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AUGUST 1990

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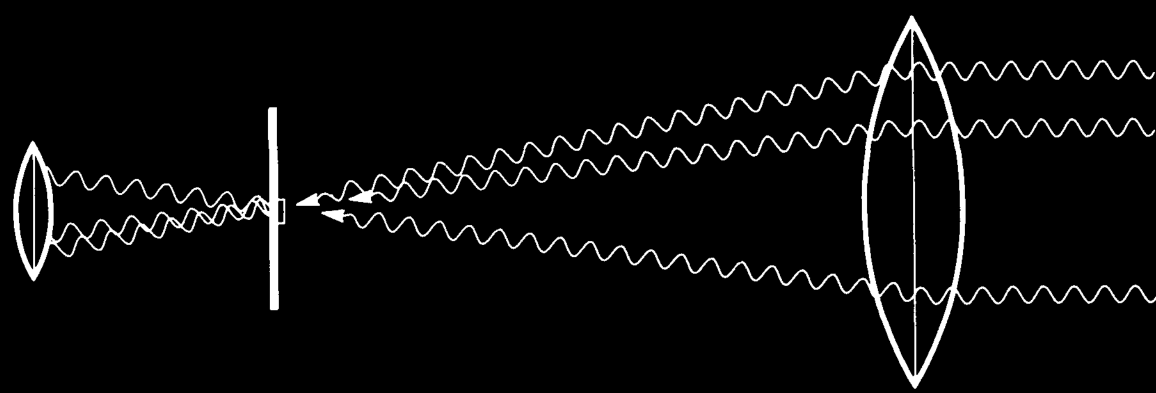

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TABLE OF CONTENTS

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AUGUST 1990 QEX ADVERTISING INDEX

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1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters
2) document advanced technical work in the Amateur Radio field
3) support efforts to advance the state of the Amateur Radio art.

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Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and double spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black-and-white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Empirically Speaking. . .

Spinoffs

Sunday, January 21, 1990, marked the launching of four AMSAT Microsats and two UoSATs into low Earth orbit (LEO) aboard an Ariane V-35. While four of the satellites are in fine shape, two exhibited problems after the launch, though there's still a chance of nursing them back to health. The feat was nothing short of fantastic, and worthy of a year the ARRL Board of Directors has designated "the year of the amateur-satellite program."

Officially, the Microsat program started on November 8, 1987, in Southfield, Michigan, when the AMSAT Board of Directors approved the program. It took about two years of engineering to turn the plan into satellite hardware ready for launch. Software development paralleled that of the hardware; in a sense, it's never done, as new programs can be uploaded to the on-board satellite computers.

The *concept* of the Microsat project, more specifically a packet-satellite (PACSAT) project, was talked about in October 1982. That was the meeting, called by Tom Clark, W3IWI, that resulted in general agreement on AX.25 as the link-layer protocol for amateur packet radio.

AMSAT didn't *invent* the idea of PACSATs, coin the name, or even come up with the concept of small satellites. What AMSAT did, however, was to bring some fine engineering talents together, get there before the crowd, build operational systems, and do so on a budget smaller than the rounding error of aerospace firms.

There's no doubt that the technology developed in AMSAT has migrated to the commercial world and advanced the state-of-the-art. There are two direct spinoffs of AMSAT engineering to Oribital Sciences Corporation (OSC) and Volunteers in Technical Assistance (VITA), a second-order spinoff to STARSYS and possibly some family resemblance with Motorola. All these groups have new small-satellite projects.

The OSC and STARSYS satellite system proposals are similar, are both looking for mobile-satellite service allocations in the 137-138 MHz and 148-149.9 MHz bands. The viability of these systems depends upon the outcome of WARC-92 agenda item 2.2.4(d), "consider possible allocations of up to 5 MHz of a frequency band below 1 GHz to low-orbit satellites on the basis of appropriate sharing criteria." Motorola's "Iridium" system is to use the 1.5/1.6 GHz mobile-satellite band.

There's another (amateur-to-commercial) spinoff area of interest these days. Many are aware that packet radio technology developed by hams has been used by other radio services, commercially and by governments. But there's one, in particular, that not only uses the amateur (AX.25) protocol but does so in the 902-928 MHz amateur band. That system is known as LAWN, which is a radio local-area network operating under the Part 15 provisions for 1-watt spread spectrum. LAWN is not the only Part 15 spread-spectrum system operating in this band, but is the only one we know of using AX.25. As these spread-spectrum systems are relatively new, we have not received reports of interference from them to amateur operations in the 902-928 MHz band. If there are some, the ARRL Technical Department would appreciate the information.

Whatever emerges from the small-satellite initiatives, mentioned above, will change the way the world telecommunicates, particularly in less densely populated areas that are not economical for cellular radio. Amateur radio and satellite experimenters can take some pride in their contributions to this field. One has mixed feelings about commercial systems that benefit from Amateur Radio technology and want to operate in our spectrum, however.—*W4RI*

Calibrating Diode Detectors

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Introduction

A diode detector provides one of the simplest methods of measuring the voltage level of an RF signal. It is effective over a wide frequency range—from audio frequencies to microwave frequencies. It can handle a wide range of voltages—from signals with millivolt levels to signals with tens or even hundreds of volts. It can be added to many circuits for only a fraction of a dollar in parts cost. However, it suffers from one major disadvantage. It is not very accurate unless it is calibrated in the circuit in which it is used. More than one article has stated that the detector circuit, which you have just constructed, needs to be calibrated with a signal of known power level. This statement cleverly conceals the fact that, if you had known the power level of the signal or had access to expensive test equipment, you might not have had to build the detector in the first place.

This article will show one how to calibrate a diode detector to reasonable accuracy using nothing more than a good dc voltmeter and a variable voltage dc source. Three different techniques will be explored. The first is a graphical technique which should allow one to calibrate the detector to within a few dB of the true value. The second is a substitution method that can achieve an accuracy of a few tenths of a dB. The third is a detector circuit whose output will equal the peak ac input voltage to within a few tenths of a dB. All three of these methods can be effective for RF signals greater than a few tens of millivolts.

All of the results were obtained using numerical simulations of the diode detector response. Some of these results have been double-checked by experiment to make sure that the numerical simulations have yielded correct answers. The accuracies given are those predicted by the numerical simulations. Care in construction and calibration can yield real results that are quite close to the predictions.

Diode Detector Theory

A diode detector may be constructed as shown in Fig 1. There are other ways to place the diode(s) to perform the detecting function. However, all of the analysis was done for this detector configuration.

The detector circuit consists of a sinusoidal ac voltage source with a peak voltage of V_{ac}, a diode, a capacitor, and a load resistance with a value of R_L. The dc voltage out of the detector is V_o. The ac voltage source may be any source of RF as long as its source impedance is much less than the load resistance, R_L.

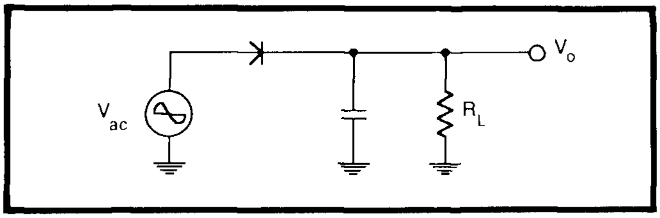

Fig 1—Diode detector circuit

The diode will conduct forward current whenever the instantaneous voltage of the ac voltage source exceeds the dc output voltage. These current pulses are smoothed by the capacitor which absorbs charge when the diode is conducting and releases it to the load resistance when the diode is reverse biased. The capacitor must have a sufficiently high value that the voltage across it will change a negligible amount over each cycle of the waveform. The load resistance will be a relatively large value if the sensing element has a high input impedance such as an op amp and a more modest value when the detector is directly driving a meter. Typically, the load resistance might vary from a few kilohms to a few megohms.

The diode used in the circuit has a large impact on the detected voltage. An ideal diode will act like a switch that is sensitive to the instantaneous voltage applied across it. When the ideal diode is forward biased, it would have zero resistance and when it is reverse biased, it would have infinite resistance. In this case, the output voltage will exactly equal the peak input voltage.

Semiconductor diodes in the real world follow a model which mathematically relates the voltage across the diode and the current through the diode[1]:

$$I_d = I_s\left[\exp\left(\frac{qV_d}{kT}\right) - 1\right] \qquad \text{(Eq 1)}$$

[1]Notes appear on page 8.

where:

I_d = Current through diode in amperes
I_s = Diode Reverse Saturation Current in amperes
q = Charge of an electron (1.6×10^{-19} coulombs)
k = Boltzmann's constant (1.38×10^{-23} joules/K)
T = Temperature in Kelvin
V_d = Voltage across the diode in volts

This equation provides an approximation to the voltage-current transfer characteristics of a diode. It will provide moderate accuracy for forward currents into the milliampere range. At a typical circuit temperature of T = 300K, this equation becomes:

$$I_d = I_s \left[\exp\left(\frac{V_d}{0.026}\right) - 1\right] \qquad \text{(Eq 2)}$$

In a detector circuit, the voltage across the diode and the current flowing through the diode will be time varying quantities. This is illustrated in Fig 2 in which a 1-volt peak sinewave is applied to the detector circuit. The computations were done for a diode with a reverse saturation current of 1 nA and a load resistance of 1 MΩ. The diode only conducts forward current for about 60 degrees of the cycle with a peak current of about 12 μA. However, the average current through the detector is only 756 nA which just equals the current through the load resistor. The peak positive voltage drop across the diode is 244 mV which is caused by the high peak current flowing through the diode. This voltage drop is much greater than what one would get if a 1-volt dc signal is applied to the diode. This article will explore methods of getting a better estimate of this peak voltage drop and using that data to calibrate the diode detector for ac voltage sources.

The reverse saturation current is dependent on the type of diode. Table 1 gives the parameters of some typical diode types. Most diodes will conduct more current than this equation predicts when they are reverse biased. Germanium diodes are especially a problem because their leakage current may be as large as the current through the load resistor if it is 100 kΩ or greater. Even with a reverse voltage of a few tenths of a volt, they show a leakage current much greater than 100 nA. This can be shown by measuring the reverse resistance of the diode; it should be a megohm or greater but is generally much less. A number of 1N34As that I measured showed a reverse resistance between 100 kΩ to 1 MΩ with a few tenths of a volt applied to them. This effect will decrease the accuracy of the calculations for germanium diodes when they are used in circuits with a detector load resistance greater than 100 kΩ.

The value of the reverse saturation current will vary with the diode current. I measured several samples of a 1N34A, a 1N5711 and a 1N4148 at several different forward diode currents. The results for the median diode are shown in Table 2. The germanium and Schottky

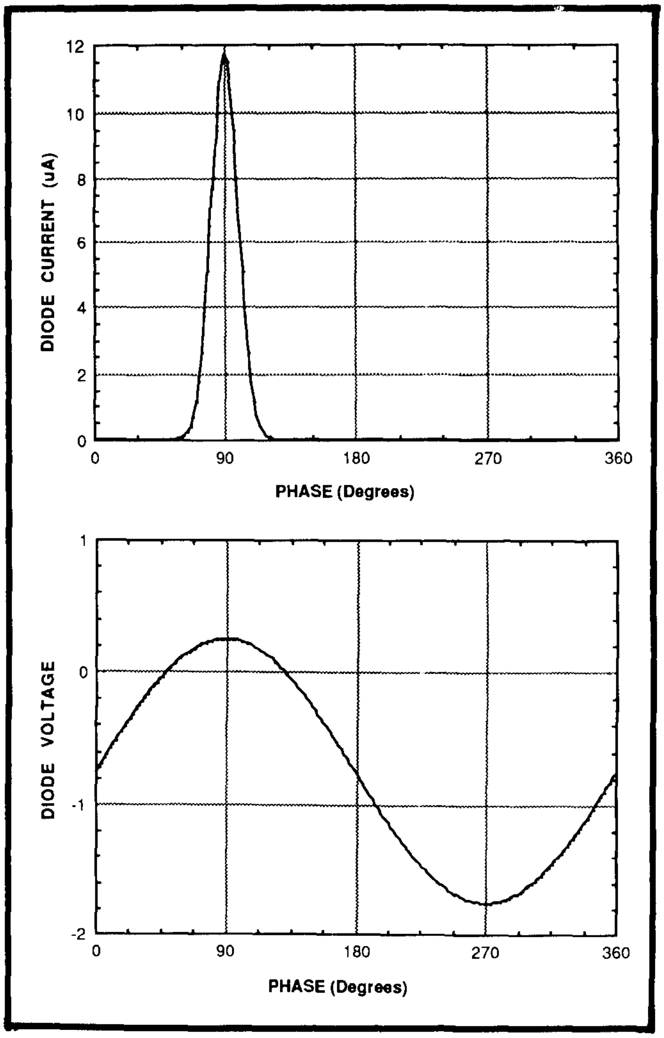

Fig 2—Current through a diode and voltage across the diode in a detector circuit. The diode has a reverse saturation current of 1 nA and the load resistance is 1 MΩ.

TABLE 1

DIODE TYPE	EXAMPLE	TYPICAL REVERSE SATURATION CURRENT
Germanium	1N34A	100 nA
Schottky	1N5711	1 nA
Silicon	1N4148	1 pA

TABLE 2

	I_s (I_d=1 μA)	I_s (I_d=10 μA)	I_s (I_d=100 μA)	I_s (I_d=1 mA)	I_s (I_d=10 mA)
1N34A	215 nA	194 nA	122 nA	33 nA	2.6 nA
1N5711	5.6 nA	5.3 nA	5.0 nA	2.1 nA	0.2 pA
1N4148	5.4 pA	0.75 pA	0.12 pA	0.020 pA	0.002 pA

diodes showed a relatively constant reverse saturation current for diode currents less than 1 mA. However, the silicon diode varied considerably even at low diode currents. Other types of silicon diodes may not vary as much as this type does.

The output of a diode detector is dependent only on the product of the reverse saturation current and the detector load resistance. Since this product is a voltage, I shall refer to it as the reverse saturation voltage:

$$V_S = R_L \times I_S \qquad \text{(Eq 3)}$$

Be aware that this quantity doesn't have any physical meaning. Fig 3 shows the relationship between the peak ac input voltage, the detector output voltage, and the reverse saturation voltage.

The diode response is also affected by the frequency of the signal going into it. At dc and low frequencies, the detector acts pretty much as in Equation 2. As the frequency is increased, the parasitics of the packaging and the characteristics of the semiconductor device can significantly change the results. I have measured several 1N5711s and found them to be flat within ±0.1 dB up to 200 MHz. However, Lewallen found that a 1N914B was down −1.3 dB at 65 MHz and −7.7 dB at 500 MHz[2]. We have both measured 1N34As and found that some samples were flat to 500 MHz while others were down −1 dB at 100 MHz and as much as −4 dB at 500 MHz. Construction techniques for the circuit containing the diode can also play an important role. Long lead lengths will surely cause increased errors at the higher frequencies.

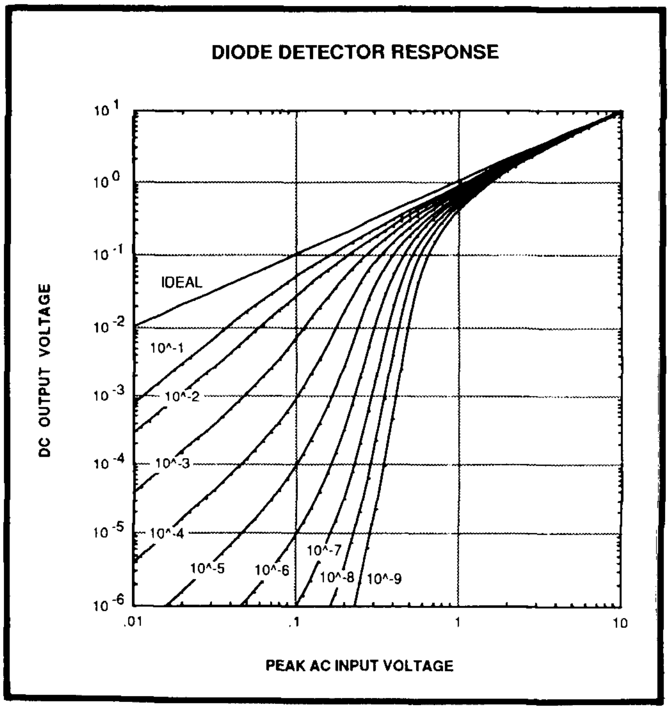

Fig 3—Chart showing the relationship between the peak ac input voltage, the detected output voltage, and the reverse saturation voltage of the diode detector. The curves are shown for an ideal diode and diodes in detector circuits with reverse saturation voltages of 10^{-1}, 10^{-2}, 10^{-3}, 10^{-4}, 10^{-5}, 10^{-6}, 10^{-7}, 10^{-8}, and 10^{-9} volts.

Graphical Calibration

Fig 3 could be used as a graphical calibration aid. However, it has limited accuracy. A better chart for doing this is given in Fig 4. It relates the peak diode voltage drop to the dc output voltage of the detector for several different reverse saturation voltages. To use the chart, one finds the measured detector output voltage on the horizontal scale and slides up to intersect the estimated line for the actual diode reverse saturation voltage. The peak diode voltage drop can then be read off the vertical scale. This peak diode voltage drop is then added to the detector output voltage to get the estimated peak ac input voltage.

The graph is quite regular, so I felt it would be very easy to model an equation to fit the graph. The best fit equation for the peak diode voltage drop is:

$$V_p = [(0.0716 - 0.0606 \times \mathrm{LOG}(V_S)] + [0.0863 + 0.0009 \times \mathrm{LOG}(V_S)] \times \mathrm{LOG}(V_o) + 0.0056 \times [\mathrm{LOG}(V_o)]^2 \qquad \text{(Eq 4)}$$

It should only be used over a range of detector output voltages, V_o, from 10 mV to 10 V and reverse saturation voltages, V_S, from 10^{-2} to 10^{-9} volts. Over this range it predicts the peak diode voltage drop to within a few millivolts. This equation is not effective for reverse saturation voltages that are greater than 10^{-2} volts.

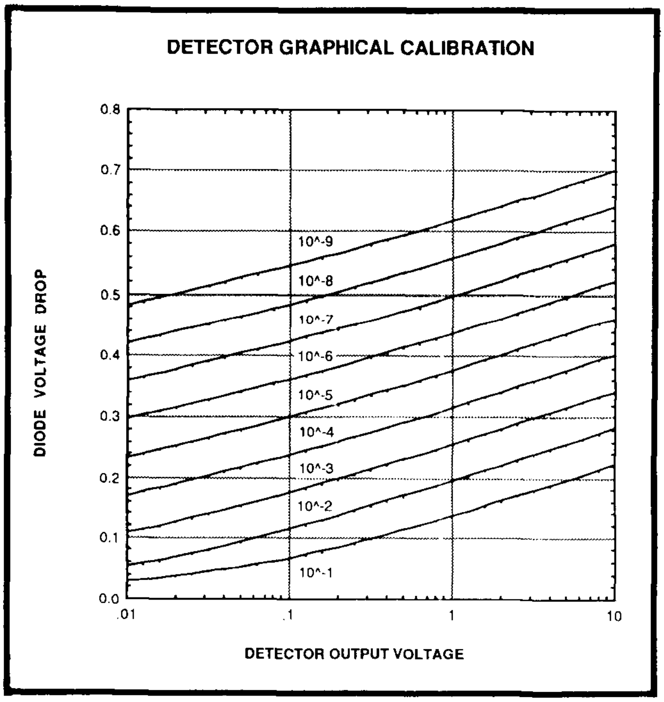

Fig 4—This shows the peak voltage drop across the detector diode as a function of the detector output voltage. The calculations were done for reverse saturation voltages of 10^{-1}, 10^{-2}, 10^{-3}, 10^{-4}, 10^{-5}, 10^{-6}, 10^{-7}, 10^{-8}, and 10^{-9} volts. The peak voltage drop is added to the detector output voltage to obtain an estimate of the peak ac input voltage.

In order to use this technique, one must determine the detector load resistance and the diode reverse saturation current. The detector load resistance is best determined by measuring it, since its value may not be obvious especially if a meter is used in the circuit. An ohmmeter needs to be connected across the detector load resistance, making sure that the voltage used to measure the resistance reverse biases the detector diode. If in doubt, either remove the diode or make the measurement twice, reversing the ohmmeter leads between measurements. The highest value will be the correct one. If an op amp or other semiconductor device is connected to the load resistor, it may also be necessary to disconnect it so that the semiconductor devices are not forward biased.

The diode reverse saturation current can either be estimated using Table 1 or can be directly measured with a relatively simple circuit shown in Fig 5. The measurement is made at a particular diode current which is set by R_S. The diode current for the measurement should be at roughly the same peak current at which the diode is used. The ratio of the peak diode current to the average current is roughly 10:1. For instance, if the detector load resistance is 1 MΩ and the detector output voltage is 1 volt, the average diode current will be 1 mA and the peak current would be about 10 μA. If V_b is 10 volts, then R_s should be set equal to 1 MΩ to yield a measurement current of 10 μA.

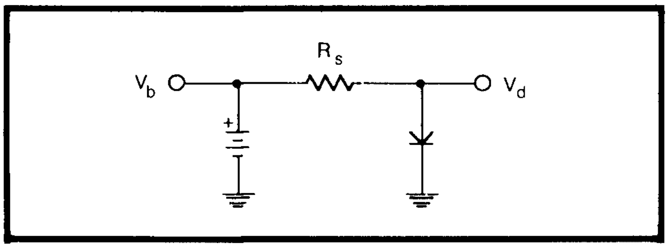

Fig 5—Circuit used to determine the diode reverse saturation current.

V_b may either be a battery or a dc voltage source. Once R_s is determined, V_b and V_d are measured with a high-impedance voltmeter. The voltmeter should have an input impedance of 10 MΩ or greater in order to avoid excessive loading of the circuit. The diode current and reverse saturation current may be computed by:

$$I_d = \frac{V_b - V_d}{R_s} \qquad \text{(Eq 5)}$$

$$I_s = \frac{I_d}{\exp\left(\frac{V_d}{0.026}\right) - 1} \qquad \text{(Eq 6)}$$

For greatest accuracy, the graphical correction should be done only at a measured reverse saturation current. However, even changes of 50% in its value won't have a big impact when using the graphical method of calibration. The two methods that follow this one avoid the problem of varying reverse saturation current by calibrating the detector without needing to know its reverse saturation current.

DC Voltage Source Calibration

A more accurate method to calibrate an individual diode detector can be accomplished using a dc voltage source in place of the ac source. This method takes advantage of the fact that the peak current through a diode detector is many times larger than the diode current for a dc voltage equal the peak ac voltage. A resistor, R_p, is placed in parallel with the detector load resistance in order to cause an increased current to flow through the diode. An adjustable dc voltage source is then used to replace the ac signal source and its voltage is adjusted to the peak voltage of the ac source at each calibration point.

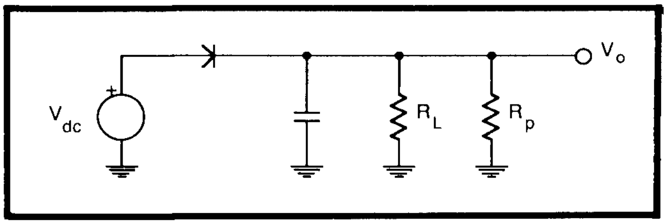

Fig 6—Circuit used for the dc substitution calibration method.

The method requires that the load resistance, R_L, be determined first. This may be done as described in the previous section. A resistor, R_p, is then temporarily connected across the load resistance. The value of R should be 15% of the value of R_L which will yield a calibration accuracy of $\pm$0.2 dB. If one is only interested in peak ac input voltages greater than 300 mV, then R_p should be set equal to 10% of the value of R_L and this will yield a calibration accuracy of $\pm$0.1 dB.

The RF end of the diode detector is disconnected from its source and connected to a dc voltage source. The dc voltage source is set equal to the peak ac voltage at each calibration point and the corresponding output voltage or meter reading is then measured. This reading will be equal to the value that would be obtained if the same peak RF voltage was present at the input of the diode detector.

Fig 7 shows the calibration error for several values of R_p as a function of the peak ac input voltage. The data was done for $V_s = 10^{-3}$ volts which is equivalent to a germanium diode with a load resistance of 10 kΩ or a Schottky diode with a 1-MΩ load resistance. The values of R_p are given in terms of the percentage of R_L. As can be seen from the graph, a value of $R_p = 10$-15% of R_L gives the minimum error. Also shown is the error for an uncalibrated diode which is significant even at a 10-volt peak ac input signal.

There is very little voltage out of the detector when the peak ac input voltage is small. For instance, in this case the dc output voltage will only be 70 mV when the

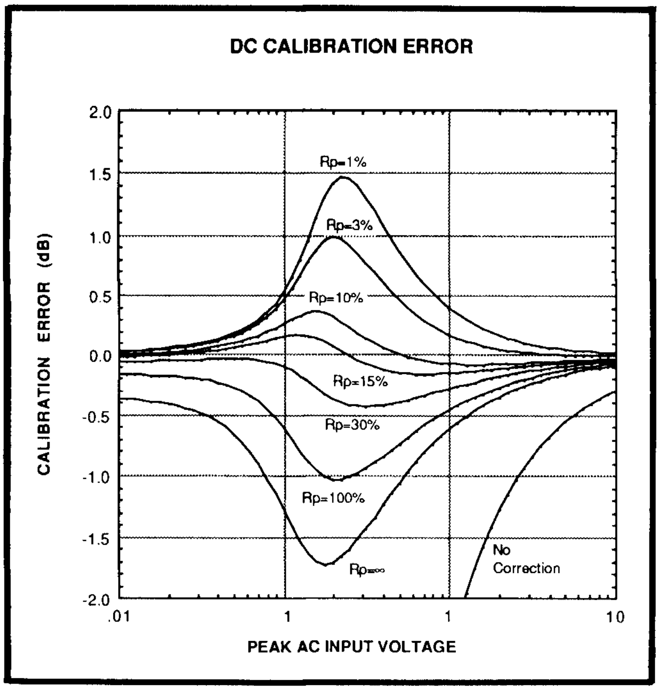

Fig 7—Measurement error for dc calibration method with $V_s = 10^{-3}$ volts. The values of R_p are given in terms of the percentage of R_L.

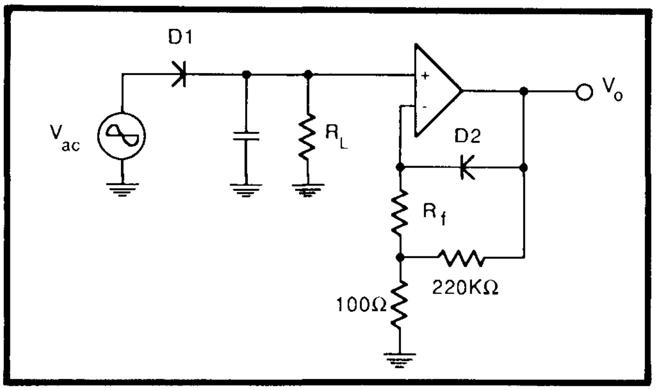

Fig 8—Circuit diagram of a linear detector whose output equals the peak ac input voltage. See text for details on the op amp requirements and the value of R_f.

peak ac input voltage is 0.1 volts. This small voltage may be difficult to measure. The circuit in the next section overcomes this limitation by using an op amp with non-linear feedback to increase the dc output voltage so that it will equal the peak ac input voltage over a wide range of peak ac input voltages.

Linear Diode Detector Circuit

A linear detector would be a circuit whose output was proportional to its input over a wide range of input voltages. The circuit in Fig 8 shows a method of constructing a circuit whose dc output is equal to the peak ac input[3]. This circuit is effective for voltages greater than 30 mV when constructed with 1N5711 diodes and a 1-MΩ load resistor.

D1 and D2 must be a matched pair of diodes in order to minimize the measurement error. They can be matched by measuring the value of the reverse saturation current using the circuit in Fig 5. The diodes should be matched to within 20%. The most critical point for matching the diodes occurs when the peak ac input voltage is between 30 and 100 mV. This implies a 10- to 100-nA peak diode current with 1N5711s and a 1-MΩ load resistance. For this case the diodes should be matched at a diode current of 1 μA since it is not feasible with amateur test equipment to measure the reverse saturation current at smaller diode currents. Data that I took on five 1N5711s shows that this is not an unreasonable requirement.

R_f is set in the range of 8-18% the value of R_L. This causes a larger current to flow through D2 than the average current that flows through D1. This larger current approximates the peak current that flows through D1, which means that the voltage drop across D2 is roughly equivalent to the peak voltage drop across D1. The voltage drop across D2 is added by the op amp circuit to the detected voltage which yields an output voltage that is roughly equivalent to the peak ac input voltage. The 220-kΩ and 100-Ω resistors limit the maximum voltage gain of the op amp to 2200 which makes it easier to adjust the op amp for a zero voltage output with no signal input.

The circuit places some stringent requirements on the op amp. The op amp voltage offset must be much less than the detected voltage. A Schottky diode with a 1-MΩ load resistor ($V_s = 10^{-3}$) will have a detected voltage of 400 μV for a 30-mV peak ac input signal. This means that the op amp input offset voltage must be significantly less than this. An offset voltage nulling circuit may be incorporated to accomplish this. Many op amps provide two terminals to connect a potentiometer for null adjustment. The offset voltage nulling circuit should be adjusted so that the op amp has zero output voltage with no input signal.

The op amp must also generate very little offset voltage from its input current. For this example, the op amp input current must be significantly less than 400 pA if the offset voltage generated by the op amp bias current is to be less than input offset voltage. The rest of the op amp parameters are not critical and there are a number of possibilities. I've used Texas Instrument's TLC271/272/274 series with success in this circuit.

Table 3 gives the recommended values for R_f that will minimize the measurement error. I've also included the estimated error range and minimum peak ac voltage that can be detected within this error range. The values of R_f are given as a percentage of R_L.

A graph of the measurement error is shown in Fig 9 for the circuit when used with Schottky diodes and a 1-MΩ load resistance. The error is plotted for various values of dc feedback resistance. From the curves, it is obvious that one should use a value of $R_f = 180$ kΩ for an overall minimum error and a value of $R_f = 100$ kΩ if

TABLE 3

V_S (Volts)	R_f (As % of R_L)	Maximum Error (dB)	Minimum Peak AC Voltage (mV)
10^{-1}	8.2%	±0.2	90
10^{-2}	15%	±0.3	40
10^{-3}	18%	±0.3	40
10^{-4}	18%	±0.3	50
10^{-5}	15%	±0.3	150
10^{-6}	12%	±0.2	250
10^{-7}	10%	±0.1	300
10^{-8}	10%	±0.1	400
10^{-9}	10%	±0.1	500

one wishes to minimize the error for input voltages above 300 mV.

Conclusions

This article has shown a number of methods for calibrating a diode detector without using expensive laboratory test equipment. The graphical method is useful when one wishes a fast answer without tremendous accuracy. The dc substitution method allows one to calibrate a diode detector with good accuracy using only a dc voltage source and accurate dc voltmeter. The linear diode detector provides a dc voltage output that is equal to the peak ac voltage input. The latter two calibration methods will retain a fair degree of accuracy even when the diode significantly deviates from the ideal diode equation or the input waveform is not sinusoidal.

With these methods, the average amateur should be able to know with a reasonable degree of certainty that they are correctly measuring an RF voltage. This knowledge will be useful in constructing power meters, directional wattmeters, or other circuits that require the magnitude of the RF voltage in order to make useful measurements.

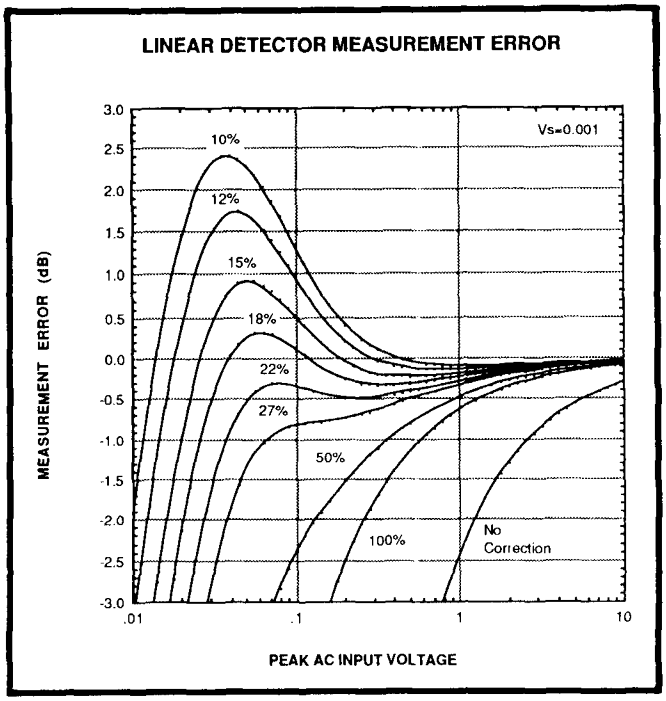

Fig 9—Measurement error for the linear detector circuit. Computations done for $V_S = 10^{-3}$. Measurement error is shown for both no correction and correction with feedback resistances given as a percentage of the load resistance.

References

[1]*ARRL Handbook,* 67th Edition, Chapter 4.
[2]Roy Lewallen, W7EL, private communication.
[3]John Grebenkemper, KI6WX, "The Tandem Match—An Accurate Directional Wattmeter," *QST*, January 1987.

Bits

LCA-1

Tatum Labs, Inc, announces a new release of their Logic Circuit Analysis program, LCA-1, a computer aided engineering (CAE) software package used by engineers for simulating digital circuits. This release adds Boolean expression definable components, internal node initial state definition and ATE interface capability. The product continues to be a cost-effective PC-based logic simulator, having 4-state logic simulation, with tabular and real time graphical output.

The **ATE Interface** offers "hot links" to other software so that the computed results can be compared to ARE checks running in real time. LCA-1 records every point within a circuit, together with every event, and allows data to be spooled to external applications.

Constant signals for ASIC designs—LCA-1 now supports permanently high and low signal lines permeating all modes. When using ASICs with redundant pins, high and low lines are managed via macros that mimic in-circuit behavior.

The **initial states of internal nodes** may be defined, eliminating any ambiguities and/or simulating an actual hardware configuration.

In addition to the convenient expandable primitive component syntax now available, LCA-1 will also accept **Boolean expression defined components** using either Boolean expression words or symbols.

LCA-1 runs on IBM PC/XT/AT and compatibles, requires 256 kbytes of RAM and MS-DOS 2.0 (or later). A one time update is available for current users, with multiple copy discount available. For a new installation, the cost is $495. A free demo is available on request.

For additional information or demo, contact: Fred Ebert, Tatum Labes, Inc, 3917 Research Park Drive B-1, Ann Arbor, MI 48108, tel: 313-663-8810.

Small Aperture IR Optical Links Using LED Light Sources

By Lawrence E. Foltzer
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In the March, April and May 1990 issues of *Ham Radio* magazine, Bob Atkins, KA1GT, described a long-distance (DX) atmospheric optical communications link using a Helium-Neon (HeNe) laser in the transmitter. In contrast, this article describes the capability of an optical communication approach based on the use of Infrared (IR) Light Emitting Diodes (LEDs), large area PIN photodiodes and small-diameter lenses.

Although it is true that lasers will outperform LED-based transmission systems in terms of distance (they can reach several 10s of miles), line-of-sight limitations and weather conditions can substantially reduce this capability. Simple IR LED-based systems, on the other hand, can reach distances of ½ mile or so, which will fulfill the requirements of many of the potential applications for optical communication links.

This article concentrates primarily on the basics of optical system design, ie, the characterization of LEDs for use in free-space links, and how to use manufacturers' data or your own LED characterization data to predict system performance and range limits. Three examples are provided to illustrate the range of performance that can be obtained when using these simple devices.

Advantages of LED-Based Systems

There are many advantages to using LEDs in optical communications links. One of the more important advantages is that they are offered in packages that include integral lenses that concentrate and focus their optical power output into beams with various degrees of divergence.

Another advantage of using LEDs is their relatively simple power-supply requirements. In most applications, LEDs will require less than 100 mW of power, permitting portable battery operation. HeNe lasers, however, require about 100 volts to operate and may consume a few watts of power.

LEDs also provide safety benefits when contrasted with the laser. A typical HeNe laser may emit a couple of milliwatts of optical power in a highly collimated beam whose intensity remains high and potentially dangerous at a significant distance from the laser. An LED, on the other hand, may emit a few tens of milliwatts, but from a relatively large surface area and with a high degree of angular divergence. The LED's large beam divergence causes the power density and potential to cause eye damage to fall off rapidly with distance. You can collimate the output of an LED with lenses to achieve longer system ranges, but to do so one must expand the beam diameter, which reduces the flux density at the output of the transmitter.

Modulation is much more easily accomplished for LEDs than for an HeNe laser, and can be accomplished directly, ie, without using external beam-modifying devices. The simplest example of the use of direct modulation is digital (on/off keying) modulation, where one simply turns the LED drive-current on and off in accordance with some preestablished transmission code protocol.

For analog transmission, three basic options exist:

1) One can continuously vary the LED drive current to follow the signals amplitude behavior (direct baseband modulation). This approach, however, is practical only for short distances or low fidelity transmission because LED linearity is impaired at high drive-current levels.

2) AM or FM subcarrier modulation: This method first places the data to be transmitted on an RF carrier using AM or FM modulation techniques. The modulated RF carrier signal is then used to amplitude modulate the LED. Since this modulation approach translates the baseband data frequencies to a higher frequency range, it facilitates a high degree of receiver noise immunity to low-frequency optical interference, eg, 60 Hz from incandescent lamps and dc overload due to high ambient lighting conditions.

FM-subcarrier modulation is recommended when a high degree of transmission linearity is required. This is because the data is encoded as a variation in RF carrier frequencies rather than as amplitude variation. High-quality fiber-optic video transmission systems commonly employ this approach.

3) Analog-to-digital (A/D) conversion followed by digital transmission as described previously.

Infrared LEDs for Communications

Infrared LEDs come in a wide range of colors, package styles and output power levels. The wavelength or color of a particular LED is dependent on its material composition. Typical emission wavelength regions are:

Table 1

LED Specifications

Device	*Vendor*	*WL (nm)*	*Angle/2*	*Intensity (Ie)*	*Package*
LD271	Siemens	950	30	>10 mW/sr @ 100 mA	T-1 3/4
LD274	Siemens	950	10	>30 mW/sr @ 100 mA	T-1 3/4
SFH484	Siemens	880	8	>50 mW/sr @ 100 mA	T-1 3/4
SFH485	Siemens	880	20	>16 mW/sr @ 100 mA	T-1 3/4
TIL38	TI	940	25	15 mW/sr @ 100 mA	T-1 3/4
TIL39	TI	940	10	35 mW/sr @ 100 mA	T-1 3/4
TIL905	TI	880	25	———	T-1 3/4

Table 2

PIN Diode Detectors for IR Communications

Device	*Active Area*	*Responsivity*	*Cap*	*Package*
BP104	2.2 mm × 2.2 mm	0.56 A/W, 880 nm	14 pF, 5 V	2 pin DIP
BPW34F	2.7 mm × 2.7 mm	0.56 A/W, 880 nm	20 pF, 5 V	2 pin DIP
SFH206	2.7 mm × 2.7 mm	0.56 A/W, 880 nm	20 pF, 5 V	Tall TO-92
TIL100	3.0 mm × 3.0 mm		28 pF, 5 V	Custom 1
TIL413	2.1 mm × 2.1 mm		15 pF, 3 V	Custom 2

940-950 nm for Silicon (Si)-doped Gallium Arsenide (GaAs) LEDs; 900 nm for Zinc-doped GaAs LEDs; and 790 to 890 nm LEDs made from various composition ratios of Gallium Aluminum Arsenide (GaAlAs).

Zinc-doped GaAs LEDs represent the oldest technology and the lowest power category of LED. They are, however, high-speed devices with typical turn-on and turn-off times in the 5- to 50-ns range. Because of their relatively low radiant intensity and more restricted availability, they will not be discussed further here.

The Si-doped GaAs LED family is also a rather old technology, but its high radiant intensity continues to make it a contender for use in remote-control applications. Perhaps its biggest drawback is its slow (300 ns to 1 μs) response time. For all but the most demanding applications, however, its speed is adequate.

The GaAlAs material system has the advantage that the emission wavelength can be shifted by changing the relative amount of Gallium versus Aluminum in the Ga(x)Al(1-x)As recipe. From this flexibility, faster, more-powerful IR emitters have become available, while maintaining a good spectral match to silicon detectors.

Table 1 lists some of the plastic-packaged LEDs that are available today that are good candidates for low-cost IR transmission links. The Table illustrates the dependence of an LED's radiant intensity on the beam-width/light-concentrating power associated with the molded-lens package.

The most important characteristic of LEDs for use in through-the-air links is their radiant intensity. Radiant intensity is expressed in terms of power per unit of solid angle (watts/steradian (sr)). Using this parameter, one can easily predict the maximum system range to be expected based on the "effective" cross-sectional area of the receiver. In addition, since LEDs behave linearly (assuming you follow the manufacturer's duty factor and pulse-width rules for high-current conditions), you can scale the LED's intensity in direct proportion to the increase over the reference drive current on the device data sheet. Examples of how to make range calculations are presented in the section labeled "The Optical Range Equation."

Silicon PIN Photodiodes

Table 2 lists some of the PIN photodiodes that are available today, and are intended for use in IR applications. (PIN refers to the device's physical structure, a layer of "I" [Intrinsic] (pure/undoped) semiconductor material sandwiched between layers of "P" and "N" doped materials.) What makes these devices specifically useful in IR links is that they all have built-in absorption filters designed to pass the light produced by IR sources, while rejecting unwanted optical wavelengths.

There are many advantages to using PIN photodiodes over other detector types, such as the phototransistor. Its performance is predictable from device to device so that performance is repeatable. It is easy to model for the purpose of predicting maximum transmission distance. And it provides the highest degree of flexibility in terms of the trade-off of bandwidth versus sensitivity. PIN diodes are also inherently low-noise devices, which means high-sensitivity receivers can be built using them, despite their lack of internal gain.

To operate the PIN photodiode effectively requires only a couple of volts of reverse bias and a series resistance to develop a signal in response to its photocurrent. The reverse bias forms a high electric-field region in the "I" layer of the device that separates and collects the charge carries created as a result of photon absorption.

The PIN diode is also easy to model (represent mathematically) for the purpose of range calculations because the size of its active area is normally provided on the manufacturer's data sheet. Its uncomplicated structure also makes it useful in characterizing the performance of LEDs (determining radiant intensity) for use in atmospheric links, so don't throw away those junk-box LEDs until you've read the section entitled "The Optical Range Equation."

The choice of which PIN diode to use is application dependent. Since all of these detectors are made of silicon, it is not surprising that they all have a similar responsitivity, so device selection boils down to issues like package style, active area and capacitance. If the range of your application is short, and the bandwidth is low, you may not need to use a lens to increase the effective collection area of the receiver, just pick a large-area photodetector. In most applications, though, the smaller the detector the better; the smaller device has less capacitance, which leads to greater bandwidth and receiver sensitivity. The only exception to the smaller-is-better rule is that it becomes more difficult to axially align the receiver to the transmitter (greater pointing accuracy is required) as the detector size diminishes.

The Optical Range Equation

The task of determining what distance can be supported between a receiver and transmitter can be easily determined by using Equation 1 when the source radiant intensity, receiver sensitivity and aperture are known.

$$d = \mathrm{SQRT}(Ie \times Ar / Pr) \qquad \text{(Eq 1)}$$

where

d is the maximum link range,
Ie is the source intensity in watts/steradian,
Ar is the effective receiver cross-sectional area, and
Pr is the receiver sensitivity.

To understand Equation 1's origin, the concept of source intensity must be understood. The units of intensity, watts per unit solid angle (the steradian), defines the total power that a light source emits into a standard-sized cone when placed at the apex of that cone. The size of this so-called standard cone could have been arbitrarily selected, but was standardized in a particular way to simplify the mathematics. In any case, if you envision an LED emitting into two cones of a particular angular size, where one is twice as tall as the other, you would find the flux density at the open end of the smaller cone to be four times that at the end of the taller cone, as predicted by Equation 2, which was obtained by rearranging Equation 1.

The angular size of our standard cone is one that subtends 1 steradian (65.5 ° total included angle). The area of the base of a 1-steradian cone is $1 / (4 \times \pi)$ that of the area of a sphere of the same radius. Noting that the area of a sphere is $4 \times \pi \times R^2$, and that the area of a cone that subtends 1-steradian angle is, by definition, $1 / (4 \times \pi)$ that of a similar sized sphere, it is easy to see that the area at the end of our standard cone is simply R^2. This leads to the conclusion that the flux density at distance R from a source emitting Pled watts into a 1-steradian cone is P_{LED} / R^2. The power intercepted by a receiver then becomes the product of the flux density at a distance R from the source, and the cross-sectional area of the receiver's aperture.

By rearranging Equation 1, you can determine what signal amplitude will be detected at a particular distance from the transmitter.

$$Pr = Ie \times Ar / d^2 \qquad \text{(Eq 2)}$$

where

Pr now represents the power that should be received at distance d.

Testing Equation 2's result against the receiver's sensitivity will serve as a guide in the design and optimization of the optical system.

By rearranging Equation 1 again, obtaining Equation 3, you can characterize and/or validate the performance of LEDs. Just make sure that the distance between the source and detector is great enough that the detector is in a region of uniform intensity distribution. (If your receiver's diode active area represents 1/100th of a steradian or less, your measurements will be accurate.)

$$Ie = Pr \times d^2 / Ar$$

where

Pr is the received power at a distance d,
d is the measurement system range, and
Ar is the cross-sectional area of the detector.

Experimental Transmitter Design

A schematic diagram of a transmitter circuit suitable for conducting optical-link range experiments is shown in Fig 1. The transmitter drives the LED with a low duty factor pulse of high current to increase source peak power and intensity, while reducing electrical power requirements.

U1, an LM311 voltage comparator, is used to make a low-frequency oscillator (7.7 kHz) that determines the pulse repetition rate of the transmitter. The RC network that couples the oscillator output to Q1 differentiates the oscillator output and provides the charge to drive Q1's base. The RC time constant of this network determines the pulse width of the output current, while resistor Rs, in the collector of Q1, sets the peak current drive to the LED. For pulse duty factors of < 10%, pulse widths less than a few microseconds, and drive currents less than ten times the maximum recommended dc drive current,

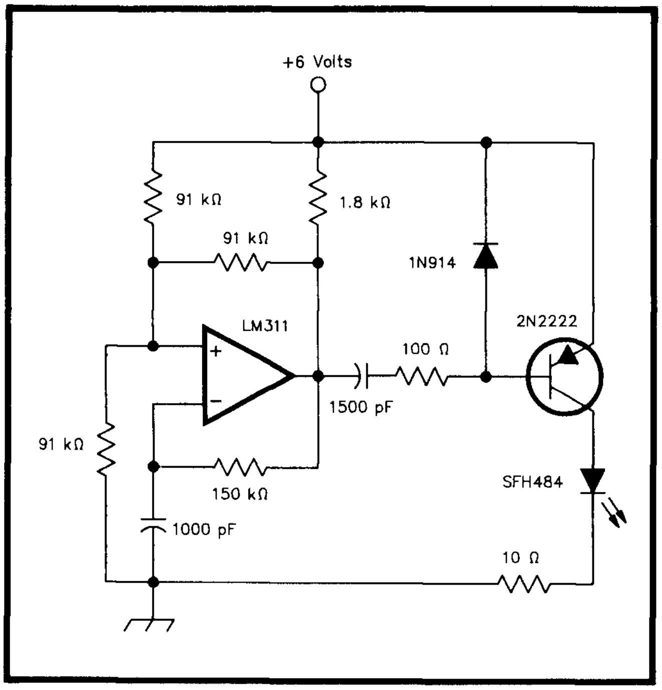

Fig 1—Experimental optical transmitter circuit

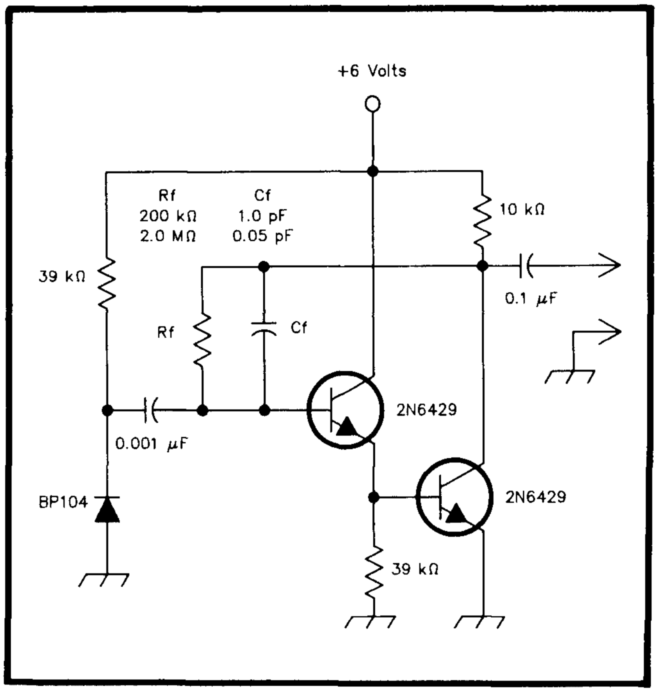

Fig 2—Experimental optical receiver circuit

you may linearly scale LED intensity in proportion to the LED's data sheet levels for use in the range equations.

Experimental Receiver Design

Fig 2 shows the schematic of the optical receiver used to conduct the experiments described later in this article. This receiver is basically a current-to-voltage converter that is often referred to as a transimpedance amplifier. This receiver preamp/front-end is best used by itself to make system performance measurements since the widest receiver dynamic range is obtained in this way.

The resistor labeled Rf on the schematic sets the gain, bandwidth, and sensitivity of the receiver. For all practical purposes, the transfer function (gain) of a wide-band/non-integrating transimpedance amplifier, when used with silicon detectors, will be on the order of Rf/2 volts per watt. Increasing the value of Rf will simultaneously increase the receiver gain and sensitivity while reducing the front-end bandwidth. However, as Rf is increased to optimize performance, the receiver will tend to integrate the signal and the transfer function will be less than predicted by the Rf/2 rule.

The bandwidth of this receiver is purposely made wider (Rf = 200 kΩ) than it would have to be for most of the applications listed above, so that the transmitted pulse shape and amplitude may be more accurately observed on an oscilloscope.

The sensitivity of a bipolar transistor-based transimpedance receiver, and used in this application environment, is predominantly limited by the input transistors base current dependent shot noise, and the dc photocurrent caused by ambient light interference. Without going into an extensive analysis of receiver noise, however, one can expect the receivers output noise level to be in the neighborhood of 1-mV RMS. For digital or FM subcarrier modulation schemes, low error rate (1 error per billion bits) performance is achieved for a peak-to-peak signal to RMS noise ratio greater than 12:1. Based on this information, we can arbitrarily define a 10-mV peak-to-peak signal amplitude at the receiver output as our detection limit for determining system range. Using these rules of thumb for receiver gain, noise and signal amplitude, we can use Equation 4 to predict the receiver's sensitivity (Ps) for error-free performance.

$$Ps = 0.02 / Rf \qquad \text{(Eq 4)}$$

Optical System Performance Experiments

In this section the experimental results for three optical systems configurations are presented. These results were obtained using the transmitter and receiver circuits described earlier. In all cases, the LED was driven at a peak current level of 280 mA. For consistency, only one LED, the SFH484, was used in these experiments to illustrate their broad application range.

Table 3 summarizes the results of the optical system-range experiments, and comments on each of these configurations are given in the following paragraphs.

Experiment 1

The optical system used in this experiment represents the simplest optical configuration in that the LED and detector did not employ external optics (lenses) to enhance transmission system gain. The transmitter

Table 3

Experimental Results

Experiment	*Transmitter Configuration*	*Receiver Configuration*	*Rf (kΩ)*	*Measurement Distance (ft)*	*Signal Strength (mV)*	*Calculated Range (ft)*
1	SFH484	BP104	200	9	10	9
2	SFH484	BP104 +	200	22	1000	220
		6-in FL, 2.1-in dia	200	100	46	214
			2000	100	180	425
3	SFH484 + 6-in FL, 2.1-in dia	BP104 + 6-in FL, 2.1-in dia	2000	200	1800	2680

used an SFH484 LED with its integral molded lens. The receiver used a BP104 PIN photodiode, which has an aperture that is approximately 0.1 inches on a side.

The 10-mV minimum signal was obtained at a distance of nine feet between the transmitter and receiver. Using Equations 3 and 4 and the PIN diode's data from Table 2, we determine the LED's radiant intensity to be 194 mW/sr at 280 mA, which scales to 37 mW/sr at 100 mA. This result, although about 25% lower than the LED manufacturer's specification, is reasonable given the uncalibrated nature of the homebrew equipment used to make the measurement, and that our range predictions are based on relative signal levels.

Experiment 2

In this experiment, the receiver photodiode (BP104) was used with a 6-inch focal length (FL) lens with a 2.1-inch diameter clear aperture, in an optical system configuration as shown in Fig 3.

One of the features of this optical system configuration is that it has a built-in telescope that facilitates pointing the receiver (or transmitter) at its target transmitter (or receiver). The objective lens forms an image of the target in the plane of the source or detector, which is also the focal plane of the eye lens. When looking through the telescope, one sees both the target and the source or detector in focus, and simply has to obscure the target with the source or detector to achieve alignment.

The field-of-view of the receiver for this optical configuration is determined by the size of the active area of the detector, and the focal length of the objective lens. The receiver's field-of-view is 1 square inch at 5 feet (0.048 °, half angle), which translates to a 40- × 40-inch spot at a range of 200 feet.

At a range of 33 feet, with Rf = 200 kΩ, a 1-V signal was detected. Based on this level of performance, a maximum range of 220 feet is predicted. To validate this prediction, a second experiment was performed at a range of 100 feet, where a 46-mV signal was measured. From this data, a maximum range of 214 feet is predicted, validating the range prediction method.

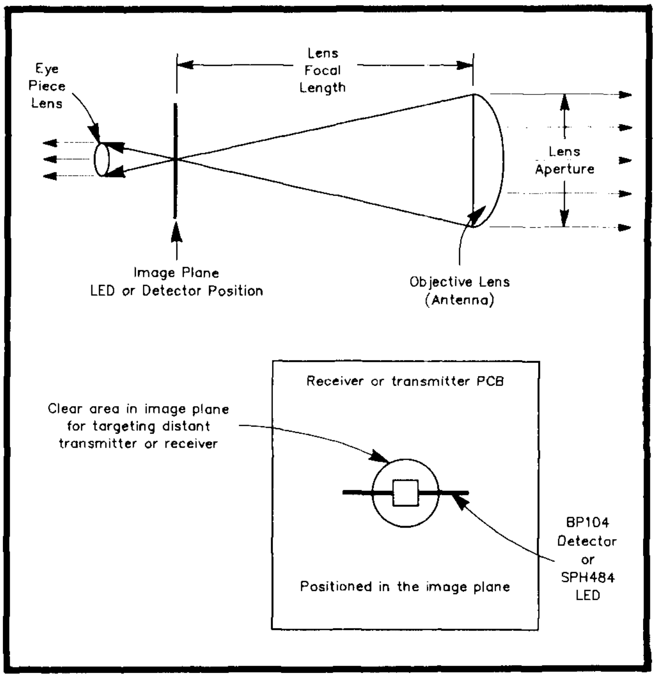

Fig 3—Long-range optical system

To demonstrate how receiver sensitivity is affected by the value of Rf, measurements were also made for Rf equal to 2.0 MΩ, at the 100-foot range. This configuration yielded a 180-mV signal, that extrapolates to a maximum system range of about 425 feet. Based on Equation 4, one would have perhaps expected a ten times increase in signal amplitude, in direct proportion to the increase in the value of Rf. However, the resultant decrease in receiver bandwidth does not allow the receiver output to follow the optical signal, which has the effect of reducing the apparent receiver gain. In addition, with Rf equal to 2.0 MΩ, the shunting effect of the 39-kΩ PIN diode-biasing resistor also comes into play, shunting away part of the signal current. The PIN photodiode biasing resistor might be increased, but at the expense of photodiode bias voltage in high ambient or background illumination. A decrease in bias will increase PIN diode capacitance, further reducing bandwidth and apparent gain.

Experiment 3

In this experiment the optical system used for the receiver in the previous experiment was applied to the transmitter. The initial experiments that were conducted used LEDs without integral lenses, but proved to be difficult to align due to the small projected spot size of the source (6.4- by 6.4-inch spot at 200 feet).

To relax the pointing accuracy of the transmitter to somewhat match that of the receiver, we have two options. One is to use a larger area LED, and the other is to add another element in the optical path to magnify the active area of more typically sized LEDs.

As it turns out, the effective size of the SFH484 LED chip is magnified to a size nearly as large as its package diameter by its own integral lens. You can verify the effective spot size by forward biasing the LED at about 20 mA (connect the LED to a 9-V battery in series with a 390-Ω, ¼-W resistor), and looking straight into it under subdued lighting. Hold the LED about a foot from your eye and slowly move it towards your eye. As you move it closer to your eye, you will see a dull red image of the LED chip and metallic contact pattern that is about as big in diameter as the package. The glow that you see is the low-level spectral component of the LED that falls into the visible portion of the optical spectrum.

As Table 3 shows, this transmitter configuration yielded the highest potential range (2680 feet) of all the configurations tested and was quite easy to boresight on the receiver. There was no attempt made to optimize the received signal strength through iterative transmitter pointing, so it would not be too surprising for one to obtain somewhat better performance than reported here. Short-range laboratory measurements of this transmitter configuration's radiant intensity yielded about 5.7 watts per steradian, which, from Equation 1, predicts a range of more than 2300 feet, which agrees well with the experimental results.

Corrigendum

As some of you have noticed, the references listed at the end of "A Compact 1-kW 2-50 MHz Solid-State Linear Amplifier," by H.O. Granberg, K7ES/OH2ZE, July 1990 *QEX*, were incomplete.

Here's a complete list of references:

References

The circuit boards and other components for this design are available from Communication Concepts, Inc, 508 Millstone Drive, Xenia, OH 45385, tel 513-429-3811/220-9677.

[1]Motorola, Inc, Semiconductor Sector Application Notes AN-749 and AN-1035.

[2]Hilbers, A.H., "Design of HF Wideband Power Transformers," Amperex (Philips) Application Laboratory Report ECO6907 and ECO7213.

[3]Blocksome, Roderick K., "Practical Wideband RF Power Transformers, Combiners and Splitters," *Proceedings of RF Expo*, February 1986.

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[5]Granberg, H.O., "New MOSFETs Simplify High Power RF Amplifier Design," *RF Design*, October 1986.

[6]Wakefield Engineering, 60 Audubon Road, Wakefield, MA 01880 and 7261 Mars Drive, Huntington Beach, CA 92647.

[7]Granberg, H.O., "MOSFET RF Power: An Update," *QST*, December 1982, January 1983.

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High Efficiency Power Amplifier SIMulator

Design Automation, Inc, announced a new computer program in the "High Efficiency Power Amplifier" series, HEPA-SIM, simulates high-efficiency (Class E) RF power amplifiers making it easier to design these desirable circuits. The program computes the steady-state periodic time-domain waveforms in a single-ended switching-mode RF power amplifier. It also computes the dc input power, RF output power, all of the power dissipations, and the RF output spectrum. HEPA-SIM can perform a series of analyses, sweeping any of the nine major circuit parameters, and plot the computed results vs the swept parameter. Waveforms can be plotted on the monitor screen and printed on a dot-matrix or laser printer.

The program can accept input from a companion program, HEPA-DESIGN (available soon) which designs a circuit to meet user-specified power and bandwidth requirements, and can pass its output to a companion program, HEPA-OPT (available soon) which optimizes a design automatically, to user-specified criteria.

HEPA-SIM requires an IBM PC/XT/AT/PS-2 or compatible computer with 385 kbytes of RAM. A floating-point numeric coprocessor is recommended, but not required. Prices range (in North America) from $595 to $995, depending on quantity, for a single-payment perpetual lease. Demo disk with all program function is available for $30, postpaid.

For further information contact Nathan Sokal, Design Automation, Inc, 809 Massachusetts Avenue, Lexington, MA 02173-3992; tel: 617-862-8998; fax: 617-862-3769.

QEX 1990 Reader Survey

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VHF + Technology

By Geoff Krauss, WA2GFP
16 Riviera Drive
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This and That

My sincerest apologies to the San Bernadino Microwave Society which I incorrectly identified, not once, but in both my April and June '90 columns, as the nonexistent "Santa Barbara" club; I just got my wires miscoded! The SBMS (ha!, no way I can get *that* acronym wrong, I hope!) is still active and sent along a copy of its newsletter containing several items of interest. In answer to a SBMS question, *I* think that it *is* a shame that *QEX* now has the only regularly scheduled VHF-and-up technical column (and even that's bimonthly) of the established American ham sources; but the *QEX* management tells me they would be happy to receive good VHF+ technical items (articles or notes) for future publication.

I also received, at about the same time, several issues of the newsletter of the North Texas Microwave Society, a very active group in the Dallas/Fort Worth area. There is a wealth of good information in the late '89/early '90 issues, especially in an article for building a 5760 SSB transverter (this changes my comment in the June 1990 column about what state-of-the-art is with respect to the availability of a US-originated 6-cm transmitter design). Now the question is whether boards/kits/units will become available for this design. This is especially germane as I know that a large number of 3.4- and 5.6-GHz QSOs were made last weekend, in the June '90 contest, using microwatt output sources (eg, diode multipliers) over 60+ miles, and a 10-mW 5760-MHz QSO was done over a 170+ mile path (of course a 3-foot dish was on one end and a 6'er with a good LNA was on the other). Nevertheless, SOTA is again advancing, even as I write.

However, even if this column is on time (being scheduled to appear in the August '90 issue), I may be late in writing the next column or two at the normal time for two reasons: Eye surgery (maybe, but not certainly, exacerbated by possible exposure to leakage of really high-power microwave stuff back in my ECM-design days) was quite successfully done on June 25, but will keep my vision poor for about eight weeks; and then a job transfer/move to the Philadelphia area will keep me even busier (note that in looking around for a new job I considered: Dallas, Texas [the NTMS locale]; Beaverton, Oregon [home of the Tektonix Radio Club 47-GHz builders]; and north Philadelphia [near to the Mt Airy VHF Club, possibly the largest such Club in North America]—my wife isn't quite sure what criteria I'm using in my job hunt, but I bet many of my readers understand. In any event, I'm looking for affiliation with a club that can use some help in mounting a true contest challenge of the W2SZ/1 group, who appear to finally have broken the 1-million-point mark in an ARRL VHF contest.

From the stories I've heard, it appears that the big contest battles are now between W2SZ/1 in western Massachusetts and N6CA in California; Chip Angle's bunch appears to be as big and well-equipped as the Easterners', but apparently lost this June '90 'test because awful weather discouraged some of their microwave rover stations. In contests where almost as many 220- to 432-MHz QSOs are made as on 50 MHz, and more 2.3- or 3.4-GHz contacts than on 902 MHz, the importance of the rover is heavily felt. Many of you may have received a polling on rover/contest rule change questions from the *National Contest Journal*; please answer. These questions were passionately argued at the 1990 Northeast VHF/UHF Conference at Nashua, New Hampshire, in mid-May; the overwhelming majority of those present appear to favor expanded use of rovers, microwaves and a 3-person/3-band limited multiop class. Many attendees apparently did something about their feelings as the total number of microwave contacts made in the East (and especially those between new grid combinations) was very large in the recent contest. A particularly heavy increase was noted at 2.3/3.4/5.6 GHz, and there was much activity at 10 and 24 GHz.

State-of-the-Art in Our 10- and 24-GHz (and up) Bands

As always, my first move in determination of amateur SOTA was to check my boxes of articles for these bands. Unbelievably, in the last ten years there has been more published for 10.0-10.5 GHz than for all of 902-5925 MHz combined! Not much has been written for 24.0-24.25 GHz, and almost nothing for 47.0-47.2 GHz (perhaps because this now-narrow band will not allow relatively unstable FM-modulated oscillators, nor will the second harmonic of a 24-GHz Gunn oscillator fall within the band, as it did when the band was 48-50 GHz). Outside of a brief mention of some work I had done several years ago, there was nothing about our 76-81 GHz band, and there was definitely not one thing for our 119.98-120.02 GHz band (of course, if you can figure out how to generate a signal at this frequency in the first place, much less one stable to within 15 ppm at this sub-millimeter wavelength, you don't need any articles!) or our 142-149 or 240-250 GHz bands. To say nothing of all that space above 300 GHz. Hi, hi, hi! Naturally, I leave out laser communications because it is really an entirely different form and not *really* micro/millimeter wave work.

If there are so few people working on 3456 and even fewer on 5760, can there be very many 10 giggers? Yes, indeed. Just look at contest results; there are twice as many 10-gig (short for gigahertz, of course) users as 3.4 or 5.6. In fact, you may find about as many 24 giggers as 5.7 giggers! Why? Because most 10/24 GHz work is done using duplex wide-band FM transceivers, predominantly the M/A-COM Gunnplexer® units (available from ARR in Connecticut). This unit has a waveguide-cavity Gunn-diode oscillator, with mechanical tuning over the whole band (500 MHz on 10 and 250 MHz on 24); a varactor for electronic tuning/modulation is provided; as is a waveguide mixer, placed in front of the oscillator. A waveguide isolator is placed in front of the mixer. A portion of the transmitted signal is reflected back into the mixer by a screw placed between the isolator and the mixer, and serves as LO for down conversion of the received signal. A good low-noise IF preamp and simple FM IF strip complete the rig. QSOs over many miles (under line-of-sight, or LOS, conditions) can be made with the 17-dB horn antenna normally supplied with the unit. Other 10-GHz FM transceivers exist, mostly formed from Doppler radar units intended for intrusion alarms; these units are generally not as sensitive as the Gunnplexers, but may cost much less. Separate transmitters and receivers, such as the dielectric-resonator/GaAsFET units available from Mitsubishi, can be used with a wave-guide transfer switch (if you can find one in the surplus market), and offer the possibility of adding power/low-noise amplifiers, filters, etc. The possibility of locking the unit oscillators to a crystal standard has been covered in numerous articles, but is rarely done (probably because the degree of difficulty is the same as building a narrow-band CW/SSB unit instead).

One comment must be made about the WBFM units: You can get the police awfully mad at you if you are not careful in the use of these units! The X-band speed radars use a frequency of 10.525 GHz, which is just above the amateur band. These Doppler radars are not particularly well filtered in their font ends, and 10-GHz amateur signals can do nasty things to the units. There is no illegality here, as long as you are safely within the amateur band, but try proving that to a nonscientifically educated cop. Even more ominous is the fact that the Ku-band police radar operates at 24.15 GHz, right in the middle of the shared band. Even though the range is less, the potential for interference is greater; true, it is a risk that FCC regulations say all users should be aware of and must tolerate, but try convincing a local judge of that some Saturday afternoon during a Summer contest! To make matters even worse, a number of people have actually used ham rigs to jam police speed radars; it is not too difficult to get the radar to display false speeds, if you know how. It is for reasons such as these that older, obsolete radar units, which police used to auction off, or just plain give away, to amateur microwavers, are now trashed; another victim of over-doing it! It might be mentioned here that speed radar units *can* be meaningfully used for long (50+ miles, using the internal antenna) paths; the RF head is mated with a modulator and IF strip, after the speed detecting electronics are removed. How's that for equipment that's not particularly reliable over distances of a few hundred feet in its normal surveillance role?

Antennas for these bands are limited; horns (rectangular, circular, and compound) are used by themselves or as feeds for parabolic dishes (perhaps with a hyperbolic subreflector, if you are lucky enough to be able to build/find a Cassegranian system). Surfaces are almost always solid as gain is affected by openings of more that $\frac{1}{4}$ λ, or surface irregularities of more than about 1/10 λ.

More susceptible of innovation is design of SOTA weak-signal equipment; only one CW/SSB transverter is available commercially—from SSB Electronics GmbH (via Gerry Rodski, K3MKZ, 124 Cherrywood Drive, Mountaintop, PA 18707). The LO/Tx/Rx module set produces about 0.1-W output Tx and a NF listed as "about 2.5 dB," for $600+. There is a WB5LUA dual-GaAsFET amplifier unit (from Down East Microwave) which can be used for a slightly better receiver noise figure, but any higher transmit power requires that you build your own.

As with other bands, receiving converters are superheterodyne mixers; the 10-GHz mixer, preferably, is preceded by an image-rejection filter after an 'LUA preamp (or two). Surplus filters, mixers, couplers and the like, are still the main source of subassemblies at 10 gigs, and almost totally so at 24 GHz or higher. It was true, up to a very few years ago, that the only decent 3-cm receiving converter was the G3JVL wave-guide design, with a cavity and post form of input filter followed by a cavity mixer, after a cavity and post multiplier/filter, with low-side LO injection. If you have access to engineering equipment and techniques, better receive converters can be built using active mixers or harmonic diode mixers, with microstrip techniques and filtering. Of course, North American VHF+ers await a truly reproducible design.

As with other microwave bands, the biggest problem is still the generation of a highly stable signal above 10 GHz. Most solutions center around the use of surplus parts, or entire oscillator multiplier chains from so-called "brick" sources to which one applies power, and possibly an external frequency-controlling signal (if the brick source does not have an internal crystal/oscillator) and can obtain appreciable 10-12 GHz power output; changing frequency becomes only a matter of changing the crystal to one at a new frequency and ordered from any one of the several suppliers advertising in the usual ham magazines. Doubling of a 13-GHz signal to 24 GHz is difficult, but not unknown. Generation of any amount of transmit power is almost always by TWTs or solid-state devices such as power GaAsFETs. Expect to see more experimentation and publications on these bands in the future!

Conducted by: Stan Horzepa, WA1LOU
75 Kreger Drive
Wolcott, CT 06716-2702

ARRL/CRRL 9th COMPUTER NETWORKING CONFERENCE

The 9th Computer Networking Conference, jointly sponsored by ARRL and CRRL, will be held Saturday, September 22 in London, Ontario, Canada. The conference hours are 9 AM to 5 PM and the conference site is the London Regional Art Gallery and Museum, 421 Ridout St N in downtown London, overlooking Harris Park at the forks of the Thames River. There is adequate free parking nearby.

A registration fee of $US 20/$CDN 25 includes a copy of the conference proceedings and a catered hot lunch.

London, Ontario, population 270,000, is located in Southern Ontario midway between Detroit, Michigan, and Buffalo, New York, and is accessible by car via Highway 401, rail or air. While no major airlines have direct service to London, commuter service to Toronto and Detroit is provided by Air Ontario (Air Canada) and Canadian Partner (Canadian Airlines International). ComAir (Delta) provides connector service from Cincinnati and Cleveland. For those in the US that might want to take in a bit of countryside, flying to Detroit, Buffalo, or Niagara Falls and renting a car for the 2 to 3 hour trip to London can be a cost-effective alternative to flying directly. Check with your travel agent for details.

Conference organizers have negotiated a special flat rate of $CDN 85 a night (no limit to the number of people allowed to stay in one room) at the 322-room Radisson Hotel, London Centre, located about four blocks from the conference site. It is highly recommended that conference participants stay at this hotel to facilitate organizing Friday-night dinners and informal get-togethers. (A list of alternate accommodations can be furnished on request.) Conference participants must make their own reservations at the Radisson. Use the toll-free number 800-333-3333 and mention the conference.

Past computer networking conferences have attracted 120 to 150 participants from all over the US, Canada and beyond. Conference speakers share the results of recent work at the leading edge of packet radio. All participants hear all speakers as there are no concurrent presentations. Although this is not the place to find out how to get into packet radio, if you are a beginner and you do attend, you are certain to develop an enthusiasm for this wonderful mode.

Since this is a conference and not a hamfest, there is nothing to look at or buy, but that doesn't preclude a few interesting displays and demonstrations or the deal of a lifetime made in the parking lot.

Also, conference organizers will make arrangements so that everyone who wishes can have dinner together and a night out at a popular restaurant.

To register, send $US 20/$CDN 25 to:

9th Computer Networking Conference
c/o Harry MacLean, VE3GRO
500 Riverside Dr
London, ON N6H 2R7
Canada

Include your name, address and call sign, if any. You will receive confirmation in the mail, along with maps and additional information related to the conference.

MICROSAT STATUS REPORT

AO-16

AO-16 is in good shape. RAM memory tests will be done in preparation for uploading the BBS software. According to Harold Price, NK6K, at the present time, the BBS software for AO-16 is targeted to be installed and operational by August. Until that time digipeating operation will be possible.

DO-17

DOVE is still transmitting on S-Band only. Full recovery has been delayed due to an unexplained anomaly in the S-Band transmitter. Bob McGwier, N4HY, has to rely on the S-Band transmitter because the 2-meter transmitter blocks the 2-meter receiver on DOVE. This anomaly has required N4HY to use his considerable programming skills to write special digital signal processing software to work around this problem. Another problem plaguing the DOVE recovery is that N4HY's hastily assembled S-band station is about 5-dB SNR lower than what is required for reliable reception. N4HY has been able to use the high quality S-Band station of Bill McCaa, KØRZ, in Boulder, Colorado, via phone patch during mutually visible passes for software loading. N4HY has given the DOVE recovery top priority. On July 7, DOVE had again crashed as evidenced by Microsat Bootloader transmissions from the S-Band beacon. Analysis was under way to determine whether the crash was due to a problem in one of the tricky software upload sessions or, less likely, another hardware problem.

WO-18

WO-18 is in good shape. The third version of the imaging software was uploaded last week. The Weber State University engineers have added more "tuning knobs" to the picture taking software in order to improve the contrast of the earth images. Additional information has been added to the image header giving the solar array currents and horizon sensor values when the picture was snapped. These added items will provide more information about where the lens was pointed when the picture was taken. The impact detector, which was originally designed to indicate when WO-18 was hit by micrometeorites, is being used to verify that the CCD camera iris shutter has clicked. When the shutter is snapped, the impact detector instantaneously sees the vibration that the shutter operation causes in the spacecraft structure.

Each day several images have been downloaded and those who have WeberWare 1.0 can turn the binary data into pictures on CRT screens. Finally, the 1.2-GHz fast-scan TV receiver was tested and the horizontal sync was detected. A full image is expected to require more gain and/or much more accurate antenna pointing.

One good quality picture of earth features has finally been taken and will be transmitted every evening for the next several days so that WeberWare 1.0 users will be sure to have a chance to capture it. Other testing will continue each day.

LO-19

LO-19 is in good shape. The RF power output of LO-19 is thought to have been varying because of the simultaneous operation of the CW beacon. Once the RAM memory-checking software has been finished and checked out on AO-16, the same RAM memory check will be performed on LO-19. Eventually the AMSAT-LU Group will take over all the software development for LO-19 after the BBS software is installed. At the present time all general housekeeping and monitoring tasks are performed by the AMSAT-LU team.

from AMSAT

MICROSAT BBS PROGRESS REPORT

Harold Price, NK6K, who is working in close collaboration with Jeff Ward, K8KA, at the University of Surrey (UoS) on the Microsat BBS software, provided the following report concerning its progress.

It was decided early on to use UO-14 as a test bed for all Microsat BBS development. NK6K points out that there are several advantages to this. First, all software developed for UO-14 is completely portable to the Microsats due to a device programmers call high-level application programming interfaces (API). APIs hide the differences between the different computer architecture of the Microsats and UO-14. APIs perform the nitty-gritty tasks that are peculiar to the hardware of a particular computer system. For example, one of the most common tasks performed by all computer systems is input/output (I/O) operations. Thus, to the programmer it makes little difference if the BBS software he is writing is for the Microsats or for the Packet Communication Experiment (PCE) on UO-14, as long as he uses the APIs.

Another advantage for using UO-14 as a test bed is the 9600-baud uplink/downlink that it uses. NK6K says that he can load new software in two passes as compared to eight passes for the Microsats (the Microsats use 1200 bauds). Also, UO-14 has a completely separate computer system for the PCE operations. The UO-14 on-board computer (OBC), which controls the satellite, is unaffected by what happens with the secondary PCE computer. So, if the BBS software crashes, UO-14 will continue to operate normally. If the BBS software causes a crash of the OBC on a Microsat, a great deal of software needs to be reloaded. Although an OBC crash doesn't put a Microsat at great risk, it is inconvenient to the users. Except for DOVE, most software reloads of a Microsat OBC are not very difficult to accomplish.

from AMSAT

SAREX STATUS REPORT: MINI-HAMFEST IN SPACE

When space shuttle mission STS-37 is launched there will be a pile-up on both ends of the QSOs. When the crew for STS-37 was chosen, Ken Cameron, KA5EWP, was the only licensed amateur on the list. In May, Mission Specialist Jay Apt passed the Technician Class license test. Recently, Ken upgraded to General Class and Mission Commander Steven Nagel and Mission Specialist Linda Godwin passed the Technician Class. Meanwhile, the entire shuttle fleet has been grounded by NASA until further notice.

from AMSAT

PACKET-RADIO LINKS ON 23 CM

This story describes the efforts being made to design a 23-cm packet-radio link by Geff Mather, G8DHE, a member of the Worthing and District Video Repeater Group (WDVRG).

The desire to develop a 23-cm radio link suitable for packet radio came about from the groups interest in putting on a PBBS (GB7VRB) and node (BR2:G6WØR). As a site already existed, using 23 cm for a video repeater, it was thought that a 2-meter, 70-cm and 23-cm node would be useful to cover a similar area (Brighton to Chichester, coastal strip) as that served by the video repeater. Also, since there was considerable concern by others that video and packet radio would mutually interfere with each other on the same band, we wished to test this and, if possible, prove that this would not be the case.

The group already has a successful design for a 1-watt, PLL-controlled transmitter suitable for medium to wide bandwidths, which is ideally suited for use on packet radio. Since a lot of amateur TV stations employ a wideband down converter feeding a suitable video IF and demodulator, it is believed that if a suitable IF could be designed for packet-radio use, then the two IF strips could work in tandem following a PLL-locked version of the down converter. This would have the advantages that both modes could be used together and the cost of additional front ends and aerials at 23 cm would also be eliminated.

To simplify the IF and to enable a cost efficient design, the TDA7000 radio chip was considered. This chip has all the elements required to receive a signal in the range of 2-110 MHz and provide a demodulated signal O/P with an IF bandwidth of 150 kHz and good demodulated frequency/phase response up to about 30-40 kHz. This would be more than adequate for a 9600-baud or higher packet-radio link. To this end, a basic design was put together (with PCB) which has been proven to work successfully within the shack. However, no filtering was included in the initial design between the down converter and the IF, resulting in a horrendous S/N ratio! The next stage of the design is to put a roofing filter at the front of the IF to improve this factor.

The modulation of the 1-watt transmitter at a bandwidth of only 150 kHz (normal bandwidth for video being 12 MHz or more) has proved quite practical. This bandwidth was chosen as it would enable bauds in excess of 9600 to be easily achieved.

It should be noted that the overall cost of the system is not massively less than a commercial 23-cm rig, however, it can be used on both packet radio and video. Also, expansion to more than one channel is very easily achieved in a variety of ways depending on intended purpose.

It is intended that once a full working system has been set up and tested for a month or so, then WDVRG will make the IF kits available in the same manner as the transmitter kit and add-on PLL unit that are already available.

by Geff Mather, G8DHE
from *Connect International*

PACKET-RADIO FREQUENCY COORDINATION?

The rapid emergence of packet radio has produced some variety of problems. It is easy for those of us involved with voice repeater coordination to compare packet radio to our own chosen mode. However, this simple comparison overlooks some very substantial differences. The greatest similarities, I believe, is the network system that provides so much of packet radio's utility. Of necessity, these stations must be in operation on known frequencies. It is the more casual packet-radio operator who poses problems because of the available agility of changing frequency with changing conditions of band use. It is also this class of operator who finds little utility in belonging to packet-radio organizations. A by-product of this situation is that it has been difficult for packet-radio interests to consolidate into state-wide organizations comparable to repeater councils. Because of the differences in modes, the model of coordination developed for FM repeaters over a 30-year period is not directly applicable to packet radio. Additionally each packet-radio station is, in a sense, a potential repeater. Packet has experienced an explosive growth over the past few years. The similar growth of FM repeaters over a long period of time has allowed for some consolidations which the packet-radio service, at least in many cases, seems unable to achieve. Coordination à la FM hardly seems the answer for packet radio.

The failure of state packet-radio operators to assemble into state-wide organizations has lead some repeater councils to attempt some form of coordination. Obviously the registration of every packet-radio station is impossible. Certainly it is desirable to collect and publish data on those stations which are assembled in networks. In a few cases, the efforts of repeater councils to assemble and publish this data has been rebuffed and considered an intrusion into packet-radio affairs. The need seems to exist and the efforts of FM coordinators will probably continue until the packet-radio interests consolidate their activities. It will be unfortunate if the efforts are taken as intrusive instead of one of assistance and support. As a well-organized service, the FM repeater community is in a better position to offer this service than packet-radio interests at this time. As the packet-radio service matures and becomes capable of supporting state-wide organizations, adopts rational systems of frequency assignment, they will assume the duties that are presently being offered by some repeater councils. This should not become a cause for warfare between the services. Packet should remain free to develop its own practices and assignments and its own frequency structure. It must also develop some method of discipline which corrals maverick packet-radio operators into defined frequencies away from interference with established repeaters. Until the packet-radio interests are able to organize on a state-wide basis, OARC, and I believe other FM coordinators, will be willing to offer support and assistance without intrusion into the legitimate activities of the packeteers. The two services are mutually disruptive when operation on the same frequencies is attempted. Cooperation between the modes should result in a desirable growth of packet-radio organizations comparable to repeater coordinators in a far shorter time span. Obviously, this is going to require the quieting of internal competition of packet-radio groups within each jurisdiction. As an external observer, I say, the sooner the better.

by George Waldie, W8JRL
from *Repeater Coordinators' Newsletter*

PACKET RADIO AND SSTV OK'D IN USSR

Beginning 1 March 1990, Soviet "shortwavers" are authorized to use packet-radio communication (paketnaya radiosvyaz) and slow-scan television (televideniye s medlennoy razvertkoy, SSTV) by a decision of the State Telecommunication Inspectorate of the Ministry of Communication of the USSR.

The right to operate packet radio and SSTV is given to collective and individual radio stations of the 1st category, and no special permission need be obtained. When exchanging information by packet radio, it is necessary to use the AX.25 protocol and ASCII code (KOI-7 code, set Ho [No?] with an additional noninformation bit). On shortwave, frequency-shift keying of 170 or 200 Hz and an information transfer rate of 300 bit/s are used; on ultra shortwave (VHF/UHF/SHF), frequency modulation, a shift of 1000 Hz and a speed of 1200 bit/s are used.

On shortwave, the following frequency segments for packet radio and RTTY operation have been established: 1838-1842 kHz, 3580-3620 kHz, 7035-7045 kHz, 10140-10150 kHz, 14070-14099 kHz, 18100-18110 kHz, 21080-21120 kHz, 24920-24930 kHz and 28050-28150 kHz. On ultra shortwave, the following segments have been allotted for these operating modes: 144.65-144.675 MHz and 433.3-433.325 MHz.

For SSTV operation, the international standard is used, based on a mains frequency of 50 Hz. Here are the basic characteristics of SSTV according to this standard.

Frame duration - 7.2 s
Number of lines - 120
Line duration - 60 ms
Frame-synchronizing-impulse duration - 30 ms
Line-synchronizing-impulse duration - 5 ms
Synchronizing frequency - 1200 Hz
Black-color frequency - 1500 Hz
White-color frequency - 2300 Hz
Frame format - 1:1 (square)
Horizontal scanning direction - left to right
Vertical scanning direction - top to bottom.
Required passband - 1000-2500 Hz
Class of emission - J2F.

On shortwave, the following segments for SSTV operation have been established: 14225-14235 kHz, 21335-21345 kHz and 28675-28685 kHz.

edited by Aleksandr Ivanovich Gusev, UA3AVG
from *Radio*, translated by Dex Anderson, W4KM

STRAY BITS

According to a reliable source, the long-awaited PS-186, a high-speed multiport packet switch, will be available from AEA in August.

C PROGRAMMERS' GROUP

Any programmers, professional, hobbyist, or aspiring, interested in starting a C language users group which could meet via packet radio, should send the following information to:

Michael Bilow, N1BEE @ KA1RCI.RI.USA.NA
Forty Plantations
Cranston, RI 02920

Name and call sign, mailing address, telephone number (indicate if you want it listed on the roster), preferred packet-radio address, a rough idea of your experience with programming in general and with the C language in particular (advanced, intermediate, beginner, just curious), and special interests (telecommunications, C++, assembly language interfacing, etc.)

***GATEWAY* CONTRIBUTIONS**

Submissions for publication in *Gateway* are welcome. You may submit material via the US mail or electronically, via CompuServe to user ID *70645,247* or via Internet to *70645.247*@compuserve.com. Via telephone, your editor can be reached on evenings and weekends at 203-879-1348 and he can switch a modem on line to receive text at 300, 1200 or 2400 bit/s. (Personal messages may be sent to your *Gateway* editor via packet radio to WA1LOU @ N1DCS or IP address 44.88.0.14.)

The deadline for each installment of *Gateway* is the tenth day of the month preceding the issue date of *QEX*.