QEX 23 January 1984

The ARRL Experimenters' Exchange

M^2I^2 Users' Cult

It's time to start a Sony 2002 receiver users' cult (M^2I^2 UC) -- via QEX. No, the League's not pushing Sony or any other particular commercial product. The Sony 2002 represents a breakthrough in small portable receiver design and could have some interesting Amateur Radio applications. For one thing, it is fully synthesized, has digital readout and general coverage of the MF-HF bands, can copy SSB and CW, and is small (slightly over 7 x 4 x 1 inches). It's a bit more expensive than other receivers of its size, but the price of around $240 is within reach of many radio amateurs.

Oh, you're still wondering what M squared I squared stands for... 2002 in Roman numerals (MMII).

So, why don't you just say that the 2002 is a nice tiny receiver for hams to use and just let it go at that? No, that wouldn't do it justice. The 2002 has so much capability packed into its small box that it sets a new standard of what is possible in a commercially built receiver of that size. The other side of the coin is that the 2002 is not the ultimate receiver by any means; it has shortcomings.

For one thing, the manufacturer installed a little knob for fine tuning between the 5-kHz increments provided by the synthesizer and SSB signals are tricky to tune. Some experimentation in the lab showed that paralleling an external potentiometer with a bigger knob might be a satisfactory fix.

The receiver audio output frequency on either CW or SSB changes when strong signals are received. The manufacturer certainly noticed this fault as evidenced by a two-position attenuator switch. A wider choice of attenuation would help but would not compensate for fading signals. There must be a way of modifying the receiver to eliminate this problem. On the surface, it seems to be BFO pulling with strong RF inputs, not as a result of loading the power bus with heavy audio power amplifier current demand. Tracing it any further is deferred until we can get a schematic diagram.

The receiver could also use a bit of audio filtering for better CW reception. I'm not sure whether there is anything along the same lines that can be done to improve the SSB reception short of doing something to reduce the pre-detection bandwidth.

These are the main problems I have noted thus far but it probably has a few other quirks. I mention them to tantalize a few experimenters into developing a fix for these glitches and sharing the solution with the rest of us who have, or are thinking about getting, the 2002. If you can solve one or more of these problems, won't you write a short article or letter to tell the rest of us how it's done?

One application for the 2002 is to use it for HF direction finding. A project has been started in the ARRL lab to design a DF antenna system for the 2002, using a ferrite loop stick antenna band switched to cover 1.8 to 30 MHz. Hopefully, the result will be a box about the same size as the 2002 that will contain the antenna system. Undoubtedly, we will need a stepped attenuator to cope with the signal-overloading problems already mentioned. When the project is finished, we hope to have a reproducible design, with printed-circuit board, that others can duplicate. The decision is not yet final, but the object is to publish the design in QST or possibly the 1985 Radio Amateur's Handbook. Although the DF antenna system is being designed as a companion to the 2002, there is nothing to prevent it from being used with other portable receivers with equal or better operational characteristics. If you have any ideas on this or other HF DF construction projects, please drop a note to me at Hq.

So, how about it, experimenters? Are you stirred up enough to make this neat little receiver something super? Just remember, popping the cover and making modifications that void the warranty are at your own risk! -- W4RI

Correspondence

10-GHz Station List

Is there any interest in a list of stations on 10 GHz and above? If so, I would be willing to formulate the list and send it to any interested party for an s.a.s.e. The type of information I would need is: Name, call, address and a telephone number, type of equipment and status of operation.

In January, William Shaw, WA4MMP, and I will start tests on a pair of 10-mW Gunnplexers between our home QTHs. The antennas will be 2-ft snow-coasters with wideband FM at 10.7 IF. During the spring, I hope to have a beacon running at a location 20 miles east of Washington, D.C. Time will tell if that project is successful. - Jay Heller, KB3OU, 20315 Grazing Way, Gaithersburg, MD 20879.

A New Circuit for the 1968 VW Bug

I own a 1960 VW Bug with a 6-V electrical system. My problem is that I would like to run my 12-V, 2-meter radio from it. I have an extra 6-V battery to hook in series with the car battery, but I want a circuit that automatically gives 6-V to the car; 12-V to the radio and one that will maintain the charge in both batteries from the car's generator. - Scott T. Williams, AE6U, P. O. Box 73, Gualala, CA 95445.

[Ed. Note: This letter was sent to the ARRL Technical Information Service. We could only suggest that he charge in parallel and discharge in series. I recall at least one satisfactory solution from the days when the automotive industry was converting cars from 6- to 12-V electrical systems. It involves the use of an elegant solid-state (or relay) circuit. If you have one, please write Scotty and drop a copy to QEX. We'd like to know the result as there is also a problem of how to operate 24- or 28-V radios from 12-V automotive electrical systems.]

Interface for the TI 99/4A

Randy A. Kovach, KC2SD is looking for information that will help him in the construction of a cassette interface for the TI 99/4A computer. Correspondence can be sent to Randy at the upper-front apartment, 17 Liberty St., Batavia, NY 14020. - Robert Schetgen, KU7G, Technical Information Specialist, ARRL HQ.

Observations of the STS-9 - W5LFL Mission

The following is a set of observations regarding antennas and operating conditions employed for the STS-9 - W5LFL mission:

* W5LFL's signal was strong and quieting into most 2-meter FM receivers when he was heard, even handi-talkies with a rubber duckie antenna.

* Cross polarization of receiving antennas and polarization QSB effects were not noticeable at all.

* The T-R antenna (Oct. and Nov. 1983 QEX) performed well as long as W5LFL's position was greater than 10° above the horizon, but poorly below that angle. Florida passes were below 10° most of the time.

* Mobile stations with vertical whip antennas had no problem copying W5LF1 on horizon passes at ranges of more than 900 miles.

* My station successfully contacted W5LFL at only 2 to 7° elevation, even though the pass had elevations up to 40° and my 16-element Yagi was mounted on an elevation rotator. Amplifier power output level was in the 100 W category.

* Estimated ERP for the contact was 3200 W, barely enough to overcome the white noise desensing of the multitude of FM stations on each uplink channel. Other stations acknowledged by W5LFL had equally high (or higher) ERP needed to "capture" his receiver.

* Successful contacts with gain antennas required that the operator have the ability and equipment to accurately track the STS-9 spacecraft.

* I would be interested to learn if successful contacts using the T-R or whip antenna were made.

The prejudicial opinion of this writer is that linear modulation techniques should have been employed for the mission making contacts more successful. It is clearly understood, however, that the type of equipment used by W5LFL had to be channelized and compact to make it practical aboard the STS-9 -- ruling in favor of FM communication equipment. - Dick Jansson, WD4FAB, 1130 Willowbrook Trail, Maitland, FL 32751.

Comments on the Kantronics Interface Package

My new Commodore 64/Kantronics Interface/Hamtext Software package is doing a beautiful job keying my homebrew QSK kW with no problems. Wish I could type!

The Kantronics Interface Instruction Manual contains instructions for connecting their TTL compatible keying circuit to older style grid block keying circuits, i.e., T4XC. They recommend using a mechanical relay. Some of us are old fashioned, but this is a bit extreme.

The PC board does not put out enough dc and voltage to operate a transistor keying circuit, but the appropriate higher voltage can be packed off closer to Q4 as shown in Fig. 1 below.

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(Correspondence continued)

I preferred to add a new key-out jack without disabling the original keying circuit. I did not have to remove the PC board from the cabinet to make this modification and it can be performed without drilling any holes.

An out-board keying circuit is shown in Fig. 2. It can be constructed in a small box and tucked out of sight. Shielded cable should be used in a QRO shack. This circuit will key most grid block keying rigs at any speed.

The key up/down voltages shown in Fig. 2 are typical. The 1N4004 diode added in Fig. 1 lowers the Kantronics keyout voltage to about +0.75 V. This is high enough. The 1N4004's in Fig. 2 serve to clamp the key down voltage on the grid block circuit to about +0.2 V.

I just finished reading Pagel's article in the December 1983 QST (Fig. 1, p. 39), but prefer not to dig into the T4XC. His circuit could also be mounted out board. In this case, I needed 0 V key up and +1 V key down. With the Kantronics Interface at the T4XC, I had -35 V key up and 0 V key down. - Dave Kennedy, N4SU, Rt. 3, Box 100 Mtn. View Rd., King, NC 27021.

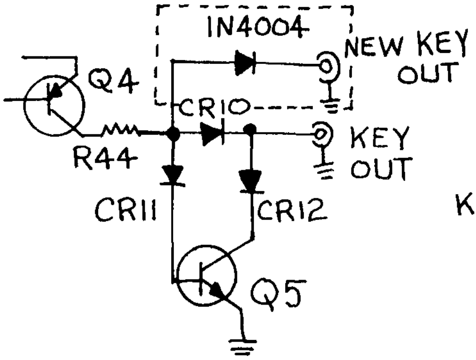

Fig. 1 -- Kantronics Interface Key out Modification

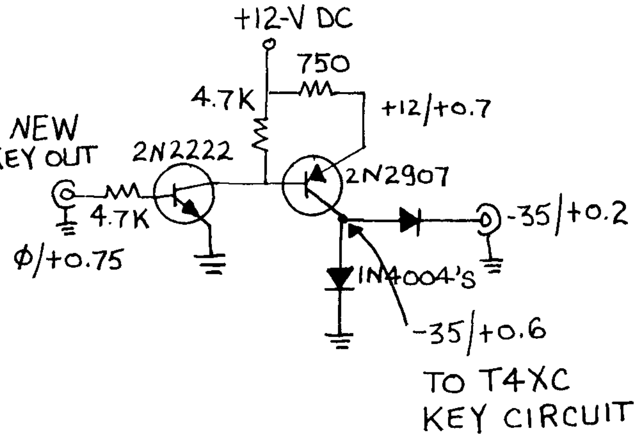

Fig. 2 -- Out board Keying Circuit. Key up/Key down voltages shown are typical.

GWU Engineering Short Courses

Continuing engineering education courses offered by The George Washington University may be of interest to professionals wanting to take advanced communications system courses. Short courses aren't cheap these days ($695 - $875), so these are not for you unless your company is footing the bill. Here are some upcoming advanced communications courses:

Data Communications Standards: Interfaces, Protocols and Network Architectures	February 6-8, 1984
Spread Spectrum Communications	February 6-10, 1984
Communications Satellite Engineering	February 13-17, 1984
Data Communications Systems and Networks	March 12-16, 1984
Synchronization in Spread Spectrum Systems	May 14-18, 1984

If interested, contact the Continuing Engineering Education, The George Washington University, Washington, D.C. 20052, telephone (800) 424-9773, (in the D.C. area 676-6106), telex 4992135.

Series Line Matching Sections for Impedance Matching

By Russell E. Prack, K5RP *

I recently purchased the latest edition of the ARRL Antenna Book and was pleased to see information on series line matching sections for impedance matching on pages 5-10 through 5-12. My initial introduction to this means of impedance matching was through Mr. Reiger's July 1978 QST article. I was surprised to learn that many I spoke to knew little about this and showed no interest. They are missing a good bet!

The design formulas are easily programmable though they look like a lot of work. I prepared such a program after reading Mr. Reiger's article and found that by entering the Zo of various cables, load impedances over a wide range could be easily matched to a 50-ohm line. I developed a table of the various lengths needed to match 5-ohm to 500-ohm loads to 50-ohm line, all computed for 1 MHz. From this table, it is only necessary to divide the lengths by the operating frequency and multiply the appropriate velocity factors to obtain the proper length.

While computing by the method described above, I discovered in many cases a distinct advantage to placing two of these sections back-to-back, thereby shortening the overall length of the entire matching system. This can be illustrated by my first attempt to use series line matching sections to match a 25-ohm load. As there is nothing spectacular about matching a 25-ohm load, it gave me "hands-on" experience with the series section method.

Fig. 1 illustrates a single series line section to match a 25-ohm load to a 50-ohm line on 3.8 MHz. It is shown using 450-ohm line and is the usual one-section method. Fig. 2 shows two series line sections in a back-to-back format to accomplish the same match.

Note that in Fig. 1 the overall length, L1 + L2, is 93.2 ft. In Fig. 2, the overall length is but 52.84 ft, or 43% shorter. These lines will become shorter when the V factors are applied.

The shorter two section arrangement in Fig. 2 has the advantage of less loss, less weight and less material. In Fig. 1, I somehow became "suspicious" of the effectiveness of very short segments such as 3 ft at 3.8 MHz. Also, because L2 and L22 of Fig. 2 have the same 50-ohm Zo, these two sections can be of one continuous length, but the length of each must be calculated separately.

Line Z at points A-A in Fig. 2 is 450 + j0. The lengths L1 + L2 take the 25-ohm load to 450 ohm at A-A and the lengths L22 and L11 carry the 450-ohm impedance' down to the required 50-ohm match. From various calculations and comparisons with the published Zo, I found the results somewhat sensitive to what the **actual** Zo of the lines are. This is true when matching low impedances. Therefore, good quality lines should be used -- lines whose specifications can be depended upon. High Zo lines should be capable of handling high current and a low Zo line should be able to handle high voltage depending upon design.

The lengths involved are not standard. They can be cut by the use of a good noise bridge at their quarter-wave resonant frequency. For example, in Fig. 2 the quarter-wave resonant frequency of L1 is:

$$\frac{246}{8.45\text{ ft}} = 29.11\text{ MHz} \quad L2 + L22$$

$$\text{Resonance is } \frac{246}{31.83\text{ ft}} = 7.73\text{ MHz, etc.}$$

This is not necessary unless a higher degree of accuracy is required. Mere line length multiplied by the V factor gives good results when using good line.

I found that these section lengths are not extremely critical. In Fig. 2 for example, if L1 is 1 ft shorter and L2 + L22 is 2 ft too long, L11 can be made a bit long and pruned to about 11.8 ft. In such a case, the impedance at the end of L11 should be 56.4 ohm and not 50 ohm. This will net an SWR of 1.13:1.

With regard to the bandwidth of the matching **system alone** in terms of SWR, if the load is kept constant at 25 ohm over the band, the SWR of the single section in Fig. 1 computes to less than 1.3:1 at 3.6 MHz and 4.0 MHz. This will not happen with an ordinary antenna. The two-section system shown in Fig. 2 computes to approximately 2:1 at 3.6 MHz and 4.0 MHz. Not bad.

The SWR was computed by shortening each segment by a factor of 3.6 to 3.8 for 3.6 MHz and lengthening each segment by a factor of 4.0 to 3.8 for 4.0 MHz. The Z (toward the generator) was then computed for the end of L1 and that complex Z was used as the load for L2. The Z was then computed for the end of L2, and so on. The final Z was plotted on a Smith Chart and the SWR scaled from that plot. Please note that these are SWR ranges for the examples shown and they could be different for other impedance transformations.

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(Prack continued)

Finally, for those interested in phase shift in this system, note that the shift is not what the total electrical length of the line is, even when matching purely resistive loads. This is because the line lengths are not multiples of a quarter wave and each matching segment has a different SWR. In Fig. 1 the total length is about 130°, but the phase shift will be 145°.

Did my two sections back-to-back work? Using a Bird Model 43 Thruline with a 100-W element, I carefully adjusted the forward power to read 100. In the reflected position I was able to see the needle move from 0. (The meter was checked prior to my tests.) The overall length of my two sections on 3.8 MHz was 41 ft after applying the V factors and a minimum pruning of L11.

In my opinion, series line sections do a good job. They are easy to make and easy to adjust. Using this method allows an infinite number of combinations and applications. My congratulations to Mr. Reiger and to the ARRL for bringing the Amateur Radio community this fine work and information.

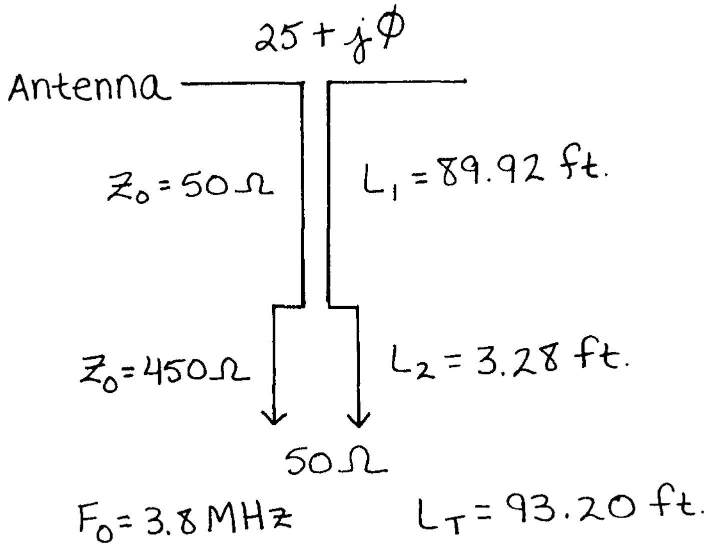

Figure 1

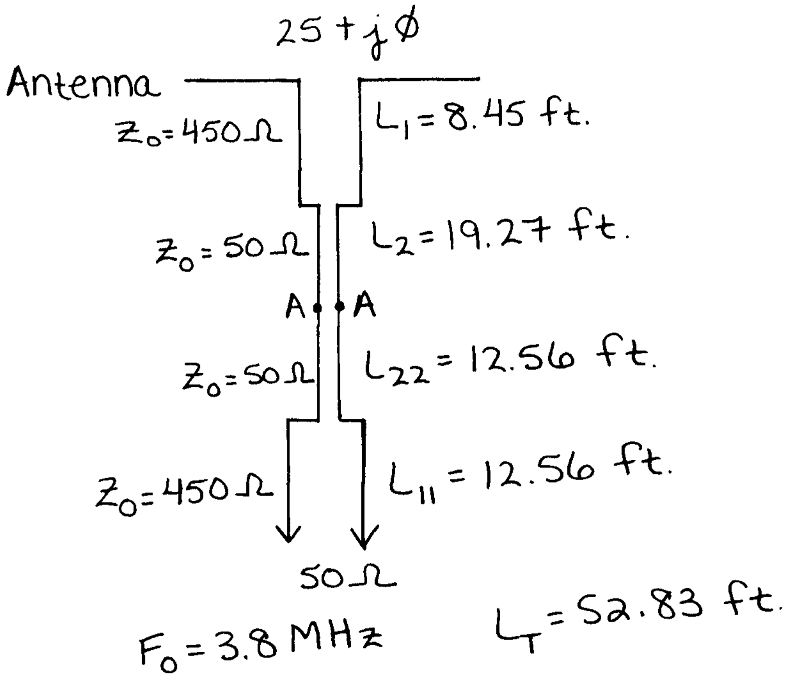

Figure 2

Data Communications

Conducted by
David W. Borden, * K8MMO

Applications of Packet Radio

Introduction

I shall embark on a series of articles explaining the various applications packet radio technology can easily be used with and will add new applications to the series periodically. Applications are important because interest dwindles in most experimental subjects as the result is achieved. For example, a large number of people jumped on the packet radio bandwagon in St. Louis, obtained TNC boards and began to bleep at each other. Finally, a CBBS (the Ward Christensen cork and pushpin computer type) bulletin board was added to the local area network. In the end, these experimenters went on to find a new horizon, using the packet system to transfer computer files to fellow experimenters.

It has happened in Washington also; St. Louis is not unique. The users become discontent typing to each other. This is normal because packet radio, in the converse mode, is actually no better than voice. The difference is its versatility. Many things can be done with packet radio and this makes it essential to the future of Amateur Radio. Digital communications is here to stay, but it must be realized that a large user base is needed to demand a real "network." The first application I shall propose is direction finding.

I. Direction Finding Network

The ability to locate a radio transmitter has preoccupied the minds of many amateurs. In Amateur Radio, it usually requires a bogus transmitter run by a third party on your repeater. This is an exercise of preparedness should a thief ever steal ham radio equipment in your area. Thinking it is a CB radio, he or she may come up on your machine. Direction finding techniques would be a plus to pinpoint the thief.

Often an amateur will interfere with repeater operations for various reasons, failing to identify. Direction finding would be ideal to pinpoint the offender. Once this is done, which authority do you contact?

First pinpoint the offender in all cases. A common cause of interference on club machines is the unintentional interference, the spur. Often, a repeater, not in the amateur service, but used by the FBI, Police, Fire, and so on, can cause an intermod product to fall on the input causing mostly noise. You can see that direction finding is helpful in finding the offender, especially in a crowded RF environment.

What role does packet radio play in direction finding? I will first introduce the hardware and will keep it simple. Each Amateur Radio station taking bearings on the target transmitter should have an inexpensive computer. I recommend the Commodore 64. The computer accepts bearing information, forms the information part of a standard AX.25 packet frame and sends the data to the packet board. AMRAD uses Vancouver boards with Host Mode ROMS. Simple enough so far, but what is taking the bearings?

This is the more expensive part of the scheme. You could use a simple radio with directional antenna, popular on "bunny" hunts. However, you get what you pay for in terms of bearing accuracy and quickness of fix. Thus, I use a DF box called the Doppler Systems direction finder. This little box, with circular LED display, is used with four omnidirectional antennas. My club's chief DFer used this box in his automobile and won a "bunny" hunt starting an hour late and beating the next winner by one hour. The kit for this device sells for $270 and can be bought from Doppler Systems, 5540 E. Charter Oak, Scottsdale, AZ 85254.

Test results found the Doppler Systems direction finder to operate poorly from fixed locations because of multiple reflections of the target signal off buildings and water towers. It is excellent mobile, however, because reflections tend to become nonexistent. My next proposal is to try a fix for this where three sets of four antennas each are mounted on a roof and are computer switched for reflection elimination. The plan from here becomes more digital.

One of the reasons I like the Doppler Systems box is that it can produce a digital representation of the bearing. This is then taken in by the computer, formed into a bearing message and sent to the Vancouver board where it is made into a packet. Thus the computer watches the DF box intermittently, forms a DF message under software control and commands the packet board to dispatch the information. Here the plan has two possible schemes for assembling the fix.

Assume that three of these systems are available and monitoring a given repeater input frequency. Word has spread that a thief has stolen a hand-held and is trying to initiate a conversation on the repeater. For the sake of our geography, the thief is located in downtown Washington, D.C. and the three DF stations are located at points around the Capital Beltway. In scheme no. 1, a master station collects all the bearings, performs the calculations required to make a fix and assembles a fix message that is sent to the Vancouver board where it is broadcast to all. Receivers of the fix then plot this on their map of the Washington Beltway area displayed on their computer. The problem with plan no. 1, however, is the

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*Rt. 2, Box 233B, Sterling, VA 22170

The Theory of Diode Voltmeters and Some Applications

By Albert E. Weller, WD8KBW *

Both simple solid-state theory and empirical observations can indicate what the current through a solid-state diode is by the "Diode Equation":

$$I = I_o\left(e^{\frac{\Delta E}{V}} - 1\right) \qquad \text{(Eq. 1)}$$

where

I = current
V = voltage
Ir = the reverse saturation current

This theory yields a value of about 0.025 volts for V, with values ranging from 0.025 to 0.050 that are observed in practice. These values may vary with the diode current.

The concept of a constant reverse saturation current proves to be oversimplified. Real diodes display features ranging from resistive "leakage" current to rapid changes in reverse current with diode voltage drop and negative resistance. The diode equation, with suitably chosen values of V and Ir, can yield a satisfactory representation of diode behavior. At low diode voltage drops I experimentally found that V = 0.035 volts with Ir = 3 EE-9 (a) for silicon diodes, and Ir = 3 EE-7 (a) for germanium diodes. These values are used as example calculations in this article. [Ed. Note: Because the QEX printer does not yet have a feature for printing subscripts, superscripts or raising numbers to a power, it should be understood, for example, that 3 EE-9 is another expression for 3×10^{-9}.]

The Simple Diode Voltmeter

Fig. 2 illustrates two equivalent circuits: The diode voltmeter or "RF Probes." We can find the diode current by applying the diode equation:

$$I = Ir\left(e^{\frac{E\sin\omega t - Ec}{V}} - 1\right) \qquad \text{(Eq. 2)}$$

The average diode current will be:

$$Ib = \frac{Ir}{2\pi}\int_0^{2\pi}\left(e^{\frac{E\sin\omega t - Ec}{V}} - 1\right)d(\omega t) \qquad \text{(Eq. 3)}$$

The integral can then be evaluated by expanding into the series of e for a result:

$$Ib = Ir\left[e^{-\frac{Ec}{V}}\sum_{n=0}^{\infty}\left(\frac{(E/V)^n}{2^n 2!}\right)^2 - 1\right] \qquad \text{(Eq. 4)}$$

Defining the function S(E/V) as:

$$S^{-1} = e^{-\frac{E}{V}}\sum_{n=0}^{\infty}\left(\frac{(E/V)^n}{2^n 2!}\right)^2 \qquad \text{(Eq. 5)}$$

there results:

$$Ib = Ir\left(e^{\frac{E-Ec}{V}}S^{-1} - 1\right) \qquad \text{(Eq. 6)}$$

Solving for the diode voltage drop:

$$E - Ec = V\ln S + V\ln\left(\frac{Ib}{Ir} + 1\right) \qquad \text{(Eq. 7)}$$

As Ib = Ec/R:

$$E - Ec = V\ln S + V\ln\left(\frac{Ec}{IrR} + 1\right) \qquad \text{(Eq. 8)}$$

The diode voltage drop is defined as the error of the circuit when it is used as a peak reading voltmeter. This error consists of two parts: a "dc error" and an "ac error." The term Vln(Ec/IrR+1) defines the dc error. Its magnitude depends on the input voltage and the diode load resistance, but goes towards zero as the product IrR becomes larger. This error also exists when the input is a dc voltage.

The ac error is defined by the term VlnS. It is independent of the load resistance and diode current, and depends only on the peak input voltage and the value of V for the specific diode. Sample values of S are given below.

E/V	S
0.01	1.0100
0.10	1.1024
1.0	2.1470
10.0	7.8227
100.	25.036
1000.	79.249

The calibration factor, X = Ec/E, for the diode voltmeter is:

$$X = 1 - \frac{V}{E}\ln S - \frac{V}{E}\ln\left(\frac{EX}{IrR} + 1\right) \qquad \text{(Eq. 9)}$$

The X curve with peak sinusoidal input voltage E, and various values of IrR, is shown in Fig. 2. For a 1N914 silicon diode and a load resistance of 22 megohms, IrR is approximately 0.066. The measured points are more consistant with V = 0.04

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*1325 Cambridge Blvd., Columbus, OH 43212

volts and Ir = 2 EE-9 (a) than with V = 0.035 volts and 3 EE-9 (a). For a 1N34A germanium diode with the same load resistance, IrR is approximately 6.6. Hence, the voltmeter with this diode has a negligible dc error while the silicon diode has an appreciable dc error.

The Compensated Diode Voltmeter

The circuit in Fig. 3 shows a way to compensate part of the diode voltmeter error. The output is taken from a tap on the load resistor with the diode equation and its integral used to obtain the two relations:

$$E = IbR + V \ln S + V \ln\left(\frac{Ib}{Ir}+1\right) + nV \ln\left(\frac{Ib}{Ir}+1\right) \qquad \text{(Eq. 10)}$$

and

$$Eo = YIbR + nV \ln\left(\frac{Ib}{Ir}+1\right) \qquad \text{(Eq. 11)}$$

where

n = the number of compensating diodes D1-Dn.

Here the compensating diodes are assumed to have the same V and Ir as the detector diode. This simplifies the analysis. The behavior of diode voltmeters with compensating diodes having different characteristics can be analyzed and is not quantitatively different.

The error of the compensated diode voltmeter is:

$$Eo - YE = YV \ln S + [n(1-Y) - Y] \ln\left(\frac{Ib}{Ir}+1\right) \qquad \text{(Eq. 12)}$$

The dc error can be eliminated by making:

$$n(1-Y) - Y = 0 \qquad \text{(Eq. 13)}$$

$$Y = \frac{n}{n+1} \qquad \text{(Eq. 14)}$$

With the output tap in this position, the diode voltmeter output is independent of the load resistance. In fact it is equal to the output of the simple diode voltmeter with an infinite load resistance. This tap position also eliminates the dependence of the error on Ir so that both types of diodes, with a differing Ir, produce the same calibration curve. The error of the compensated diode voltmeter can be made zero at any desired input voltage by setting:

$$n - Y(n+1) = Y \ln S / \ln\left(\frac{Ib}{Ir}+1\right) = YL \qquad \text{(Eq. 15)}$$

For L values different from zero, the compensated diode voltmeter error, or calibration factor, has some dependence on the load resistance. This dependence is weaker for the simple diode voltmeter.

For L greater than zero, the error is always less positive than for the simple voltmeter. Depending on the choice of L, the error can be negative (X = Eo/YE >1) at some input voltages.

Theoretical calibration curves for several choices of Y and two values of the load resistance can be seen in Fig. 4 The observed calibration differs noticeably from the theoretical and two effects are involved. First, 0.2 sec, or 12 periods of the 60-Hz calibrating voltage was the RC time constant used in the measurements. A time constant of 1000-2000 periods is required to eliminate the dependence on C. Next, the value of V is not constant for the experimental points, but increases from 0.04 volts at 2-V input to 0.12 V at 10-V input. I did not find this behavior during direct determinations of V and Ir, and have no explanation for the effect.

As a practical application, calibrations for a diode voltmeter designed to read the equivalent rms voltage of a sine wave input, (Y = 0.707), are shown in Fig. 5. A choice of n = 5 will provide a good general purpose voltmeter, although an n of 4 produces an acceptable calibration.

Note that the lower portion of the load resistor and the compensating diodes can be mounted remotely from the "probe" proper. There is a price to be paid for this compensation: Regardless of the load resistance at small output voltages, there is a reduction in current and loading by the meter used to read the output voltage becomes important. For two rms reading voltmeters, the minimum output meter resistance is as follows.

Input Voltage, peak volts	*Minimum Output Meter Resistance, Megohms* R = 2 EE6	R = 2 EE5
10	17	2
5	19	2
2.5	25	3
1.0	56	20
0.5	150	120
0.25	290	270
0.10	420	400

* For output meter current = 1/10 x diode current.

Fig. 6 shows the effect of a 22-megohm load on the output of a compensated "RMS" voltmeter with silicon diodes, n = 4 and R = 0.2 megohms. The load's effect on the input voltage is less than 1 volt, as suggested by the above table. This compensation improves the calibration factor over the simple voltmeter at all inputs above 0.25 volts. Unfortunately, the 0.2-megohm circuit has a low input resistance and might not be suitable for general use.

The Biased Diode Voltmeter

It is sometimes suggested that a diode detector be forward biased to increase its sensitivity. Fig. 7 shows such an application to the diode voltmeter. The calibration factor can be shown by:

$$X = (Ec - Eco)/E \qquad \text{(Eq. 16)}$$

$$X = 1 - \frac{V}{E} \ln S - \frac{V}{E} \ln\left(\frac{Ib + Ir}{Io + Ir}\right) \qquad \text{(Eq. 17)}$$

where

Eco & Io = output voltage and diode current

The dc error will be small if the bias voltage is made large enough so that the value of Ib will not be much different than Io. The biased diode voltmeter thus approaches the behavior of a simple diode voltmeter with infinite load resistance.

Fig. 8 shows theoretical calibration curves for several values of the bias voltage. Relatively small forward bias voltages have a substantial effect on the calibration. There is no "optimum" forward bias -- the improvement continues at a decreasing rate as the bias voltage is continuously increased.

A significant feature of this type of voltmeter is that its performance is not greatly affected by the load resistance. For silicon diodes with an IrR of 1 EE-2 and 1 EE-5 V and Eb = 2 volts, the following calibrations hold true.

Peak Input voltage, V	Calibration Factor IrR = 1 EE-2	Calibration Factor IrR = 1 EE-5
10	0.980	0.980
1	0.895	0.894
0.1	0.505	0.504

Forward biasing of the diode suggests a means of improving the calibration of a heavily-loaded diode voltmeter. However, the output device, (e.g., a microammeter), must be capable of zeroing in the presence of the standing current from the bias voltage. Fig. 9 shows two such circuit arrangements.

Square Law Voltmeters

If the ac error term is expanded as a series, there results:

$$V \ln S = V\left[\frac{E}{V} - \frac{1}{4}\left(\frac{E}{V}\right)^2 + \frac{1}{64}\left(\frac{E}{V}\right)^3 - \cdots\right] \qquad \text{(Eq. 18)}$$

from which:

$$\frac{X}{E} = \frac{E_c}{E^2} = \frac{1}{4V} - \frac{1}{64V}\left(\frac{E}{V}\right)^2 + \ldots - \frac{V}{E^2}\ln\left(\frac{E_c}{IrR}+1\right) \qquad \text{(Eq. 19)}$$

When E is sufficiently small and IrR is large, the output of the simple diode voltmeter will be:

$$E_c = E^2 \frac{1}{4V}\left(\frac{IrR}{IrR+V}\right) \qquad \text{(Eq. 20)}$$

Thus, the output is proportional to the square of the peak input voltage. E is sufficiently small if:

$$\frac{1}{16}\left(\frac{E}{V}\right)^2 \ll 1 \qquad \text{(Eq. 21)}$$

The largest possible calibration factor for square law operation is 1/4 V, or about 5-10. Hence, output voltages are small. For example, the output for the largest permissible input for a square law response, E/V = 1, will be:

$$E_c = \frac{E^2}{4V} = \frac{V}{4} = 0.008\,v \qquad \text{(Eq. 22)}$$

The square law response can be extended to higher voltages by using two or more diodes in series. The criteria for square law operation then becomes:

$$\frac{1}{16}\left(\frac{E}{nV}\right)^2 \ll 1 \qquad \text{(Eq. 23)}$$

and the calibration factor is 1/4nV.

where
n = the number of diodes in series.

Fig. 10 shows an experimental and theoretical calibration for a square law voltmeter using two 1N914 diodes in series, for which (IrR + V)/IrR = 2.094 volts.

Circuit Loading

The average forward diode current, Ib, is smaller than the peak current, Ip. The ratio of the peak to average current can be calculated from the diode equation and its integral. Following is a brief listing of this ratio and of the effective diode voltmeter input resistance based on average and peak currents. A load of 22 megohms is assumed in the calculations.

E, V	Ip/Ib	Effective Input Resistance based on Ib Megohms	Effective Input Resistance based on Ip Megohms
10.	42.4	22.6	0.53
5.	30.4	23.3	0.77
1.	14.5	27.1	1.88
0.5	11.3	31.9	2.82
0.1	13.1	84.3	6.42

While the input resistances based on the peak current may appear small, these diode voltmeters are more than adequate. This can be seen when you consider the load resistance of 22 megohms for most modern rf circuitry, where impedances in excess of one or two thousand ohms are rare.

Circuit loading by compensated diode voltmeters should be watched closely. If the diode load is increased to avoid circuit loading, the performance is degraded by the loading of the meter that reads the output. If the diode load is decreased to avoid this effect, the diode voltmeter may heavily load the measured circuit.

Figures 1 through 10 appear on pages 10-14.

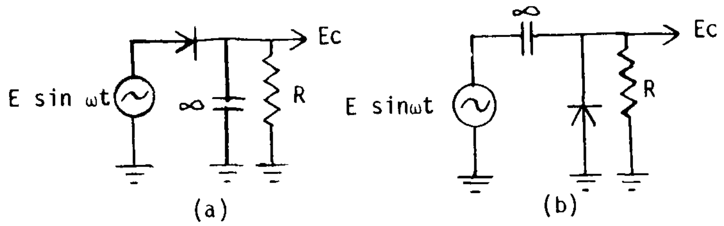

Figure 1. Simple Diode Voltmeters

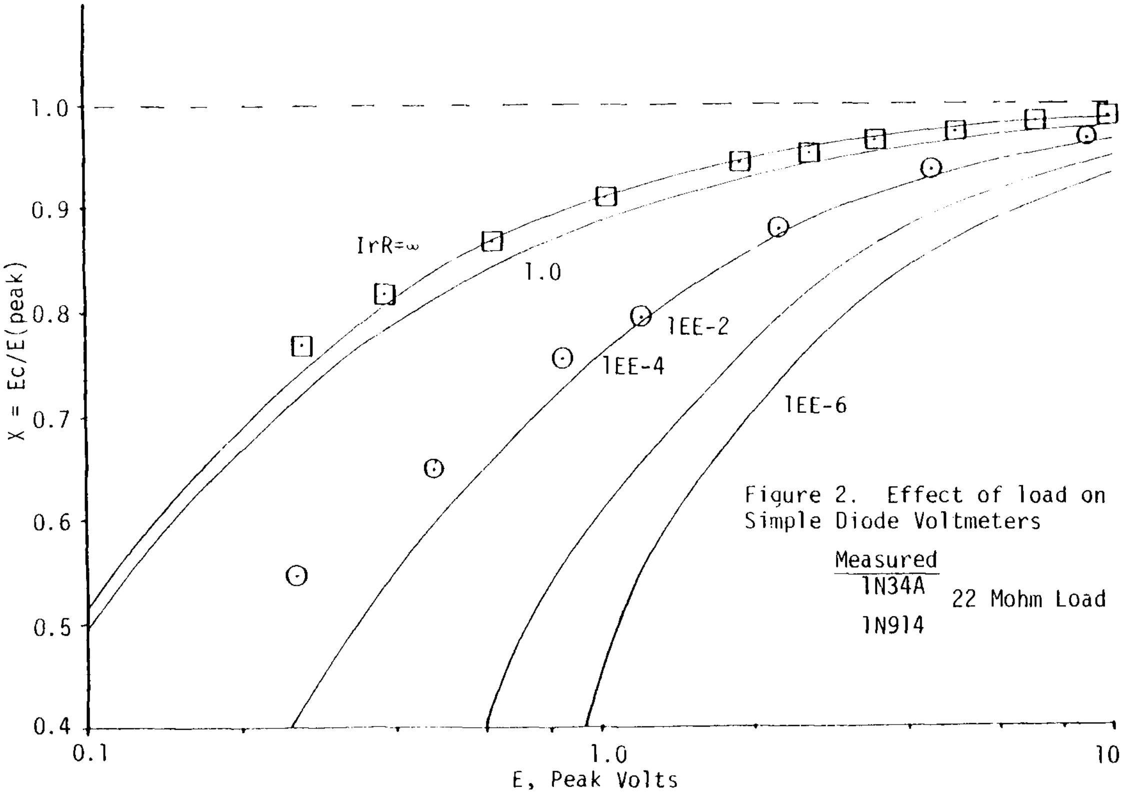

Figure 2. Effect of load on Simple Diode Voltmeters

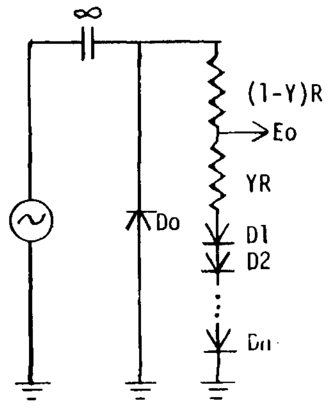

Figure 3. Compensated Diode Voltmeter

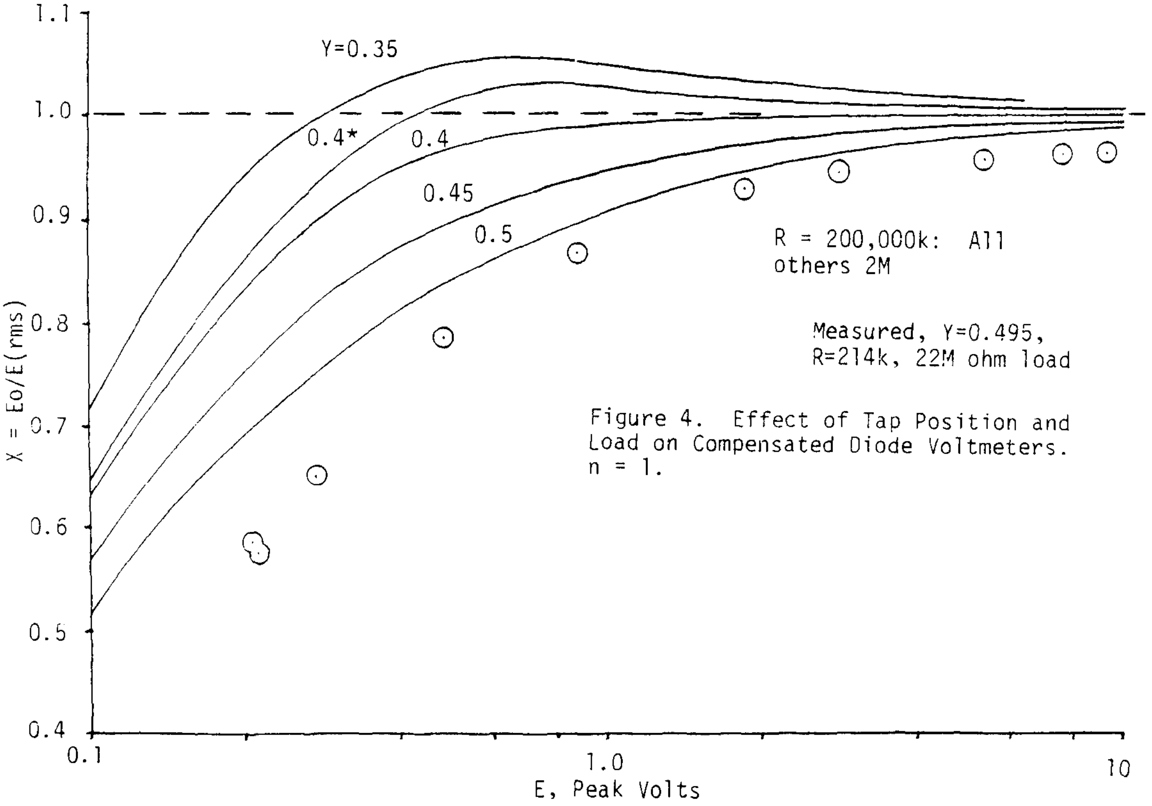

Figure 4. Effect of Tap Position and Load on Compensated Diode Voltmeters. n = 1.

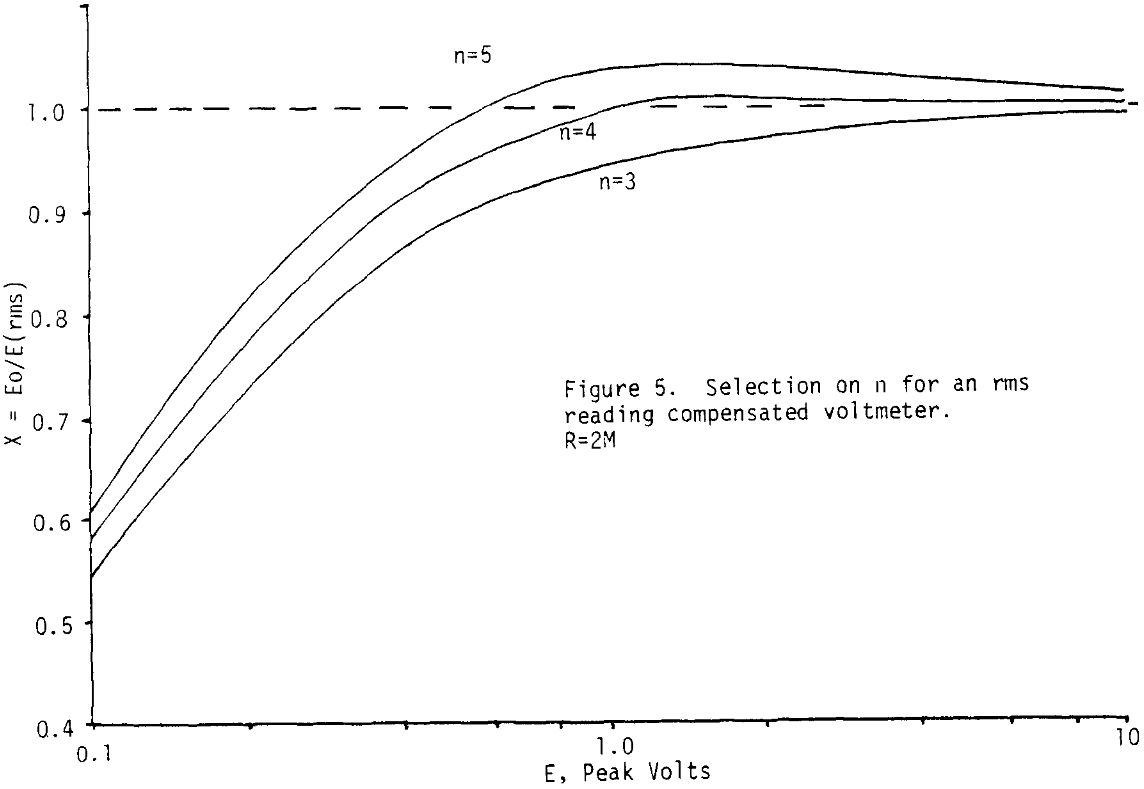

Figure 5. Selection on n for an rms reading compensated voltmeter. R=2M

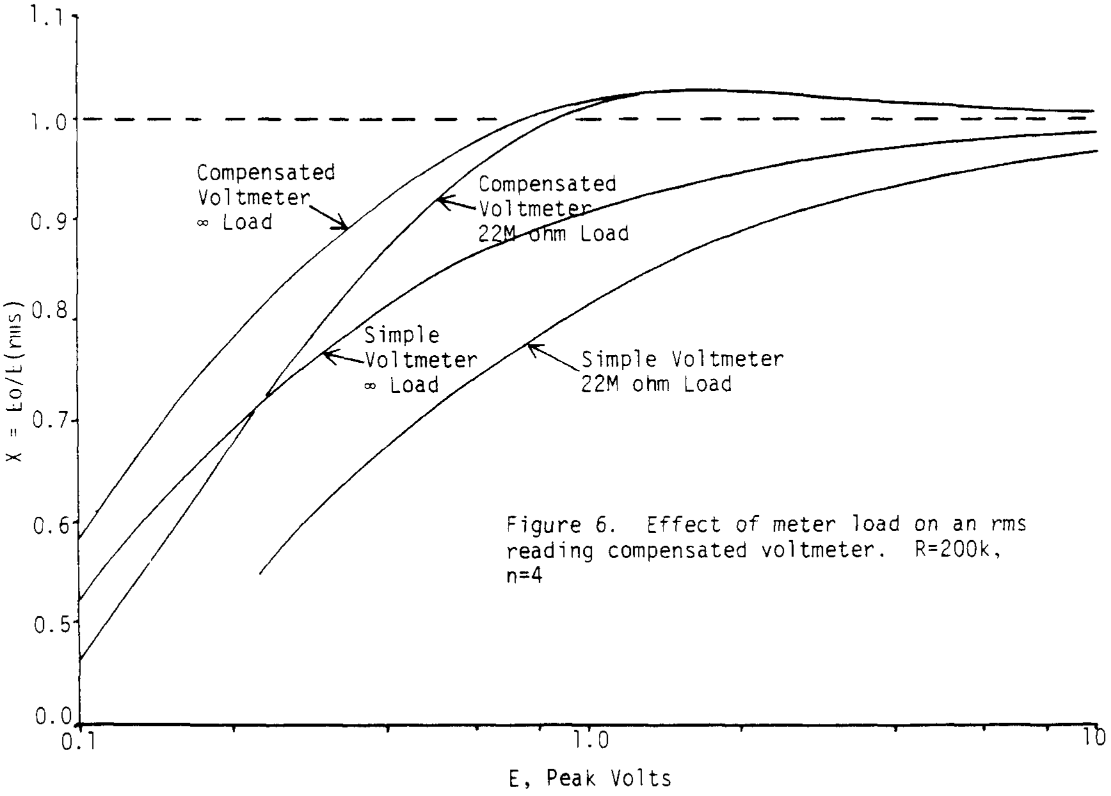

Figure 6. Effect of meter load on an rms reading compensated voltmeter. R=200k, n=4

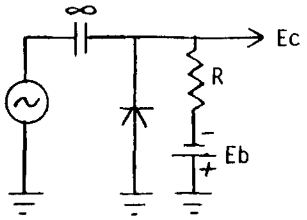

Figure 7. Biased Diode Voltmeter

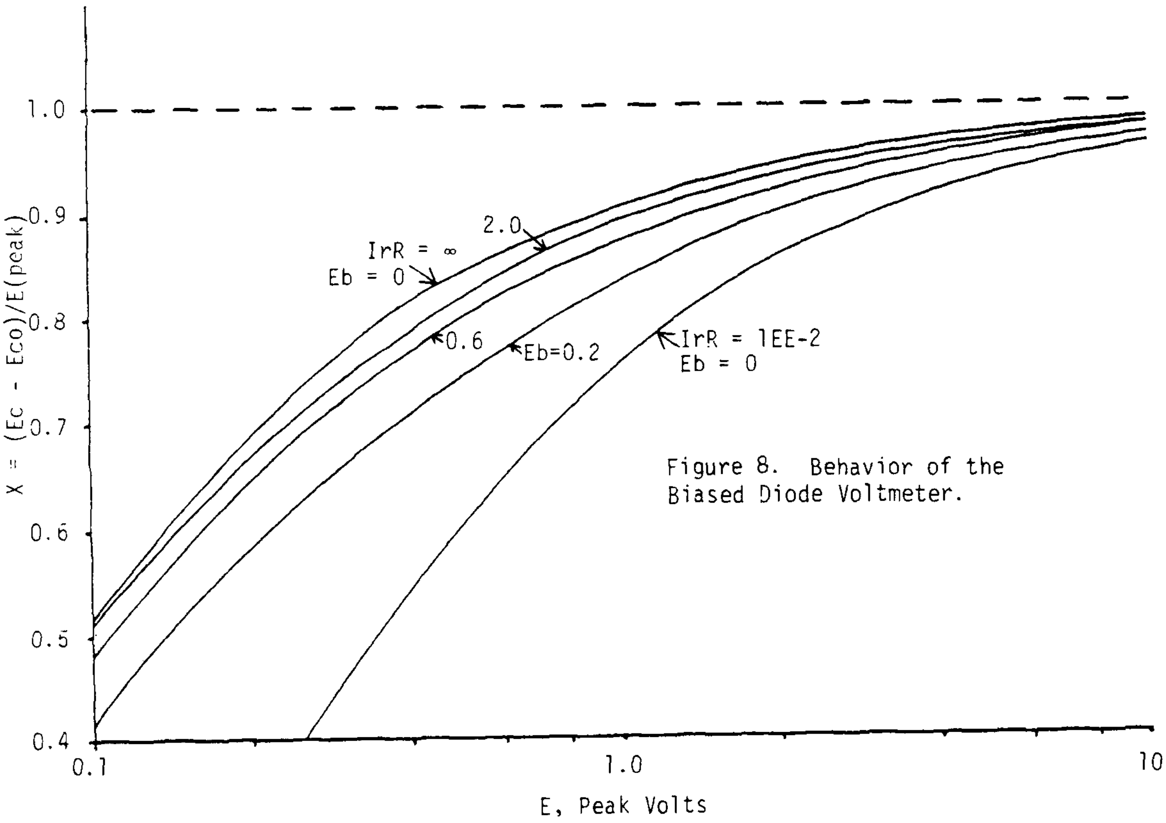

Figure 8. Behavior of the Biased Diode Voltmeter.

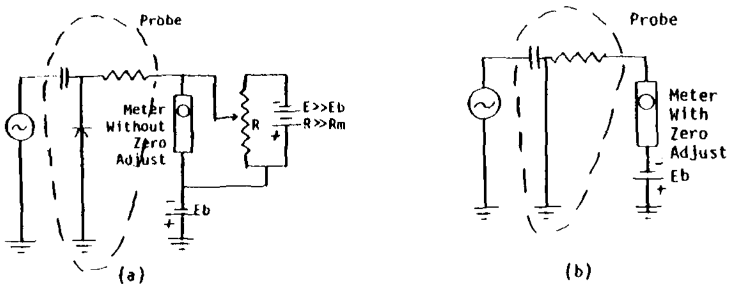

Figure 9. Practical Arrangements of Voltmeters With Biased Diodes

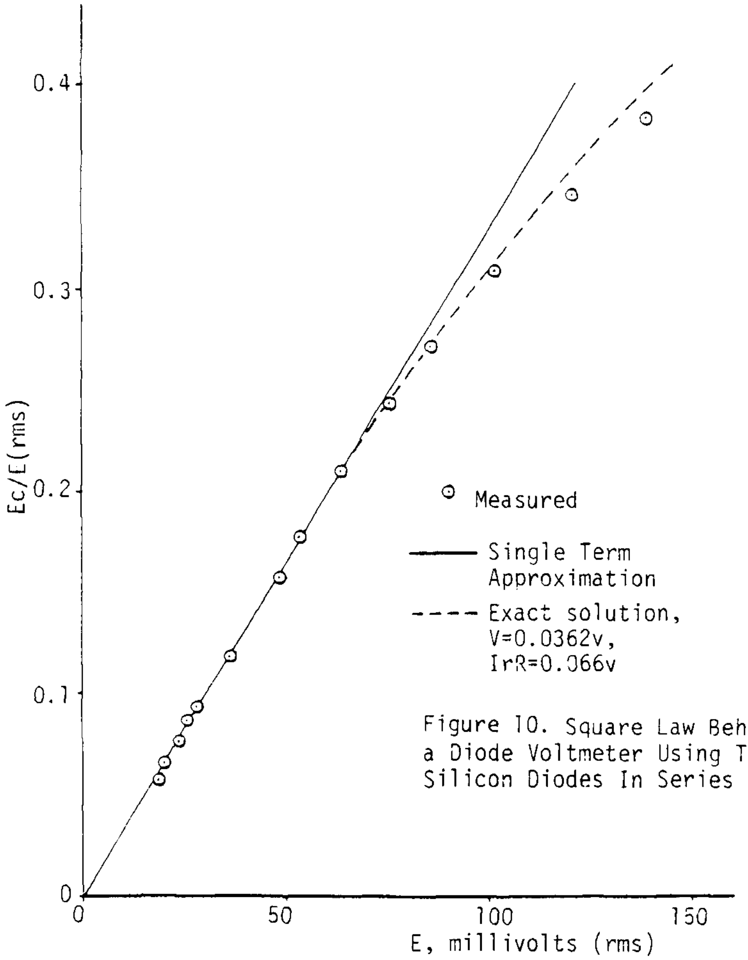

Figure 10. Square Law Behavior of a Diode Voltmeter Using Two Silicon Diodes In Series

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(Data Communications continued)

master station theory as hams do not trust them.

Scheme no. 2 is similar. Each station hears the other stations, performs the fix calculation on their computer and plots the result on their map display. I am not sure if this scheme will work with the computer short on memory with no disk. The fix program is not trivial, but it could be written in assembly language to account for size constraints in the small computer.

Both schemes have a similar drawback. Notice the use of the "monitor mode." No Vancouver or TAPR packet board is connected to any other -- all are running in monitor mode. I believe that what is needed is a master packet station maintaining multiple connections to each DF station. This super master station can perform the fix calculations and feed the fix back to each in turn. However, as I mentioned, hams are distrustful of master stations.

In future application articles, you will notice the basic hardware consists of:

1. A packet radio board (Vancouver or TAPR).
2. An inexpensive computer with a serial port similar to the Commodore VIC-20 or 64.
3. A VHF transceiver.
4. A Bell 212 standard modem (if a Vancouver board is used).

In theory, everything should run off of 12 V and be portable/mobile. I am working on that since our prototype Vancouver system built by Bob Bruniga, WB4APR, burned up after working well for several days.

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