary importance in this phase of the program. It is more significant that a relationship between the two techniques does exist and that the net effect of susceptibility testing using the stripline technique is not significantly different from susceptibility testing using radiated techniques. Therefore, susceptibility testing using the proposed stripline technique is valid.

APPENDIX C

CALCULATION OF THE EQUIVALENT NORMALLY INCIDENT (FREE SPACE) RADIATIVE AMPLITUDE AND INTENSITY, FOR SURFACE-CURRENTS INDUCED IN A METAL WALL BY A PARALLEL-RUNNING TRANSMISSION-LINE STRIP



(1)

We assume h << w for convenient approximation. (2)

The main result to be obtained (Eq. (12)) does not depend upon the dielectric constant of nonmagnetic material between the strip and the wall, nor upon whether an outside shield-strip is used with the line. Only the drive power is affected by these factors.

Because of (2), the current-density induced on the metal wall will be about equal to that on the wall-facing side of the "hot" strip:

$$S = \left(\frac{I_1}{w} \right) \quad , \quad \text{(Amps./cm, out of page).} \tag{3}$$

The surface-directed magnetic field, H_b , is given by:

$$H_b = \left(\frac{4\pi}{10} \cdot \frac{I_1}{w} \right) \quad , \quad \text{(Oersteds),} \tag{4}$$

The equivalent amplitude and intensity of interest can be expressed either in terms of I_1, or S. Eliminating I_1 between (3) and (4) yields:

$$H_b = \left(\frac{4\pi}{10} \right) S \quad \text{(Oersteds)} \tag{5}$$

The incident free-space-wave magnetic field H_i that would be needed, to produce the same boundary [magnetic field H_b upon full reflection of the wave (and, hence, the same surface-current-density)] is:

$$H_i = \left(\frac{H_b}{2} \right) \tag{6}$$

Eliminate H_b between (6) and (4), which gives:

$$H_i = \left(\frac{\pi}{5} \frac{I_1}{w} \right) \quad , \quad \text{(Oersteds).} \tag{7)(a}$$

Since $E_i = H_i$ in c.g.s. units,

$$E_i = \left(\frac{\pi}{5} \frac{I_1}{w} \right) \quad , \quad \text{("stat Volts/cm"); or:} \tag{7)(b}$$

$$E_i \doteq \left[\left(\frac{\pi}{5}\ \frac{I_1}{w}\right) \quad x\ (300)\ x\ (100)\right] \qquad , \text{(Volts/meter)}, \tag{8}$$

or:

$$E_i \approx \left(6000\ \pi\ \frac{I_1}{w}\right) \qquad , \text{(Volts/meter)}. \tag{9}$$

Conversely,

$$I_1 = \left(\frac{w\ \ E_i}{6000\ \pi}\right) \qquad , \text{(Amps)} \tag{10}$$

Specifically, in order to have:

$$E_i = 200\ \left[\text{V (RMS)/meter}\right] \tag{11}$$

(10) requires [from Equation (3)] :

$$S = \left(\frac{I_1}{w}\right) = \left(\frac{1}{30\pi}\right) \quad , \left[\text{Amps (RMS)/cm}\right] \tag{12}$$

S = 10.6 mA/cm, RMS required facing the wall

Suppose that the "hot" strip is covered by a ground-strip that is separated from it by h. Also suppose that the same dielectric-constant ε is used on both faces of the hot strip. Then the characteristic impedance of the (TEM) transmission line thus formed will be approximately:

$$Z_o \approx \left(\frac{60}{\sqrt{\varepsilon}}\ \frac{h}{w}\right) \quad , \text{(Ohms)}. \tag{13}$$

According to (1) and (12), take:

$$I_1 = \left(\frac{w}{30\pi}\right), \text{(Amps, RMS)}. \tag{14}$$

[Note that the total line-current will be $(2I_1)$.]

If the line Z_o is resistively terminated, the average power needed to excite it will be:

$$W = \left[(2I_1)^2 \cdot (Z_o)\right], \text{ or, } W = \left[\left(\frac{4}{15\pi} \cdot \frac{hw}{\sqrt{\varepsilon}}\right)\right] \quad \text{and,}$$

$$W = \left(\frac{0.085\ h_{cm}\ w_{cm}}{\sqrt{\varepsilon}}\right), \text{ Watts} \tag{15}$$

It is interesting to compare the power density in the line with the power-density of the equivalent incoming wave. The cross-sectional area of the line is 2hw, hence the line's power-density P_ℓ ; from (15), is:

$$P_\ell \equiv \left(\frac{W}{2hw}\right) = \left(\frac{2}{15\pi\sqrt{\varepsilon}}\right), \left(\frac{\text{Watts}}{\text{cm}^2}\right) = \left(\frac{42.5}{\sqrt{\varepsilon}}\right), \left(\frac{\text{mW}_1}{\text{cm}^2}\right). \tag{16}$$

The average Poynting-flux P_o corresponding to an incoming wave (11) is

$$P_o = \frac{E^2}{Z_o} = \left[\frac{3}{40\pi} \quad \frac{(200)^2}{(300)^2}\right] = \left(\frac{1}{30\pi}\right), \left(\frac{\text{Watts}}{\text{cm}^2}\right) = 10.6\ \frac{\text{mW}_1}{\text{cm}^2} \tag{17}$$

$$\therefore \left(\frac{P_\ell}{P_o}\right) = \left(\frac{4}{\sqrt{\varepsilon}}\right), \tag{18}$$

NOTE: The power-density in the transmission line is just 4 times that of the free-space normally incident wave. This is as should be expected because both are TEM, and the line's H_b must be twice the incident wave's H_i, divided by $\sqrt{\varepsilon}$.

The attenuation in the dielectric, and from radiation and other losses, limits the line length that can be used in one continuous winding, hence sets a lower limit on the width w, if a specified "wall" area is to be illuminated.

SECTION VI

ILLUSTRATIONS

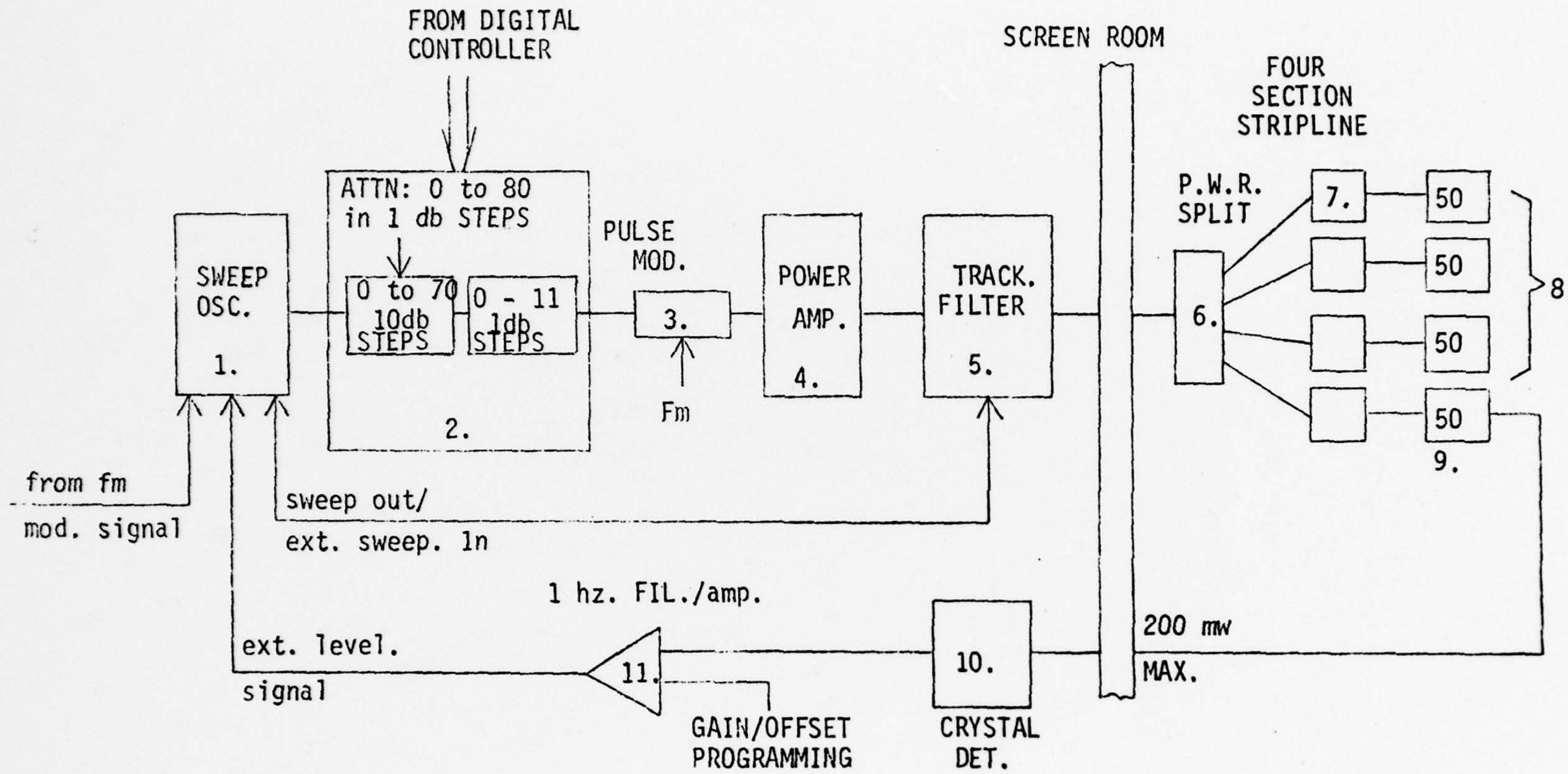


Figure 1a. Demonstration EMRS with stripline.

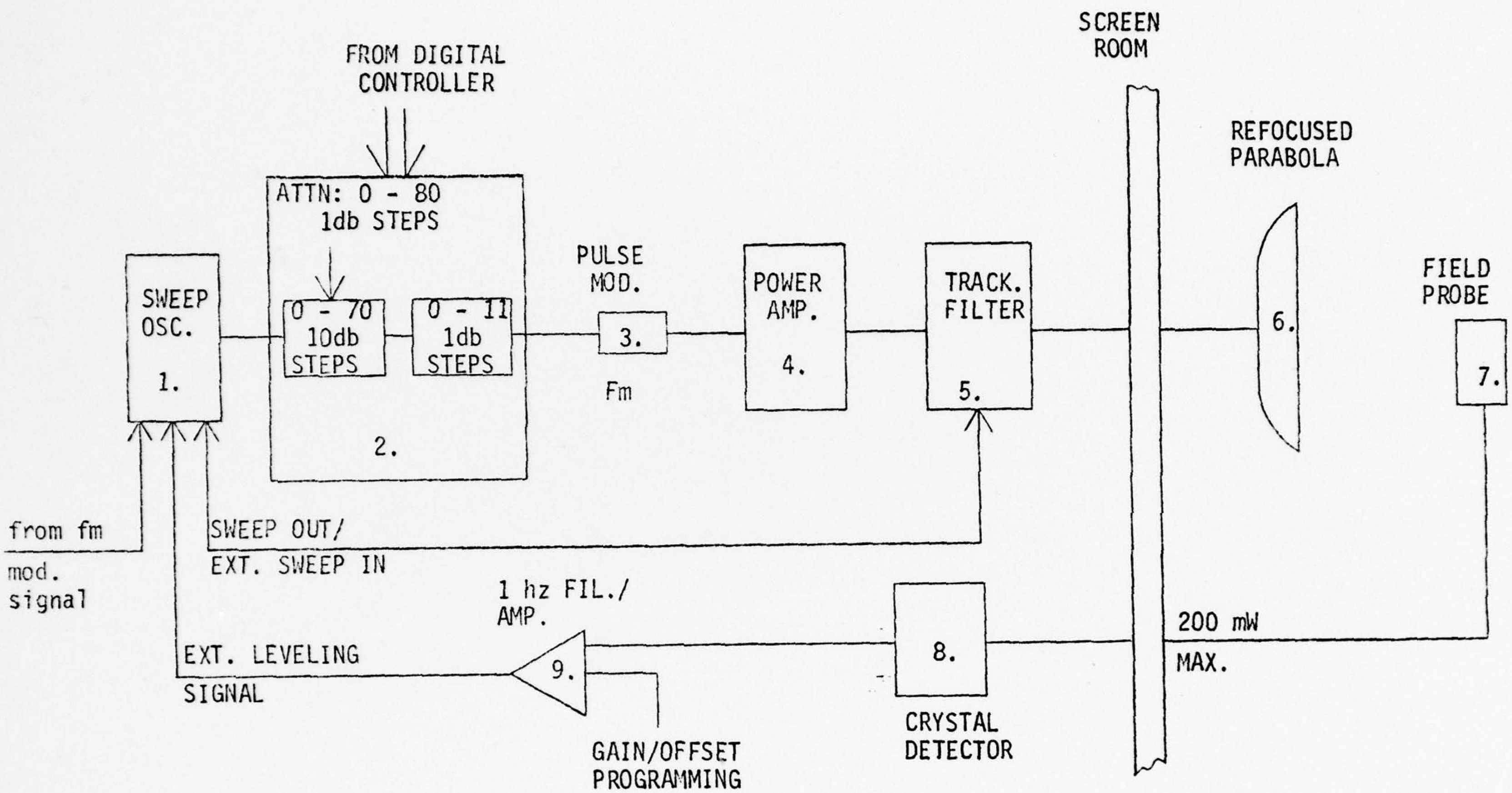


Figure 1b. Demonstration EMRS with refocused parabola.

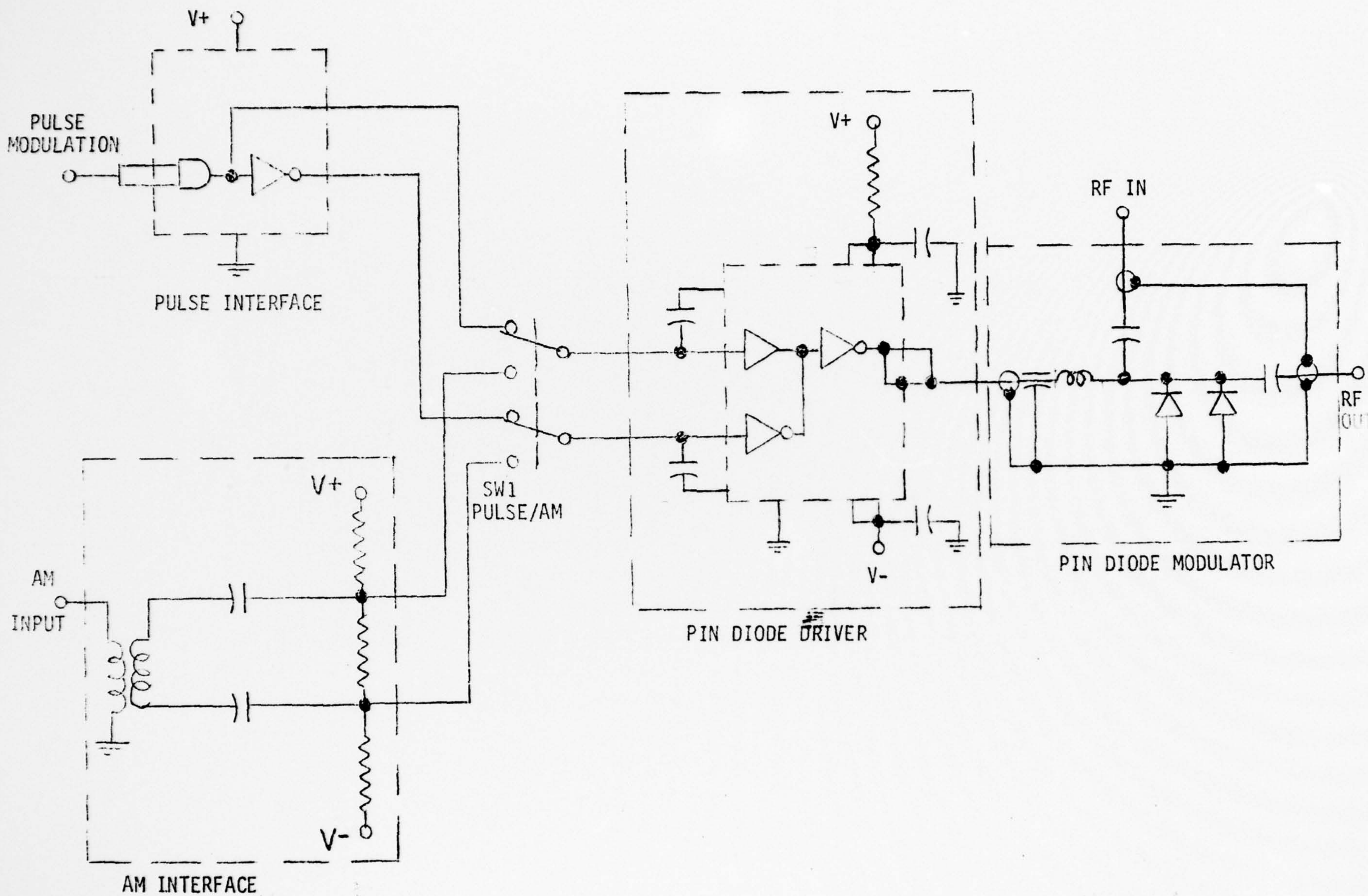


Figure 2. Fast AM/pulse modulation circuit.

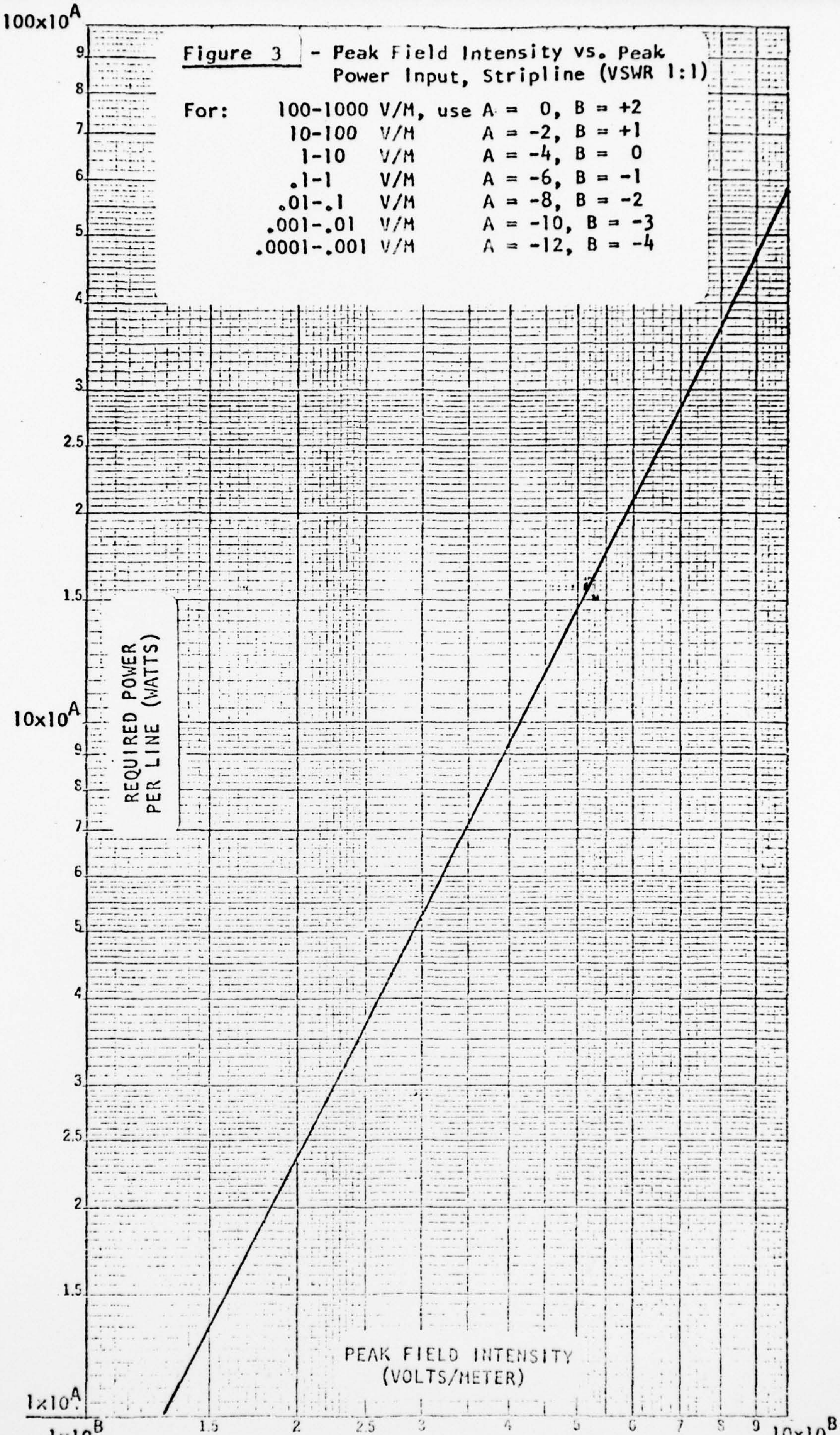


Figure 3 - Peak Field Intensity vs. Peak Power Input, Stripline (VSWR 1:1)

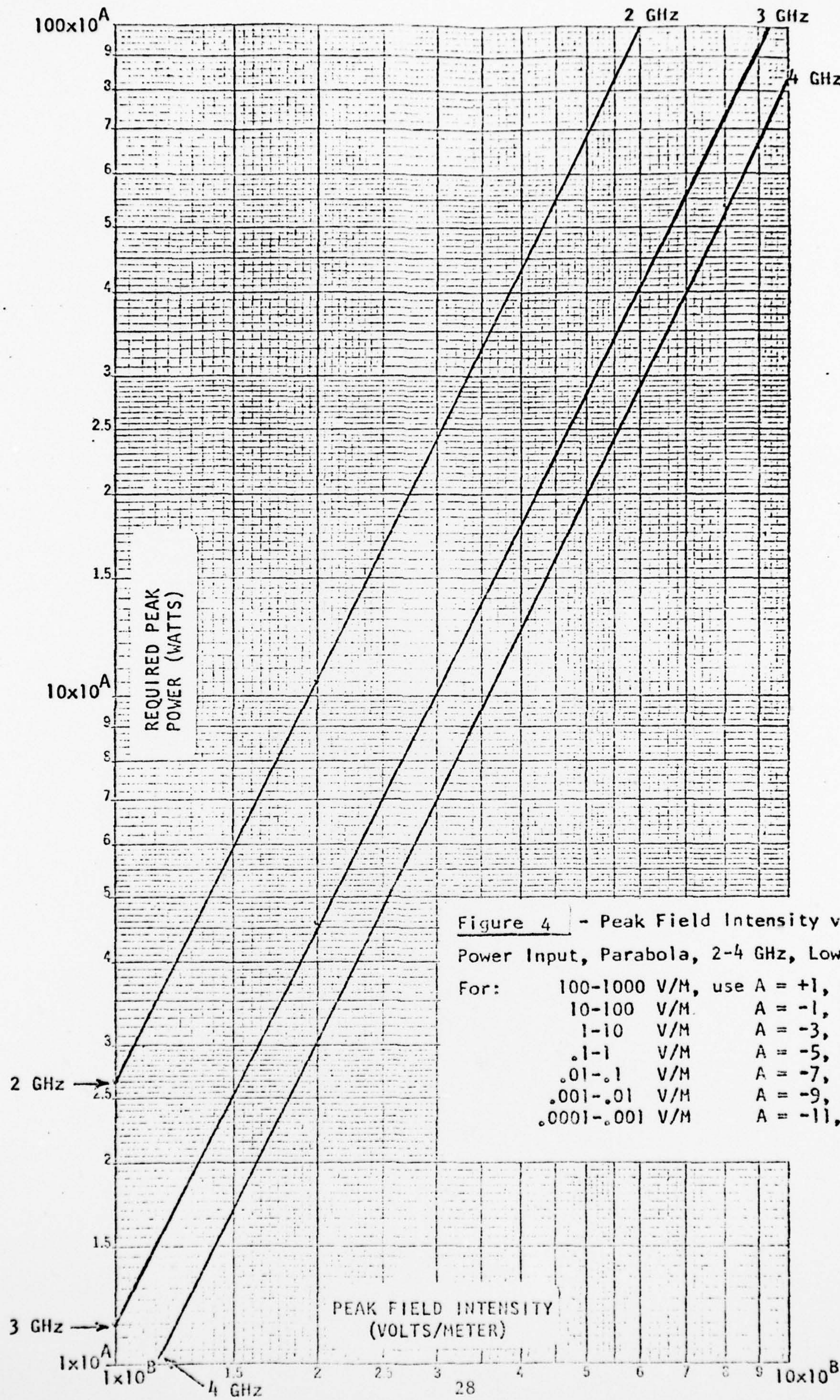


Figure 4 - Peak Field Intensity vs. Peak Power Input, Parabola, 2-4 GHz, Low Levels

For:	100-1000 V/M, use	A = +1, B = +2
	10-100 V/M	A = -1, B = +1
	1-10 V/M	A = -3, B = 0
	.1-1 V/M	A = -5, B = -1
	.01-.1 V/M	A = -7, B = -2
	.001-.01 V/M	A = -9, B = -3
	.0001-.001 V/M	A = -11, B = -4

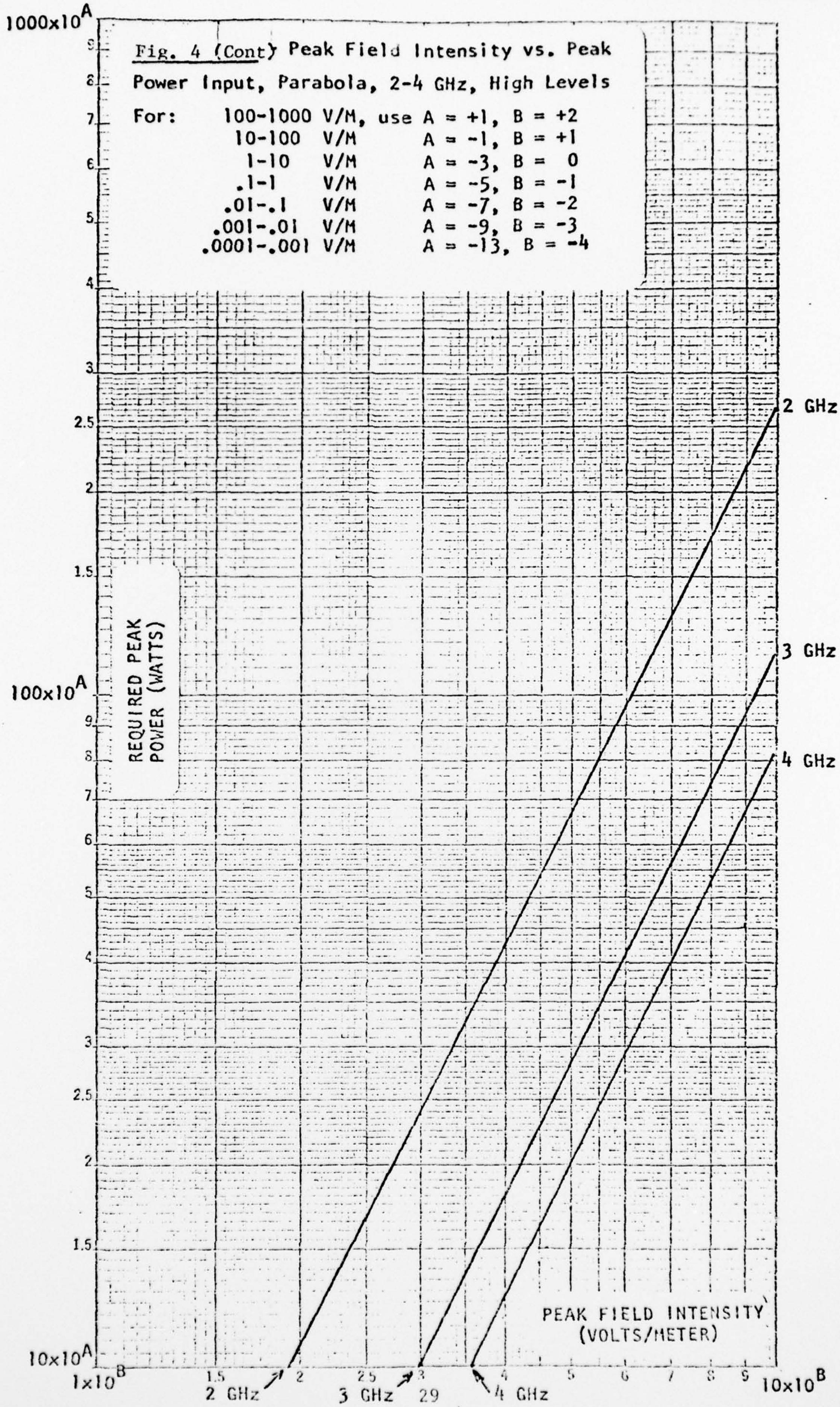


Fig. 4 (Cont) Peak Field Intensity vs. Peak Power Input, Parabola, 2-4 GHz, High Levels

For: 100-1000 V/M, use A = +1, B = +2
10-100 V/M A = -1, B = +1
1-10 V/M A = -3, B = 0
.1-1 V/M A = -5, B = -1
.01-.1 V/M A = -7, B = -2
.001-.01 V/M A = -9, B = -3
.0001-.001 V/M A = -13, B = -4

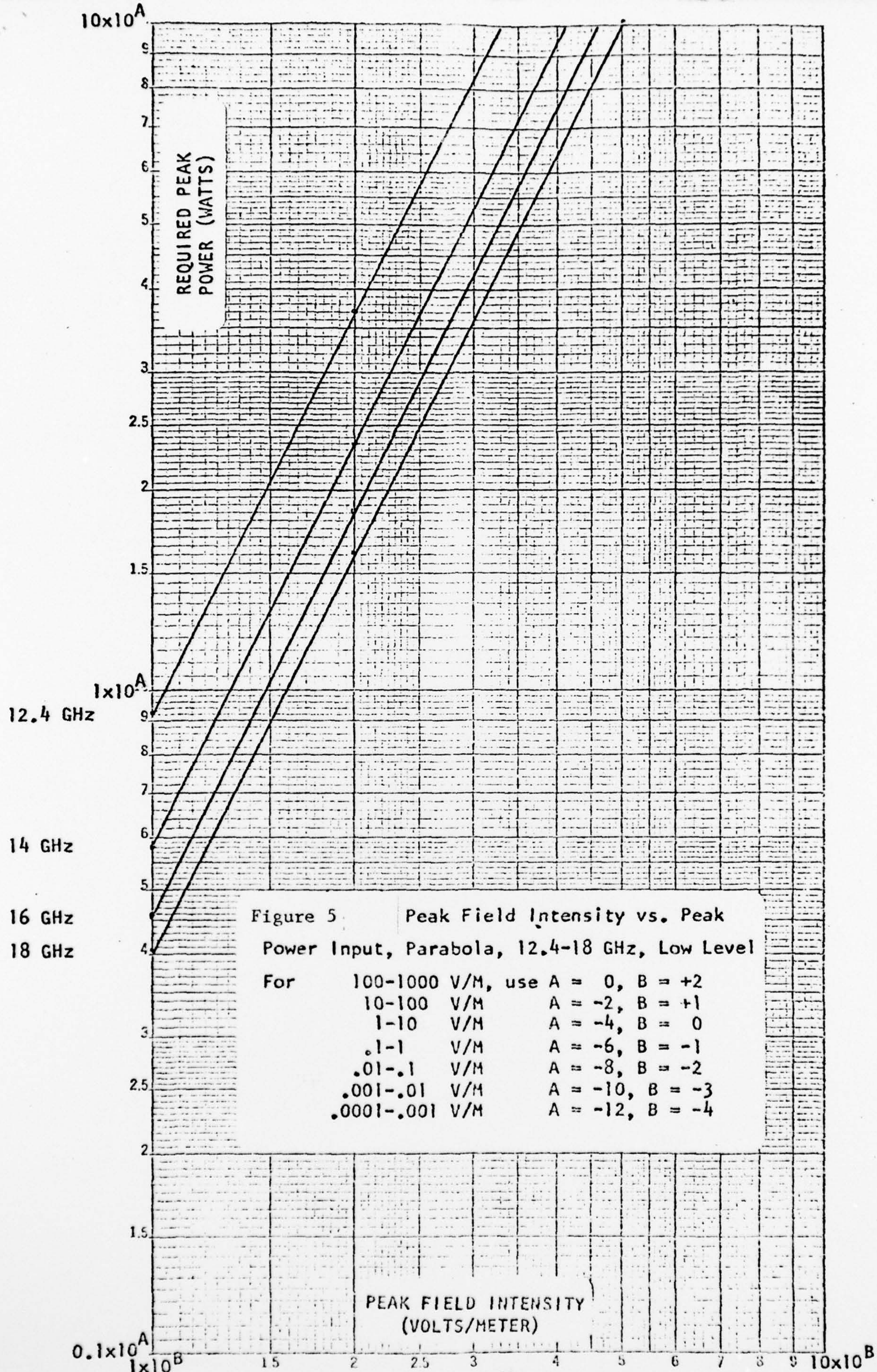


Figure 5 Peak Field Intensity vs. Peak Power Input, Parabola, 12.4-18 GHz, Low Level

For		
100-1000 V/M, use	A = 0,	B = +2
10-100 V/M	A = -2,	B = +1
1-10 V/M	A = -4,	B = 0
.1-1 V/M	A = -6,	B = -1
.01-.1 V/M	A = -8,	B = -2
.001-.01 V/M	A = -10,	B = -3
.0001-.001 V/M	A = -12,	B = -4

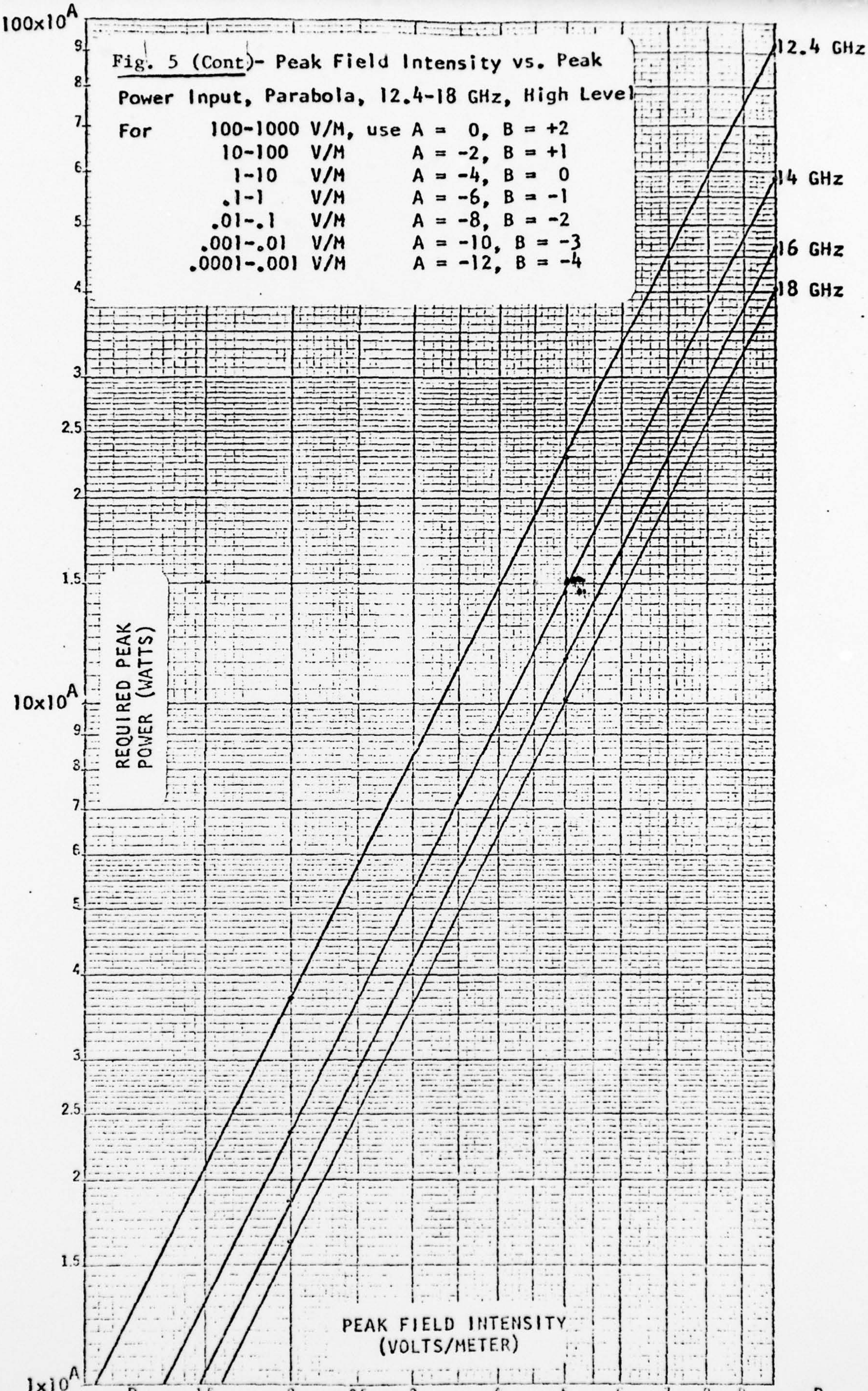


Fig. 5 (Cont)- Peak Field Intensity vs. Peak Power Input, Parabola, 12.4-18 GHz, High Level

For		
100-1000 V/M, use	A = 0,	B = +2
10-100 V/M	A = -2,	B = +1
1-10 V/M	A = -4,	B = 0
.1-1 V/M	A = -6,	B = -1
.01-.1 V/M	A = -8,	B = -2
.001-.01 V/M	A = -10,	B = -3
.0001-.001 V/M	A = -12,	B = -4

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