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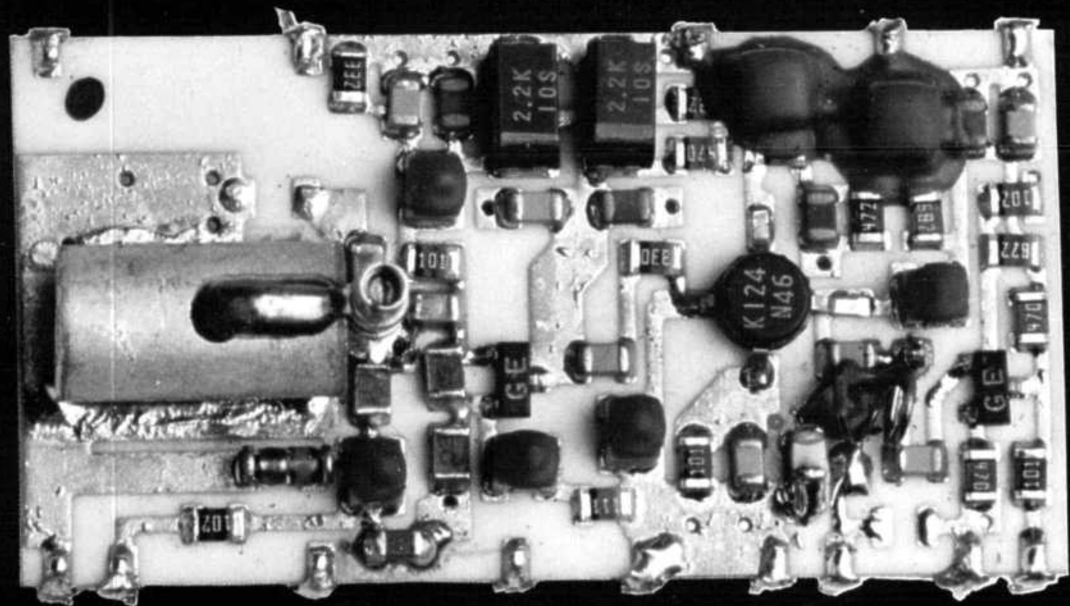
QEX

ARRL Experimenter's Exchange

October 1994



Low Phase Noise—How to Achieve It

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About the Cover:
This 900-MHz VCO based on a ceramic resonator exemplifies a low-phase-noise design, as KA2WEU describes.

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Features

Columns

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Purpose of *QEX*:

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

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and correspondence for publication in *QEX* should be marked: Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled 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

Tooling Up for RF

Previously in this space we have decried the lack of low-cost design and development tools for amateurs. While multi-thousand dollar tools abound, they are hardly accessible to the amateur in the street, so to speak.

Well, cheer up. ARRL has taken the initiative to address this problem in one area: RF circuitry. Working with Compact Software, of Paterson, NJ, ARRL will soon make available *ARRL Radio Designer*. This *Windows* program performs linear analysis of circuits. That may at first sound like a description of *PSPICE*, or *MicroCAP*, both of which are linear circuit simulation tools available at low cost (in their simpler versions, that is), but *ARRL Radio Designer* is different. What makes it different is that it really does RF, although it will do low-frequency circuits, too.

One of the frustrating problems with trying to use SPICE-type simulators at RF is that they don't speak the language. S-parameters, Y-parameters, H-parameters, group delay, reflection coefficient, VSWR, return loss, noise parameters—these are part of the *lingua franca* of both *ARRL Radio Designer* and human radio designers.

And *ARRL Radio Designer* includes capabilities usually found only in high-end programs. Want to optimize your preamp design for noise figure? Let *ARRL Radio Designer*'s optimizer find the best set of input matching circuit components to do the job—than ask it what the resulting input VSWR will be. Want to know if you really need 1% resistors in that critical part of the club-project circuit you've designed? *ARRL Radio Designer*'s statistical processing can tell you how many of the club's copies of the circuit will be unstable if you use 5% components.

There are limitations, of course. As a *linear* circuit simulator, *ARRL Radio Designer* can't provide information about distortion products or large-signal operation of circuits. But it can help you design and optimize the linear circuits used in your RF designs. Most of the circuits in a radio system are linear, after all. And, as KA2WEU shows us this month, *ARRL Radio Designer* can sometimes assist in designing even a large-signal circuit such as an oscillator.

ARRL Radio Designer's reports are available in both graphical and tabular form, as part of its *Windows* user interface. That means you can see the results on the screen right away, and you can print the resulting graphs or tables of calculated values.

You can read more about *ARRL Radio Designer*, including some sample applications, in "Introducing *ARRL Radio Designer*: New Software for RF Circuit Simulation and Analysis," October, 1994, *QST*.

This Month in *QEX*

Low phase noise is the holy grail of 1990s radio design. To achieve it, you need to be "Designing Low-Phase-Noise Oscillators." Dr. Ulrich L. Rohde, KA2WEU, discusses the design procedure and shows how modern computer circuit simulation tools simplify the problem tremendously.

Part 2 of "Practical Microwave Antennas," by Paul Wade, N1BWT, covers dish antennas—how to build them and feed them.

Need an "Inexpensive PC A-to-D" capability? Gary C. Sutcliffe, W9XT, shows how the PC's game port can serve as a simple way of measuring an external resistance, such as a thermistor.

In this month's "Digital Communications" column, Harold Price, NK6K, expands on his previous discussion of what's holding back amateur packet development.

This month we begin a new column: "Proceedings." As the various amateur technical conferences take place, we will print lists of the papers available in the conference proceedings. Our hope is that this will help you locate those papers that can help you with your experimental efforts.—*KE3Z, email: jbloom@arrl.org (Internet)*

Designing Low-Phase-Noise Oscillators

Find an oscillator design that ensures low phase noise using modern computer programs.

By Dr. Ulrich L. Rohde, KA2WEU

As the other characteristics of communication equipment improve with technology, the stability of the signals generated in systems becomes increasingly important. Most often, we use oscillators to generate these signals. Thus it becomes important to determine how best to ensure stability in oscillator designs. This is particularly true of voltage-controlled oscillators (VCO), which typically use relatively low-Q resonators (compared to crystal oscillators, for example).

Oscillator performance can be considered to be composed of both short-term and long-term stability. Short-term stability is what we refer to as *phase noise*, and has been described in detail previously.[1] In this article, we will investigate means of designing oscillators that achieve low-phase-noise performance. The trend in modern design is toward low power and small size, driven in part by the use of hand-held, battery operated equipment. Power consumption influences noise performance because the ultimate signal-to-noise ratio at frequencies far removed from the oscillator frequency is determined by the output power—and thus by the power consumed. Physical size is a limiting factor on the Q of the resonator, which is partly responsible for determining close-in phase noise; larger resonators can achieve higher Qs. This article details methods of optimizing oscillator designs based on these principles. We also briefly look at the advantages of using ceramic resonators at UHF.

[1]Notes appear on page 12.

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Oscillator Noise

The SSB phase noise of an oscillator is well described by the Leeson model and its enhancements. Leading engineers such as M. Driscoll of Westinghouse and Tom Parker of Raytheon have expanded the model to include the flicker-noise component of even passive components.[2,3] The enhanced Leeson equation, as proposed by Parker and others, is shown in Fig 1.

Numerous circuits have been developed to implement oscillators. Fig 2 shows the most common RF oscillator circuit configurations. These circuits have various advantages and disadvantages, depending on the frequency of operation and the resonator type. For circuits in the 400 to 2000-MHz range, modern oscillators tend to use transmission-line resonators and capacitive feedback of the Colpitts or Clapp type. At these frequencies, bipo-

$$S_\phi(f_m) = \left[\alpha_R F_o^{\,4} + \alpha_E \left(F_o/(2Q_L)\right)^2\right] \Big/ f_m^{\,3} + \left[(2GFKT/P_o)\left(F_o/(2Q_L)\right)^2\right] \Big/ f_m^{\,2} + \left(2\alpha_R Q_L F_o^{\,3}\right) \Big/ f_m^{\,2} + \alpha_E / f_m + 2GFKT/P_o$$

Fig 1—The Leeson equation for oscillator phase noise. *G* is the compressed power gain of the loop amplifier, *F* is the noise factor of the loop amplifier, *K* is Boltzmann's constant, *T* is the temperature in Kelvins, P_o is the carrier output power, in watts, at the output of the loop amplifier, F_o is the carrier frequency in Hz, f_m is the carier offset frequency in Hz, Q_L is the loaded Q of the resonator in the feedback loop, and α_R and α_E are the flicker noise constants for the resonator and loop amplifier, respectively.

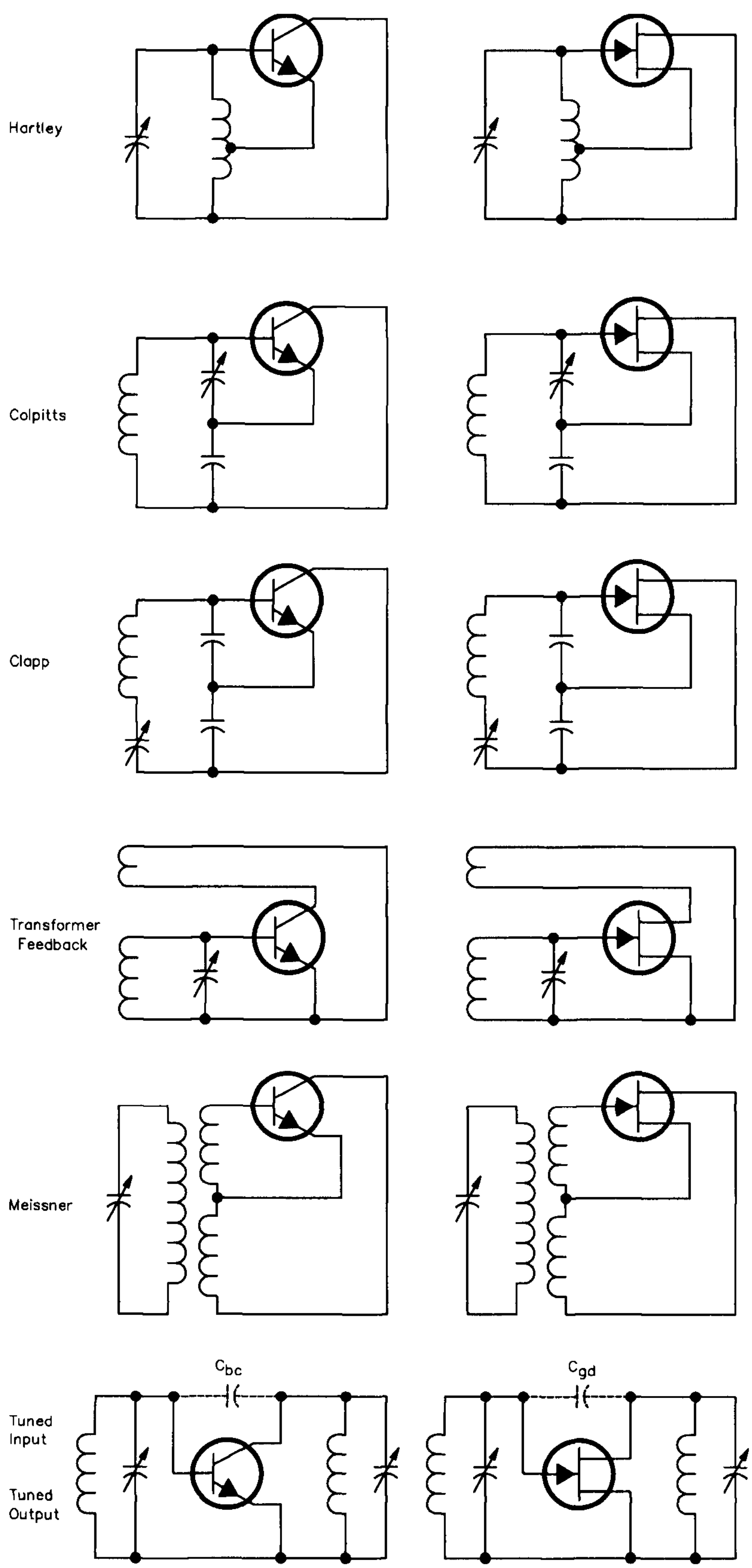


Fig 2—Common oscillator circuits for bipolar transistors and FETs.

lar transistors are generally used since few FETs have sufficient gain-bandwidth product for use in UHF oscillator circuits.

Requirements for Low-Noise Oscillators

The key elements that determine the phase noise of an oscillator are:

- the transistor's flicker-noise corner frequency, which depends on the device current;
- the loaded Q of the resonator, which depends on the coupling between the resonator and the transistor; and
- the ultimate signal-to-noise ratio, which depends on the RF output power of the oscillator and its large-signal noise figure.

Of these, the first two can be considered using linear analysis of the circuit. But the active device's large-signal operation requires nonlinear analysis techniques, without which we can make only educated approximations of the ultimate signal-to-noise ratio.

The Linear Approach

The design goal of the linear approach is to achieve the maximum loaded Q of the resonator and to keep the bias (dc) device current to a minimum. A high Q helps restrict noise components to frequencies close to the frequency of oscillation, minimizing phase noise as we move away from that frequency. The requirement for minimum bias exists because the flicker, or $1/f$, noise of the device is highly dependent on the current. Table 1 shows the flicker-noise corner frequency versus collector current for a typical bipolar transistor. JFETs have much less flicker noise than bipolar transistors, while GaAs FETs have more.

Oscillator Operation

At start-up, the oscillator's open-loop gain must be sufficient to begin

Table 1—Flicker Corner Frequency vs Collector Current for a Typical Bipolar Transistor (from Note 6)

I_c (mA)	F_c (kHz)
0.25	1
0.5	2.74
1	4.3
2	6.27
5	9.3

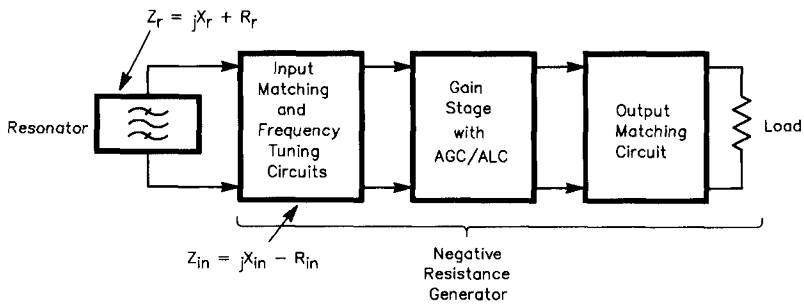


Fig 3—An oscillator viewed as a resonator and a negative resistance generator. At start-up, the resonator and oscillator reactances must be equal in value and opposite in sign, while the sum of the resonator and oscillator resistances must be less than 0. For sustained oscillation, the sum of the resistances must not become positive.

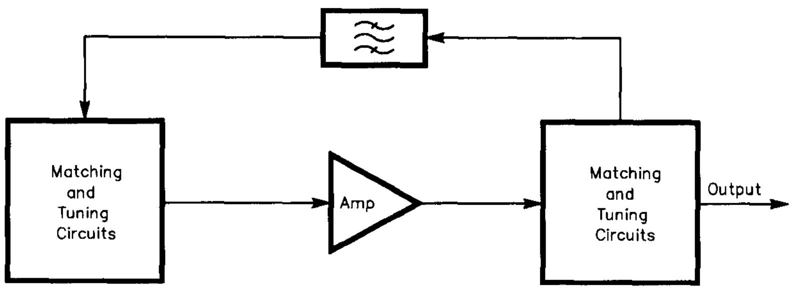
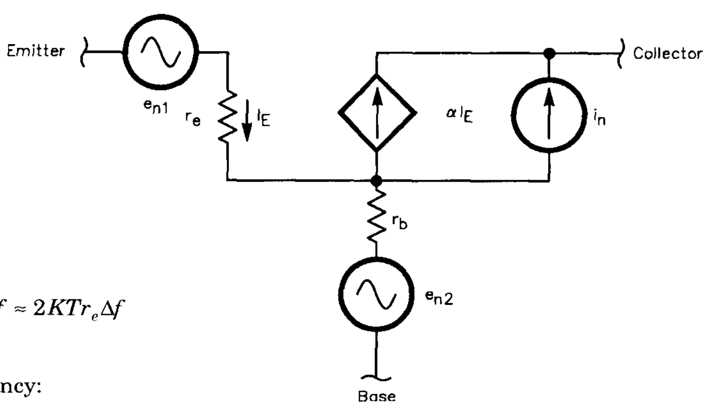


Fig 4—An oscillator viewed as a feed-forward amplifier with positive feedback through the resonator. Start-up of oscillation requires that the gain of the amplifier exceed the loss of the resonator and that the total phase shift through the amplifier and resonator be a multiple of 360°. To sustain oscillation, the phase shift must remain the same and the amplifier gain must be equal to or greater than the resonator loss.



$$\bar{e}_{n1}^2 \approx 2qI_E r_e^2 \Delta f \approx 2KTr_e \Delta f$$

$$\bar{e}_{n2}^2 \approx 4KTr_b \Delta f$$

At high frequency:

$$\bar{i}_n^2 \approx 2_q \left[\alpha_F I_E \frac{1-\alpha_F + f^2/f_o^2}{1+f^2/f_o^2} + I_{ce} \right] \Delta f$$

Fig 5—A simplified bipolar transistor noise model using white-noise sources.

oscillation. The circuit's amplitude stabilization mechanism is responsible for sustaining oscillation. We can view the oscillator as a two-terminal negative-resistance generator, as shown in Fig 3. Here, the total resistance—the sum of the resonator resistance and the resistance of the two-terminal oscillator—must be less than or equal to zero for oscillation. The net reactance will be zero at resonance. A different view is presented in Fig 4, where the oscillator is treated as a feed-forward amplifier with positive feedback. We analyze this model by finding the gain (or loss) and the phase shift of both the amplifier and the feedback network. The product of the amplifier and feedback network gains must be greater than 1, and the total phase shift should be 360° or some multiple thereof.

Although we can't precisely analyze the large-signal operation of a circuit using wholly linear techniques, we should recognize some effects that will impact our linear analysis. Chief among these is bias shift. The large signals present in the circuit in a bipolar oscillator will cause a shift in the bias current, because of the nonlinearity of the base-emitter junction. The device current may be about 10% different from the nominal (no-signal) current and may shift in either direction (more current or less). Since the flicker-noise is bias dependent, this effect is important to keep in mind.

The recommended approach to finding the bias-dependent loading of the resonator by the active device is to construct a linearized model of the device using its measured S-parameters at a particular bias point. This is especially important at higher frequencies. For simplicity, we have chosen not to do this in the example that follows, but to use a simple model.

Over a wide range of current, the device f_t remains constant. Since:

$$f_t = \frac{1}{2\pi R_d C_e}$$

where R_d is the emitter diffusion resistance and C_e is the emitter capacitance, and since:

$$R_d = \frac{26 \text{ mV}}{I_E}$$

(at room temperature), where I_E is the emitter bias current, we can therefore adjust the R_d and C_e parameters of our device model to reflect the bias current we expect to use. This will allow our linear circuit model to reflect the bias dependency of the oscillator.

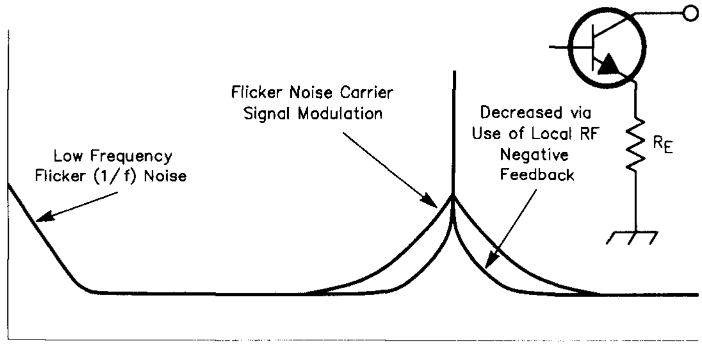


Fig 6—Adding negative feedback can reduce the amount of AM-to-PM modulation of the carrier by the transistor's flicker noise.

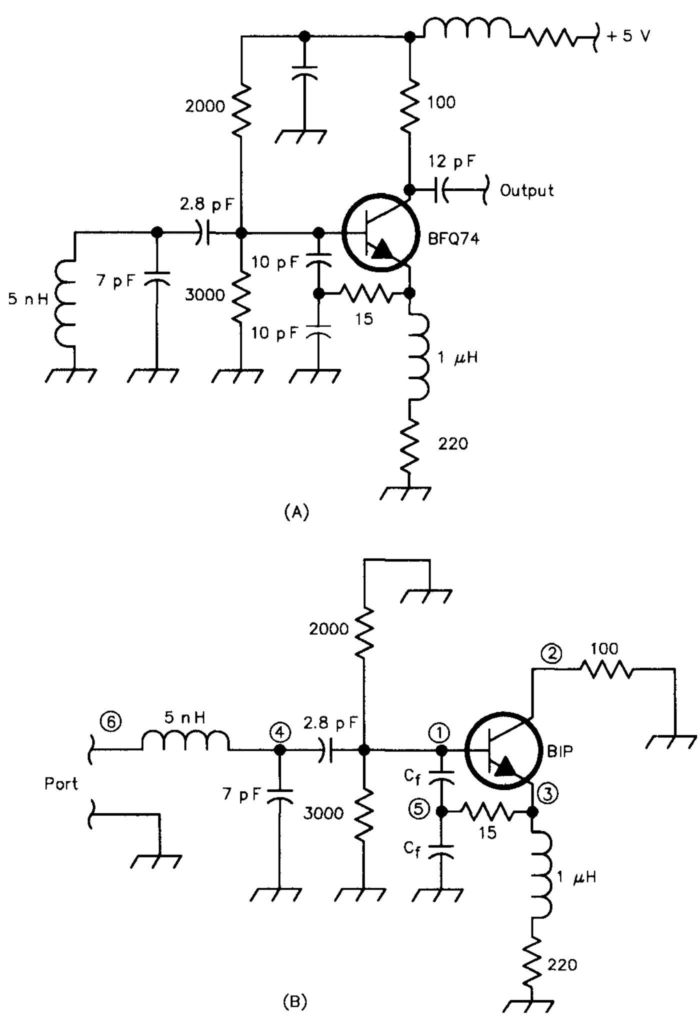


Fig 7—A simple 800-MHz oscillator circuit. The schematic is shown at A, and the circuit model for simulation is shown at B. The circuit model treats the circuit as a 1-port device, in order to investigate the impedance at the resonator.

As mentioned before, flicker noise is dependent largely on the bias current. But the *effect* of flicker noise can be reduced. This noise contributes to the phase noise by modulating the oscillator's frequency via AM-to-PM conversion. We can reduce this modulation by use of negative feedback. A simplified noise model of a transistor is shown in Fig 5. The effect of applying negative feedback to reduce phase noise is shown in Fig 6.

Linear Analysis by Simulation

By far, the easiest way to perform the linear analysis to optimize an oscillator circuit is via linear circuit simulation. Simulators such as *ARRL Radio Designer* (a subset of Compact Software's *Super-Compact* linear simulator) can provide a look into the oscillator circuit, allowing us to adjust circuit parameters to achieve the optimal circuit for the devices being used.[4]

A simple oscillator circuit is shown in Fig 7. This 800-MHz oscillator uses a Siemens BFQ74 bipolar transistor. Looking at this circuit from the standpoint of the negative-resistance generator of Fig 3, we analyze the net resistance of the resonator (L1) and the oscillator circuit. The resistance should be 0 or slightly negative.[5] We most easily view this by breaking the circuit at the ground connection of the coil and treating that point and ground as the terminals of a 1-port circuit, so we can investigate its impedance. The circuit model is shown in Fig 7B.

We want to select a bias current that is as small as possible, without reducing it to the point where the oscillator output power is too small to give a useful ultimate signal-to-noise ratio. In this case, we selected a bias emitter current of 10 mA. Now we have to find the appropriate feedback network, consisting of C1 and C2. Varying these capacitance values will vary the feedback and thus the loading of the resonator by the oscillator circuit. Finding the point at which the net resistance of the modeled circuit is just negative gives us the proper feedback; at this point the loading is the least that will sustain oscillation, and the loaded Q is therefore the highest available with the selected bias.

Fig 8 shows the circuit netlist used with *ARRL Radio Designer* to simulate the circuit of Fig 7B, and Fig 9 shows the port resistance (RZ11) and reactance (IZ11) calculated by the simulation. The traces labeled 1 are for values of C1 and C2 of 5 pF. Traces 2 are with C1 and C2 at 10 pF, and traces 3 are at 25 pF. What we are looking for here is the resistance at resonance, where the reactance is zero. For trace 3, this occurs at about 735 MHz. Here, the resistance is almost exactly zero. This allows no room for component tolerances or for adjusting the frequency upward, either of which may inhibit oscillation. Trace 2 shows a better result. At resonance, about 760 MHz, the resistance is negative, and it stays negative up through above 1 GHz. Small variations in component values, or adjusting the frequency upward, should not keep the circuit from oscillating. Trace 1 might seem to be even better because the resistance is more negative, but now we are loading the resonator—and lowering the loaded Q—more than we need to. This is shown in Fig 10, which shows the magnitude of the impedance for the same three cases.

Even though the resistive part of the impedance is negative, we can use the magnitude of the impedance to determine the loaded Q. From Fig 10 we can find the impedance at resonance, then find the 3-dB points on the curve, by multiplying the resonant impedance by 1.414. The loaded Q is then the resonance frequency divided by the 3-dB bandwidth. (This is more easily found by outputting the data of Fig 10 in tabular form.) It's obvious from the graph that the Q of trace 1 is lower than that of trace 2. For low phase noise, therefore, trace 2 is a better choice. Setting C1 and C2 to 10 pF results in certain oscillation and good phase noise.

Finally, the linear approach can be extended to further optimize the circuit. The L/C ratio of the resonator could be reduced, reducing coupling and increasing the loaded Q. If we had an ideal FET, we could theoretically achieve the performance predicted by the Leeson model and graphed in Fig 11.[6]

VCO Noise

So far, we've considered only the noise from the transistor. When we extend our design to become a VCO, by adding a tuning diode, we must also consider the phase noise introduced by that diode. Contrary to what you may have read elsewhere, this noise is not due solely to Q reduction from the added diode. The diode itself introduces noise that modulates the VCO frequency. The easiest way to analyze

```
*
*  ARRL Radio Designer 1.0
*  COMPACT SOFTWARE, Inc.
*  Copyright (C) 1988-1994. All Rights Reserved.
*
*  SIEMENS BFQ 74 BPT: Analysis of an NPN bipolar model.
*  bias network is set for Ie=10mA
*****************************************************
Cf:10pF         ; Tapped-capacitor feedback network
IE:10mA         ; Emitter current
FT:6000e6       ; Device ft in Hz
Rd:(26mV/IE)
Cte:(1/(2*PI*FT*Rd))
*****************************************************
BLK
  BIP 1 2 3 A=0.98 RB1=4 CE=Cte RE=Rd
* Feedback network
  CAP 1  11 C=Cf
  CAP 11 0  C=Cf
  RES 3  11 R=15
* Emitter feedback network
  SRL 3  0  R=220 L=1UH
* Tank circuit
  CAP 1  4  C=2.8PF
  IND 4  31 L=5NH F=800MHZ Q1=120
  CAP 4  0  C=7PF
* Collector decoupling
*   CAP 2  0  C=1NF
    RES 2  0  R=100
* DC bias network
  RES 1  0  R=2000
  RES 1  0  R=3000
  OSC  :1POR 31
END
*
************************
* FREQ: Frequency block *
************************
FREQ
  STEP 500MHZ 1000MHZ 10MHZ
  STEP 700MHz 850MHz 2MHz
END
***********************************
```

Fig 8—An *ARRL Radio Designer* netlist that describes the circuit model of Fig 7B.

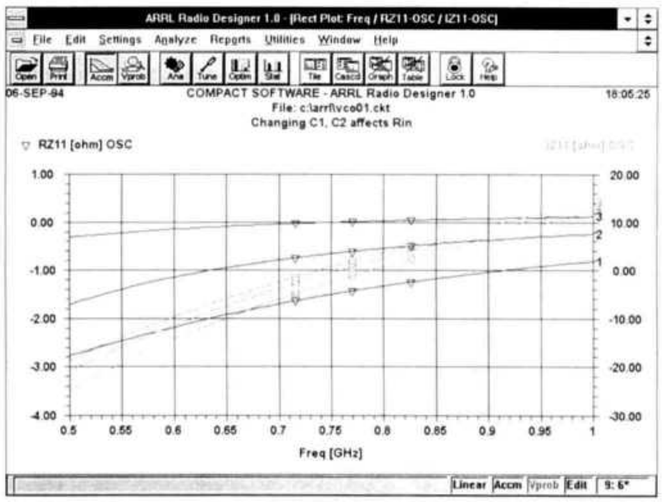


Fig 9—This *ARRL Radio Designer* analysis shows the resistive (RZ11) and reactive (IZ11) parts of the impedance seen at the port of Fig 7B. A negative resistance at resonance ensures oscillation. Trace 1 is with C1 and C2 at 5 pF, trace 2 has them at 10 pF and for trace 3 they are set to 25 pF.

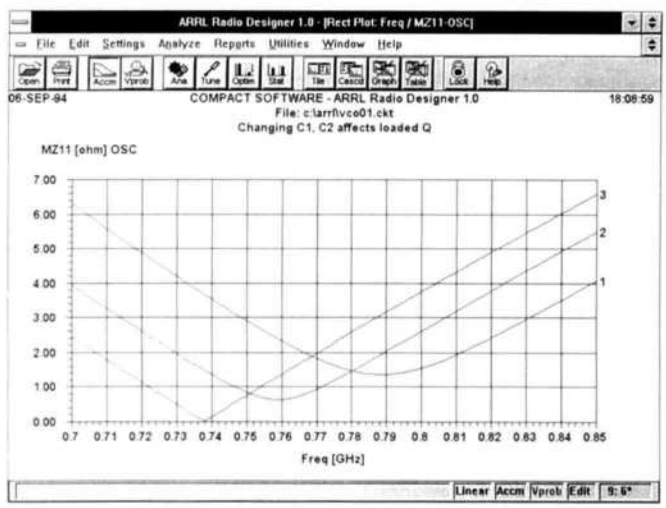


Fig 10—The magnitudes of the impedances plotted in Fig 9 are shown here. The reduction in Q as the feedback capacitors are decreased in value is dramatic.

this noise is to treat the diode's noise contribution as that of an equivalent resistance, R. This resistor can then be considered to be generating the thermal noise voltage that any resistance exhibits:

$$V_n = \sqrt{4KT_0R\Delta f} \qquad \text{Eq 1}$$

where V_n is the open-circuit RMS thermal noise voltage across the diode, K is Boltzmann's constant, T_0 is the temperature in Kelvins, R is the equivalent noise resistance of the tuning diode, and Δf is the bandwidth we wish to consider. At room temperature (about 300 K), KT_0 is 4.2×10^{-21}.

Practical values of R for tuning diodes range from about 1 kΩ to 50 kΩ. For a value of 10 kΩ, for example, we would find a noise voltage from Eq 1 of:

$$V_n = \sqrt{4\times4.2\times10^{-21}\times10{,}000}$$
$$= 1.265\times10^{-8}\ \text{V}/\sqrt{\text{Hz}}$$

This noise voltage from the tuning diode modulates the frequency of the oscillator in proportion to the oscillator's VCO gain, K_0, (the frequency swing per volt of the tuning signal):

$$(\Delta f_{rms}) = K_0 \times (1.265\times10^{-8}\,\text{V}) \qquad \text{Eq 2}$$

in a 1-Hz bandwidth. This can be related to the peak phase deviation, θ_d:

$$\theta_d = \frac{K_0\sqrt{2}}{f_m}(1.265\times10^{-8}\,\text{rad}) \qquad \text{Eq 3}$$

in a 1-Hz bandwidth, where f_m is the frequency offset of the noise from the oscillator operating frequency. Applying a typical VCO gain of 100 kHz/V gives a typical peak phase deviation of:

$$\theta_d = \frac{0.00179}{f_m}\,\text{rad} \qquad \text{Eq 4}$$

in a 1-Hz bandwidth. For an offset of 25 kHz, as might be used to find the noise in an adjacent FM channel, this gives $\theta_d=7.17\times10^{-8}$ rad in a 1-Hz bandwidth. Finally, we can convert this result into the SSB signal-to-noise ratio at the specified frequency offset:

$$L(f_m) = 20\log_{10}\frac{\theta_d}{2} = -149\,\text{dB/Hz} \qquad \text{Eq 5}$$

This kind of noise performance is typical of a high-end laboratory signal generator such as a Rohde and Schwarz SMGU or a Hewlett-Packard 8640. It's interesting to note that the resonators used by these two products are slightly different. The SMGU uses a helical resonator while the 8640 uses an electrically shortened quarter-wave cavity. Both generators are mechanically pretuned, allowing use of a relatively small electrical tuning range, at about 100 kHz/V, for FM and AFC purposes. It's worth noting, too,

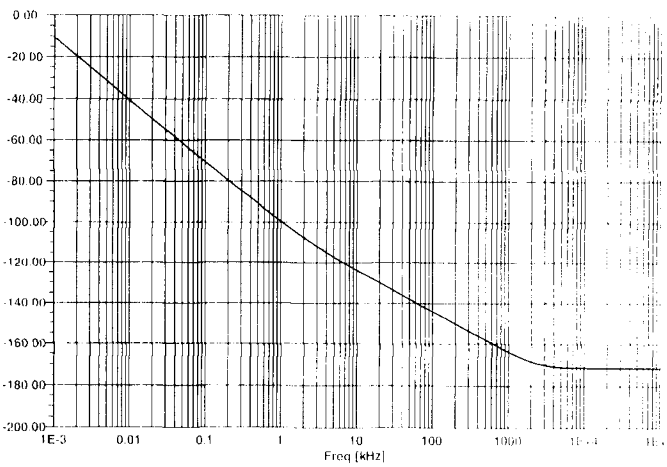


Fig 11—The Leeson equation, plotted for an ideal FET, showing phase noise in dBc/Hz versus frequency offset from the carrier. Note that the flicker-noise corner frequency is apparent at about 1 kHz.

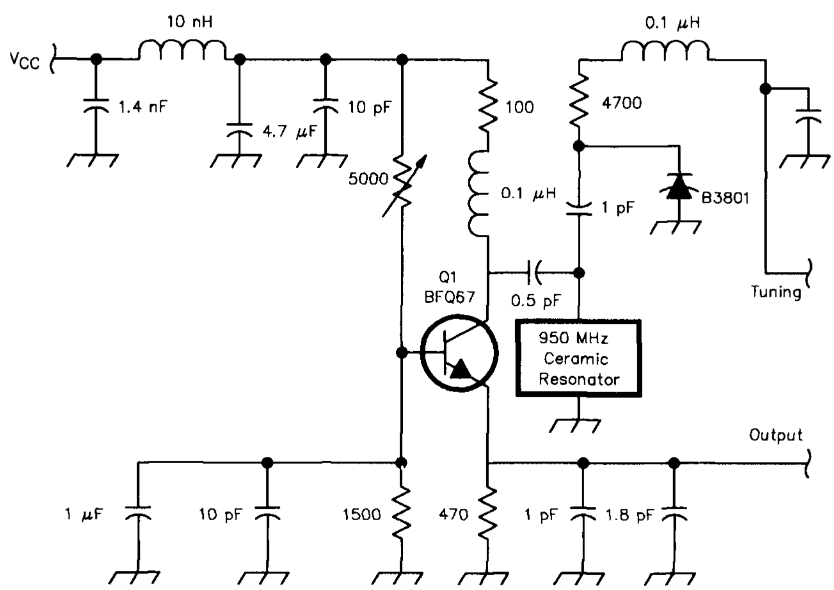


Fig 12—An example 800-MHz VCO circuit.

that the nonlinear capacitance-vs-voltage characteristic of a tuning diode results in a tuning sensitivity—and thus a noise performance—that depends on the input tuning voltage.

Fig 12 shows a typical VCO circuit. The measured phase noise of this oscillator is shown in Fig 13. While we cannot directly predict the phase noise of a circuit using only linear simulation (we will see such a prediction when we discuss the nonlinear approach), following the techniques described above can help us optimize the circuit. And we can also trade off the modulation sensitivity against the resulting phase noise, using Eqs 1, 3 and 5. This trade-off is shown in Fig 14, which graphs phase noise versus modulation sensitivity for the VCO.

Improving VCO Performance

We can improve the noise performance of a VCO by using multiple tuning diodes in the antiparallel arrangement of Fig 15. The improvement arises from two causes. First, the individual parallel diodes each have a smaller capacitance value than a single diode would need. This helps because small-capacitance diodes have less noise (lower equivalent noise resistances) than larger-capacitance diodes, due to the fabrication technology. Second, because the noise voltages from the individual diodes are uncorrelated, they do not directly sum together. Thus the effective noise is less than that of a single diode. While this scheme increases the cost of the circuit, it can reduce the diode noise contribution by as much as 15 or 20 dB compared to using a single diode. The result can approach the performance of a resonator using a single high-Q capacitor.

The Nonlinear Approach

Nonlinear circuit simulation allows us to fully predict oscillator circuit performance, including noise. In this case, we rely on a large-signal model for the transistor, such as the Curtice, Statz, TOM or Materka FET models, or the Gummel-Poon bipolar transistor model. To use this approach successfully requires careful parameterization of the device.

A nonlinear simulator such as Compact Software's *Microwave Harmonica* can generate the exact output power, bias-dependent Q and noise performance of the oscillator, using an harmonic-balance technique. In Compact's *Microwave Harmonica* and *Scope* products, a frequency-dependent color-noise transistor model is used in addition to the calculation of flicker noise. Higher-frequency transistor

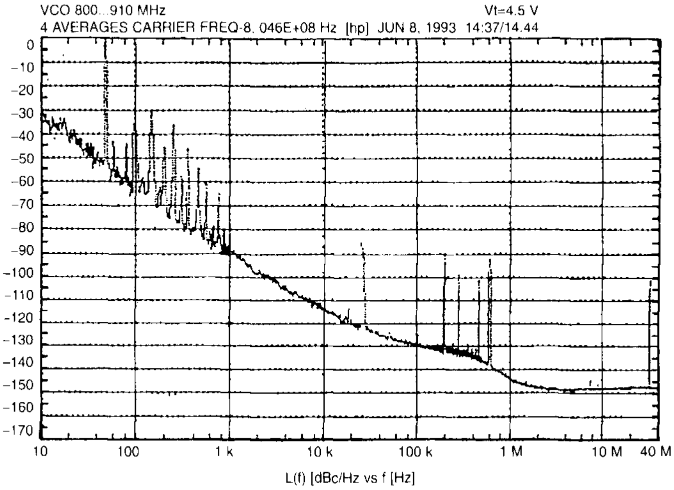


Fig 13—Measured phase noise of the oscillator of Fig 12.

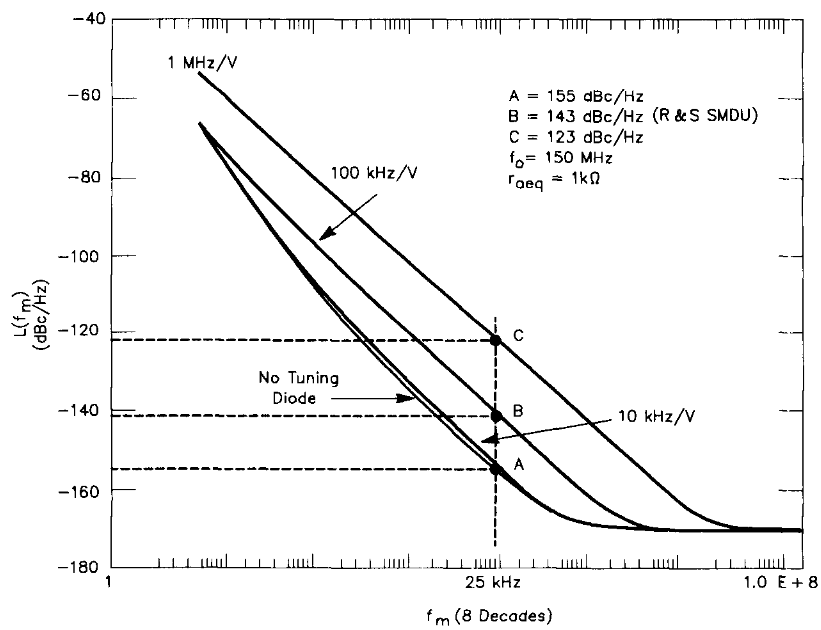


Fig 14—Phase noise increases as the tuning sensitivity of the VCO is increased, as shown here. r_{aeq} is the equivalent noise resistance of the tuning diode.

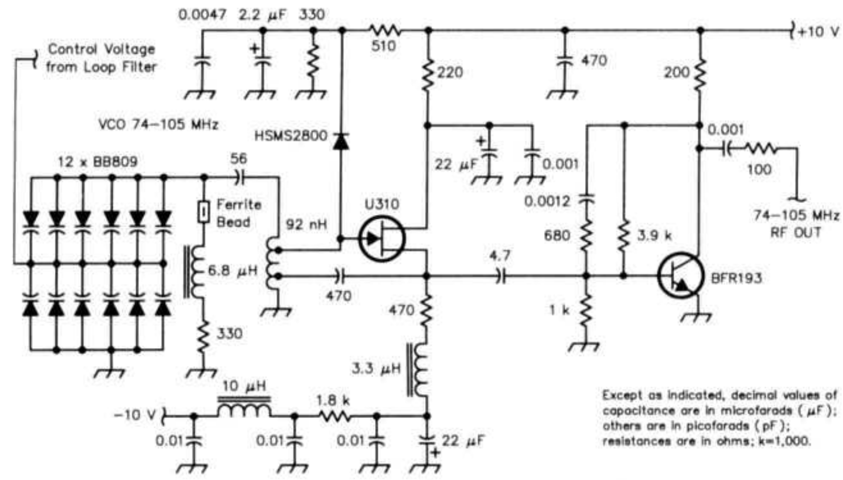


Fig 15—Improved phase-noise performance can be obtained by using a number of tuning diodes in the antiparallel arrangement shown here. The 92 nH coil is a 4-turn coil tapped at 1 and 2 turns from the ground end.

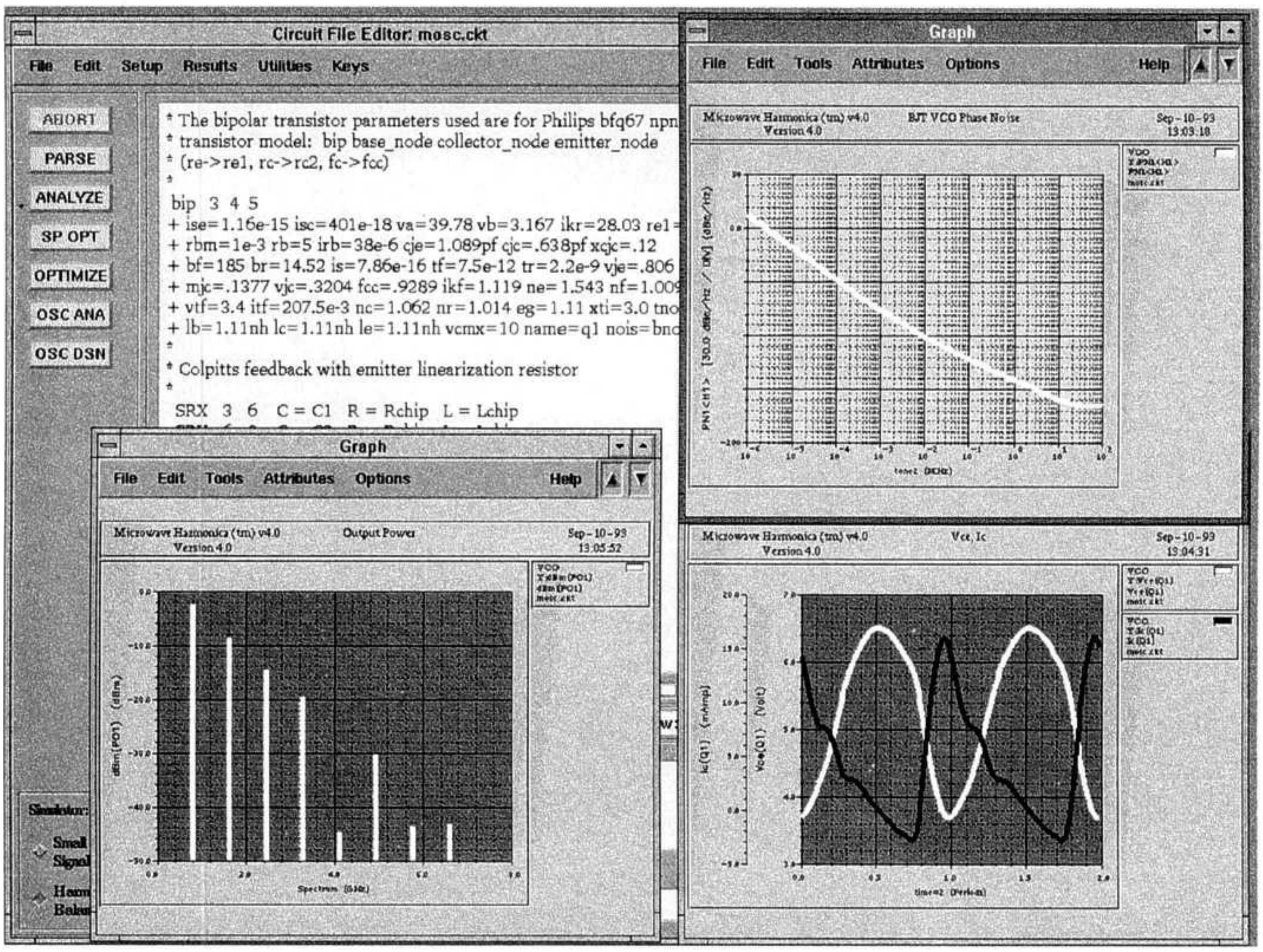


Fig 16—A *Microwave Harmonica* nonlinear analysis shows the output power spectrum, waveform distortion and phase noise of the oscillator of Fig 12.

noise is also taken into account.[7,8,9,10,11]

Fig 16 shows a *Microwave Harmonica* analysis of a Motorola 800-MHz VCO circuit similar to that of Fig 7. Included in the simulation is the noise contribution of the tuning diode.

We also can use nonlinear analysis to investigate the effect of adding a gate-clamping diode to a JFET oscillator, which is often seen in published circuits such as that of Fig 17. Figs 18 and 19 show the results of such an analysis, which shows that the clamping diode seriously degrades the noise performance of the oscillator.[12,13,14]

Coaxial Ceramic Resonators

None of the techniques we've described here can make up for a poor unloaded Q of the resonator in an oscillator circuit. But to get both small size and a high Q can be difficult. At UHF and above, one solution is the use of ceramic resonators. Siemens and other companies offer silver-plated ceramic quarter-wave TEM-mode resonators for operation at 400 to 4500 MHz. Mechanically rugged and small, these are ideal resonators for oscillators and filters in portable equipment.

These devices are shorted coaxial quarter-wave lines with silver plating on the conducting surfaces. The dielectric is one of three different relative permittivities (ε_r) of 21, 38 or 88. The needed length of a resonator can be found from:

$$L = \frac{\lambda}{4\sqrt{\varepsilon_r}} \qquad \text{Eq 6}$$

Table 2 shows typical ceramic resonator types. Larger-diameter (and higher Q) resonators can be made at customer request. Designing with these resonators is fairly simple despite the complicated procedures published elsewhere.[15]

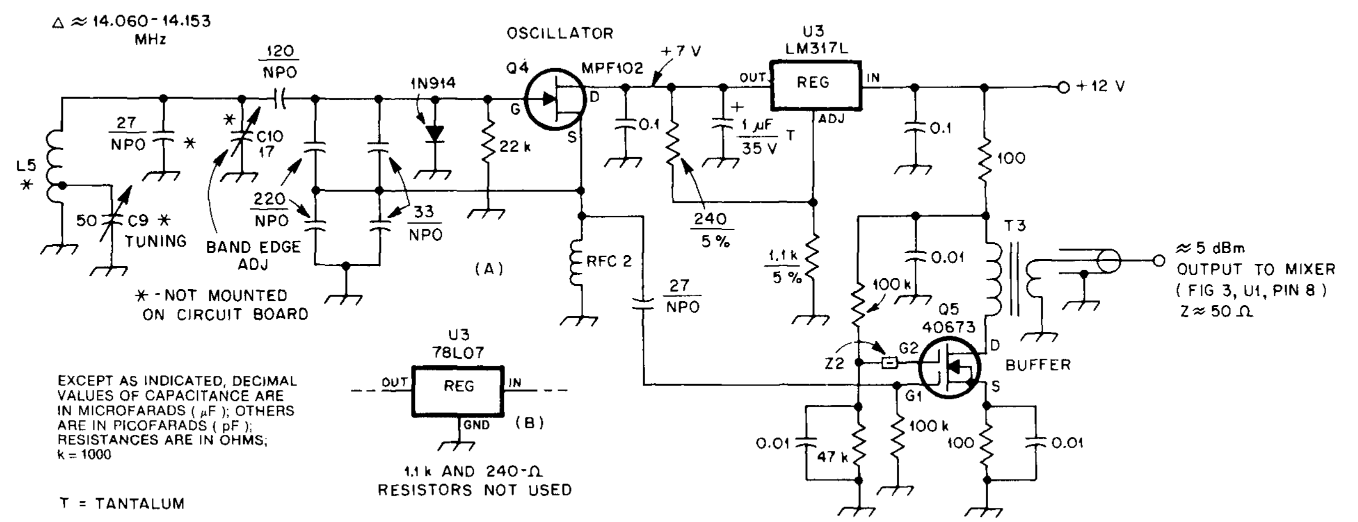


Fig 17—An oscillator with a gate clamping diode added. Predicted phase noise of the circuit is shown in Figs 18 and 19. Confirming phase-noise measurements were made at 10 MHz, after adjusting L5 for operation at that frequency.

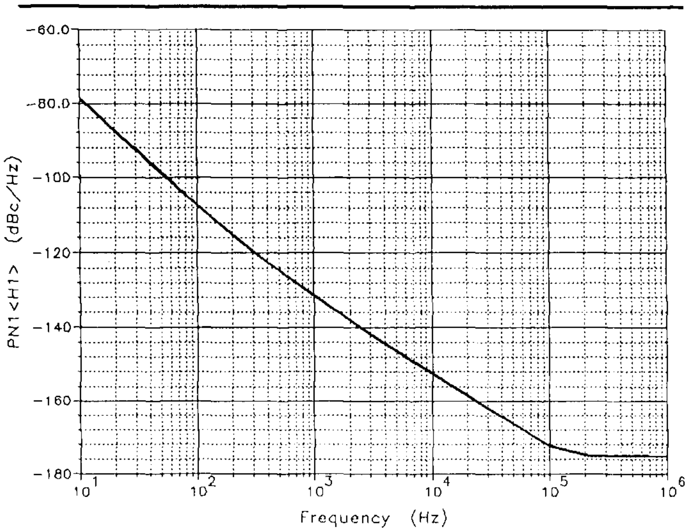


Fig 18—The phase noise of the oscillator of Fig 17.

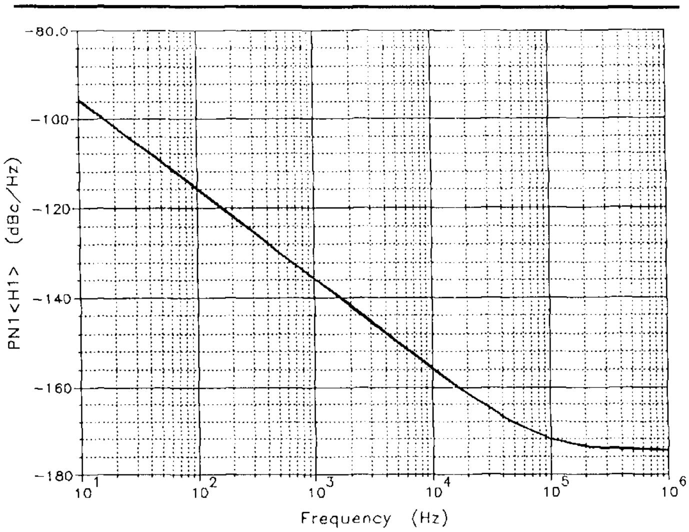


Fig 19—Without the gate clamping diode, the circuit of Fig 17 shows remarkably improved phase noise. Contrast this graph with that of Fig 18.

Equivalent Circuit Computation

Let's assume we want to design an oscillator for 900 MHz. The first thing we need to do is assume that because of loading effects we need to set the ceramic resonator frequency about 10% higher than the operating frequency. The characteristic impedance, Z_0, of a round coaxial resonator is about:

$$Z_0 = \frac{Z}{\sqrt{\varepsilon_r}} \ln(D/d) \qquad \text{Eq 7}$$

the value of Z depends on the shape of the resonator. For round resonators, Z is about 76.5 Ω. We'll use a resonator with ε_r=38, D=6.0 mm and d=2.5 mm. This gives a Z_0 of:

$$Z_0 = \frac{76.5}{\sqrt{38}} \ln(6.0/2.5) = 10.8 \ \Omega$$

or about 11 Ω. Using the fact that:

$$Z_0 = \sqrt{\frac{L}{C}} \qquad \text{Eq 8}$$

we can solve for L:

$$L = Z_0^2 C \qquad \text{Eq 9}$$

and plug this into the formula for resonant frequency:

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad \text{Eq 10}$$

Finally, we can solve for C:

$$C = \frac{1}{2\pi f Z_0} \qquad \text{Eq 11}$$

Performing these calculations for our example 990-MHz resonator gives a capacitance of about 14.7 pF and an inductance of 1.18 nH. With an unloaded Q of about 500 to 600, the parallel equivalent resistor is about 6000 Ω. The approximate length of the resonator, from Table 2, is 12.6/f mm. We now have all the needed information.

We can use the calculated L, C and R values to model the circuit. Even better, though, would be to model the resonator as a transmission line.

With the addition of an external capacitor at the open end of the resonator, the frequency can be reduced about 20%. A tuning range of about 5% is also possible. Fine tuning can be performed using a varactor diode, such as a high-Q, linear-frequency diode from Siemens. Some suggested bipolar transistors for use at 900-MHz are Siemens BFQ92P, BFS17, BFR35A and BFQ81.

To achieve good performance, the ceramic resonator should be mounted horizontally and soldered to the PC board ground on both sides along its entire length. When this is done, the resonator is wholly shielded, making it insensitive to stray fields and reactances.

Table 2—Typical Ceramic Resonator Characteristics

	ε_r=21	ε_r=38	ε_r=88
Material	(MgCa)TiO_3	(ZrSn)TiO_4	(BaPb)NdO_3
Permittivity	21 ± 2	38 ± 2	88 ± 5
D (mm)	5.5 ± 0.1	6.0 ± 0.1	5.7 ± 0.1
d (mm)	2.2 ± 0.1	2.5 ± 0.1	2.3 ± 0.1
L (mm)	16.6/f_{res}	12.6/f_{res}	8.2/f_{res}
Typical Q	800	500	300
Freq range (MHz)	2500-4500	800-2500	400-1600

Notes

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Practical Microwave Antennas

Part 2—Parabolic Dish Antennas

by Paul Wade, N1BWT

Parabolic dish antennas can provide extremely high gains at microwave frequencies. A 2-foot dish at 10 GHz can provide more than 30 dB of gain. The gain is only limited by the size of the parabolic reflector; a number of hams have dishes larger than 20 feet, and occasionally a much larger commercial dish is made available for amateur operation, like the 150-foot one at the Algonquin Radio Observatory in Ontario, used by VE3ONT for the 1993 EME Contest. But these high gains are only achievable if the antennas are properly implemented, and dishes have more critical dimensions than horns and lenses.

Background

Last September (1993), I finished my 10-GHz transverter at 2 PM on the Saturday of the VHF QSO Party. After a quick checkout, I drove up Mt. Wachusett and worked four grids using a small horn antenna. However, for the 10-GHz Contest the following weekend, I wanted to have a better antenna ready.

Several moderate-sized parabolic dish reflectors were available in my garage but lacked feeds and support structures. I had thought this would be no problem, since lots of people, both amateur and commercial, use dish antennas. After reading several articles in the ham literature, I had a fuzzy understanding and was able to put a feed horn on one of the dishes and make a number of contacts of over 200 km from Mt. Washington, in horizontal rain.

But I was not satisfied that I really understood the details of making dishes work, so I got some antenna books from the library and papers from IEEE journals and did some reading. This article is an attempt to explain for others what I've learned. The 10-GHz antenna results from the 1993 Central States VHF Conference suggest that I might not be the only one who is fuzzy on the subject—the dishes measured had efficiencies of from 23% to less than 10%, while all the books say that efficiency should typically be 55%. On the other hand, there are enough hams doing successful EME work to suggest that some have mastered feeding their dishes. One of them, VE4MA, has written two good articles on TVRO dishes and feedhorns for EME.[1,2]

There have been some good articles written by antenna experts who are also hams, like KI4VE, K5SXK and particularly W2IMU in *The ARRL UHF/Microwave Experimenter's Manual*, which is an excellent starting point. However, as I struggled to understand things that are probably simple and obvious to these folks, I did some reading and then used my personal computer to do some of the more difficult calculations and plot them in ways that helped me to understand what is happening. Many of us find a picture easier to comprehend than a complex equation. What I hope to do here is to start at a very basic level and explain the fundamentals, with pictures and graphics, well enough for hams to implement a dish antenna that works well. An accompanying computer program, *HDL_ANT*, is provided to do the necessary design calculations and to draw templates for small dishes in order to check the accuracy of the parabolic surface.

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[1]Notes appear on page 22.

HDL_ANT can be downloaded from the ARRL BBS (203-666-0578) or via the Internet from ftp.cs.buffalo.edu in the /pub/ham-radio directory.

Dish Antenna Design

A dish antenna works the same way as a reflecting optical telescope. Electromagnetic waves, either light or radio, arrive on parallel paths from a distant source and are reflected by a mirror to a common point, called the focus. When a ray of light reflects from a mirror or flat surface, the angle of the path it takes leaving the surface (angle of reflection) is the same as the angle at which it arrived (angle of incidence). This optical principle is familiar to anyone who misspent a part of his youth at a pool table! If the mirror is a flat surface, two rays of light that arrive on parallel paths leave on parallel paths; however, if the mirror is curved, two parallel incident rays leave at different angles. If the curve is parabolic ($y = ax^2$), then all the reflected rays meet at one point, as shown in Fig 1. A dish is a parabola of rotation, a parabolic curve rotated around an axis that passes through the focus and the center of the curve.

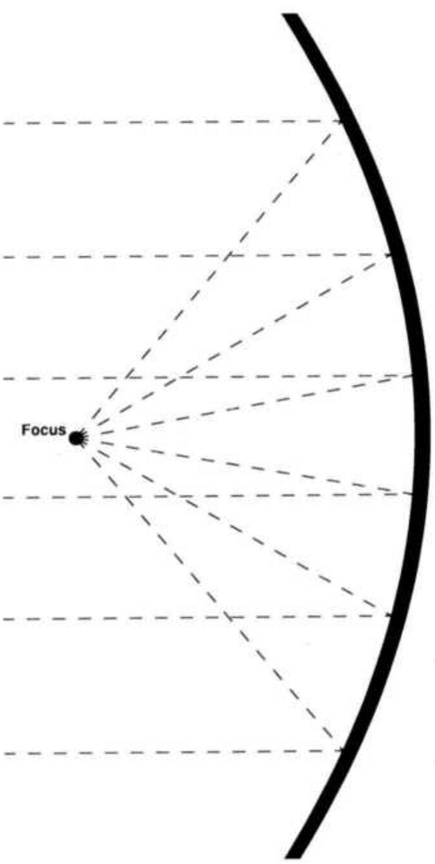

Fig 1—The geometry of a parabolic dish antenna.

A transmitting antenna reverses the path: the light or radio wave originates from a point source at the focus and is reflected into a beam of rays parallel to the axis of the parabola.

Illumination

Some of the difficulties found in real antennas are easier to understand when considering a transmitting antenna but are also present in receiving antennas, since antennas are reciprocal. One difficulty is finding a point source, since any antenna, even a half-wave dipole at 10 GHz, is much bigger than a point. Even if we were able to find a point source, it would radiate equally in all directions, so the energy that was not radiated toward the reflector would be wasted. The energy radiated from the focus toward the reflector illuminates the reflector, just as a light bulb would. So we are looking for a point source that illuminates only the reflector.

Aperture, Gain, and Efficiency

The aperture, gain, and efficiency of an antenna were all defined for antennas in general in part 1 of this series of articles. The aperture of a dish antenna is the area of the reflector as seen by a passing radio wave: $Aperture = \pi r^2$, where r is the radius, half the diameter of the dish.

If we replace a dish antenna with a much larger one, the greater aperture of the larger dish captures much more of the passing radio wave, so a larger dish has more gain than the a smaller one. If we do a little geometry, we find that the gain is proportional to the aperture.

The gain of a dish is calculated as described in part 1:

$$G_{\mathrm{dBi}} = 10\log\left(\eta \cdot \frac{4\pi}{\lambda^2} \cdot Aperture\right)$$

with reference to an isotropic radiator, h is the efficiency of the antenna. It might be amusing to calculate the gain of the VE3ONT 150-foot dish at various frequencies; use 50% efficiency to make the first calculation simpler, then try different values to see how efficiency affects gain.

How much efficiency should we expect? All the books say that 55% is reasonable, and 70 to 80% is possible with very good feeds. Several ham articles have calculated gain based on 65% efficiency, but I haven't found measured data to support any of these numbers. On the other hand, KI4VE suggests that the amateur is lucky to achieve 45-50% efficiency with a small dish and a typical "coffee-can" feed.[3]

Practical Dish Antennas

When we first described a parabolic dish antenna, we put a point source at the focus, so that energy would radiate uniformly in all directions both in magnitude and phase. The problem is that the energy that is not radiated toward the reflector will be wasted. What we really want is a feed antenna that radiates only toward the reflector and has a phase pattern that appears to radiate from a single point.

Feed Patterns

We have already seen that efficiency is a measure of how well we use the aperture. If we illuminate the whole reflector, we will be using the whole aperture. Perhaps our feed pattern should be as shown in Fig 2, with uniform feed illumination across the reflector. But when we look more closely at the parabolic surface, we find that the focus is farther from the edge of the reflector than from the center. Since radiated power diminishes with the square of the distance (inverse-square law), less energy is arriving at the edge of the reflector than at the center; this is commonly called space attenuation or space taper. In order to compensate,

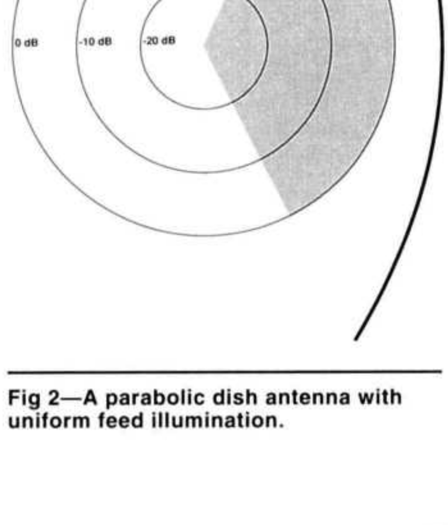

Fig 2—A parabolic dish antenna with uniform feed illumination.

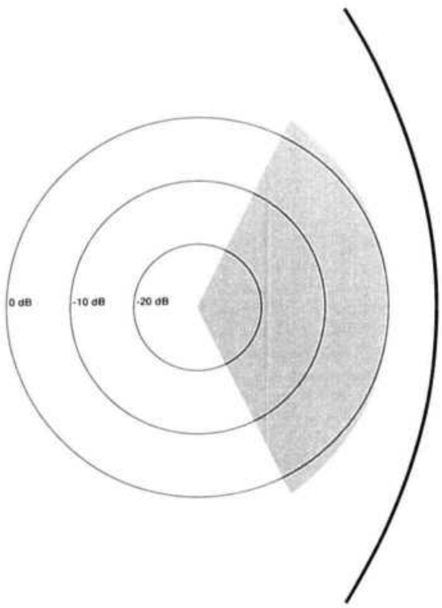

Fig 3—The desired dish illumination would provide uniform field intensity at all points on the reflector.

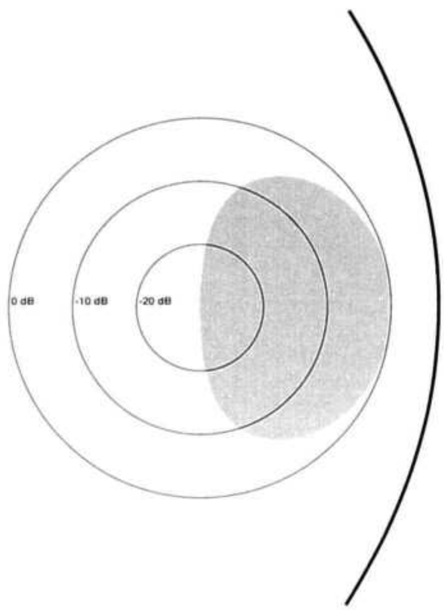

Fig 4—Typical illumination of a dish using a simple horn feed.

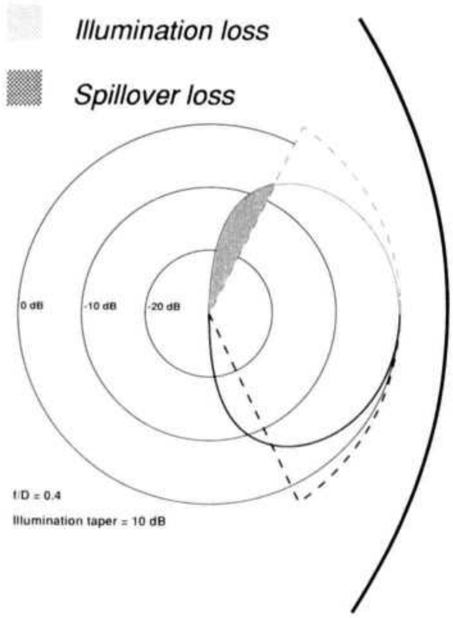

Fig 5—A comparison of typical dish illumination with the desired illuminaton.

we must provide more power at the edge of the dish than in the center by adjusting the feed pattern to that shown in Fig 3, in order to have constant illumination over the surface of the reflector.

Simple feed antennas, like a circular horn (coffee-can feed) that many amateurs have used, have a $\cos^q(\theta)$ pattern like the idealized pattern shown in Fig 4. In Fig 5 we superimpose that on our desired pattern; we have too much energy in the center, not enough at the edges, and some misses the reflector entirely. The missing energy at the edges is called illumination loss, and the energy that misses the reflector is called spillover loss. The more energy we have at the edge, the more spillover we have, but if we reduce spillover, the outer part of the dish is not well illuminated and is not contributing to the gain. Therefore, simple horns are not ideal dish feeds (although they are useful). In order to have very efficient dish illumination we need to increase energy near the edge of the dish and have the energy drop off very quickly beyond the edge.

Edge Taper

Almost all feedhorns will provide less energy at the edge of the dish than at the center, like Fig 4. The difference in power at the edge is referred to as the *edge taper*, or *illumination taper*. With different feedhorns, we can vary the edge taper with which a dish is illuminated. Different edge tapers produce different amounts of illumination loss and spillover loss, as shown in Fig 6: a small edge taper results in larger spillover loss, while a large edge taper reduces the spillover loss at the expense of increased illumination loss.

If we plot these losses versus the energy at the edge of the dish in Fig 7, we find that the total efficiency of a dish antenna peaks with an illumination taper, like Fig 6, so that the energy at the edge is about 10 dB lower than the energy at the center.[4,6] This is often referred to as 10-dB edge taper or edge illumination—often recommended but not explained.

G/T

When an antenna is receiving a signal from space, such as a satellite or EME signal, there is very little background noise emanating from the sky compared to the noise generated by the warm (300 K) earth during terrestrial communications. Most of the noise received by an antenna pointed at the sky is earth noise arriving through feed spillover. As we saw in Fig 6, the spillover can be reduced by increasing the edge taper, while Fig 7 shows the efficiency, and thus the gain, decreasing slowly as edge taper is increased. The best compromise is reached when *G/T*, the ratio of gain to antenna noise temperature, is maximum. This typically occurs with an edge taper of about 13 dB, but the optimum edge taper for *G/T* is a function of receiver noise temperature and sky noise temperature at any given frequency.[2]

Focal Length and f/D *Ratio*

All parabolic dishes have a parabolic curvature, but some are shallow dishes, while others are much deeper and more like a bowl. They are just different parts of a parabola that extends to infinity. A convenient way to describe how much of the parabola is used is the *f/D* ratio, the ratio of the focal length *f* to the diameter *D* of the dish. All dishes with the same *f/D* ratio require the same feed geometry, in proportion to the diameter of the dish. The figures so far have depicted one arbitrary *f/D*; Fig 8 shows the relative geometries for commonly used *f/D* ratios, from 0.25 to 0.65, with the desired and idealized feed patterns for each.

Notice the feedhorn patterns for the various *f/D* ratios in Fig 8. As *f/D* becomes smaller, the feed pattern to illuminate it becomes broader, so different feedhorns are needed to properly illuminate dishes with different *f/D* ratios. The feedhorn pattern must be matched to the reflector f/D. Larger f/D dishes need a feedhorn with a moderate beamwidth, while a dish with an f/D of 0.25 has the focus level with the edge of the dish, so the subtended angle that must be illuminated is 180 degrees. Also, the edge of the dish is twice as far from the focus as the center of the dish, so the desired pattern would have to be 6 dB stronger (inverse-square law) at the edge as in the center. This is an extremely difficult feed pattern to generate. Consequently, it is almost impossible to efficiently illuminate a dish this deep.

Phase Center

A well-designed feed for a dish or lens has a single phase center, as described in part 1 of this series of articles, so that the feed radiation appears to emanate from a single point source, at least for the main beam, the part of the pattern that illuminates the dish or lens. Away from the main beam, the phase center may move around and appear as multiple points, as stray reflections and surface currents affect the radiation pattern. Also, the phase center will move with frequency, adding difficulty to broadband feed design. Fortunately, we are only considering narrow frequency ranges here.

Symmetry of E-plane and H-Plane

On paper, we can only depict radiation in one plane. For a simple antenna with linear polarization, like a dipole, this is all we really care about. A dish, however, is three-dimensional, so we must feed it uniformly in all planes. The usual plane for linear polarization is the E-plane, while the plane perpendicular to it is the H-plane. Unfortunately, most antennas not only have different radiation patterns in the E- and H-planes, but also have different phase centers in each plane, so both phase centers cannot be at the focus.

Focal Length Error

When I started actually measuring the gain of dish antennas, I discovered the most critical dimension to be the focal length—the axial distance from the feed to the center of the dish. A change of ¼ inch, or about a quarter-wavelength, changed the gain by a dB or more, shown in Table 1 as measured on a 22-inch dish with $f/D = 0.39$.

I was surprised at this sensitivity, since my experience with optics and photography suggested that this is not so critical—it would be extremely difficult to adjust a lens or telescope to an optical quarter wavelength. But lenses become more critical to focus as the f-stop is decreased—an f 2 lens is considered to have a very small depth of field, while an f 16 lens has a large depth of field, or broad focus. The f-stop of a lens is the same as the f/D ratio of a dish—both are the ratio of the focal length to the aperture diameter. A typical reflector telescope has a parabolic reflector of f 8, but a dish antenna with $f/D = 0.4$ has an f-stop of 0.4, so focusing is much more critical.

More reading located an article which described how to calculate the

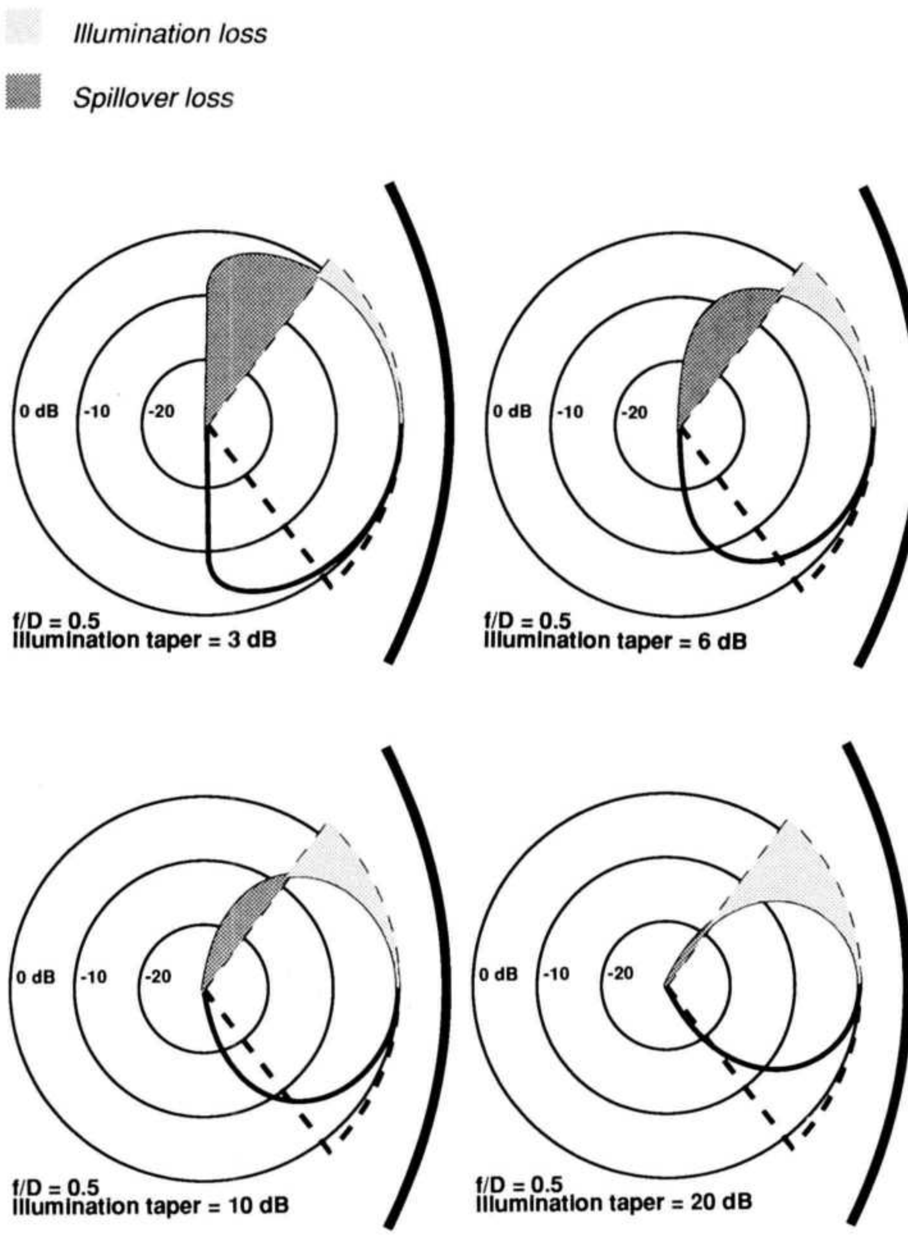

Fig 6—Dish illumination at various values of illumination taper.

Table 1—Measured Effect of Focal Length Error at 10 GHz

Feed Distance (in)	*Relative Gain (dB)*
8.125	-0.6
8.25	0
8.375	-0.3
8.625	-1.7

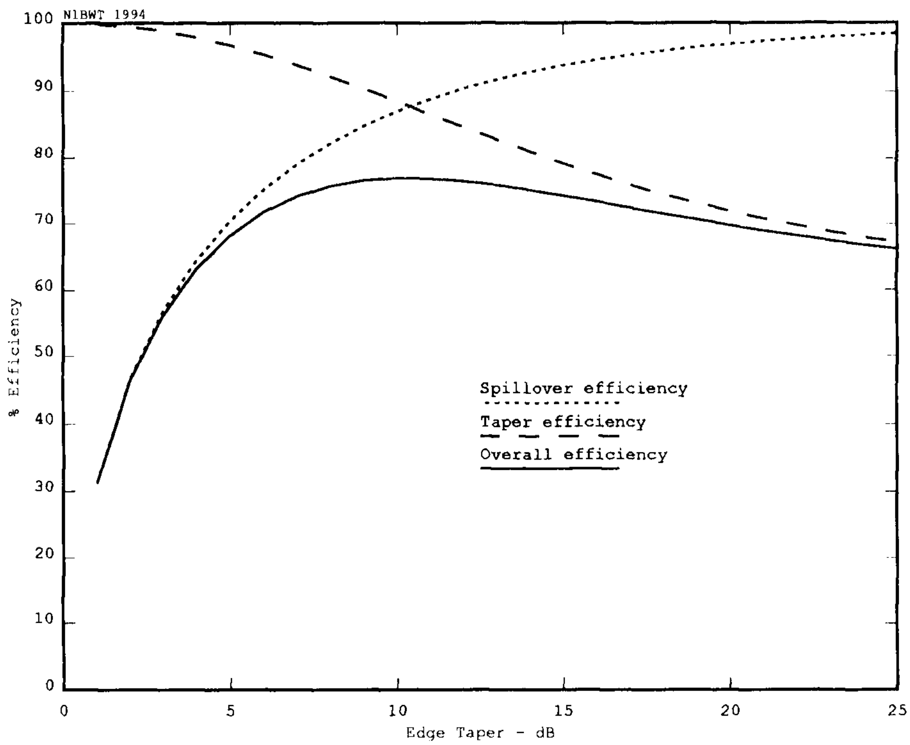

Fig 7—Dish efficiency versus edge taper. The peak efficiency occurs at a taper of about 10 dB.

loss due to focal length error.[7] Fig 9 shows the loss as the feedhorn is moved closer and farther than the focus for various *f/D* dishes with uniform illumination; the tapered illumination we use in practice will not have nulls as deep as the curves shown in Fig 9. It is clear that dishes with small *f/D* are much more sensitive to focal length error. Remember that a wavelength at 10 GHz is just over an inch.

The critical focal length suggests that it is crucial to have the phase center of the feed exactly at the focus of the reflector. Since the phase center is rarely specified for a feedhorn, we must determine it empirically, by finding the maximum gain on a reflector with known focal length.

If we are using a feedhorn with different phase centers in the E- and H-planes, we can also estimate the loss suffered in each plane by referring to Fig 9.

Lateral errors in feedhorn position are far less serious; small errors have little effect on gain, but do result in shifting the beam slightly off boresight.

Notice that the focal-length error in Fig 9 is in *wavelengths*, independent of the dish size. A quarter-wavelength error in focal length produces the same loss for a 150-foot dish as for a 2-foot dish, and a quarter-wavelength at 10-GHz is just over ¼ inch. Another implication is that multiband feeds, like the WA3RMX triband feed, should be optimized for the highest band, since they will be less critical at lower bands with longer wavelengths.[8]

Total Efficiency

It has been fairly easy to calculate efficiency for an idealized feed horn pattern due to illumination taper and spillover, but there are several other factors that can significantly reduce efficiency. Because the feed horn and its supporting structures are in the beam of the dish, part of the radiation is blocked or deflected. A real feed horn also has sidelobes, so part of its radiation is in undesired directions and thus wasted. Finally, no reflector is a perfect parabola, so the focusing of the beam is not perfect. We end up with quite a list of contributions to total efficiency:

- illumination taper
- spillover loss
- asymmetries in the E- and H-planes
- focal point error
- feedhorn sidelobes
- blockage by the feed horn
- blockage by supporting structures
- imperfections in parabolic surface
- feedline loss

KI4VE suggests that the amateur is lucky to achieve 45-50% efficiency with a small dish and a typical coffee-can feed.[3] I suspect that the only way to find total efficiency, or to optimize it, is to make gain measurements on the complete antenna.

Practical Feed Systems

An optimum feed would approximate the desired feed pattern for the *f/D* of the parabolic reflector in both planes and have the same phase center in both planes. Let's examine some of the available feed horn designs to see how well they do:

1. Dipole

Most hams know what the pattern from a dipole looks like—in free space, it looks like a donut with the dipole through the hole. If it is near ground or a reflector, the pattern in the plane perpendicular to the dipole (H-plane) is distorted to emphasize radiation away from the reflector. The shape of the radiation in this plane is controlled by the distance from the reflector, while the shape of the radiation in a plane parallel to the dipole (E-plane) does not change significantly. This suggests that the best we could do is to find a dish with an edge angle that approximates the E-plane beamwidth and adjust the reflector spacing so that the H-plane beamwidth matches the E-plane. Round disk reflectors are frequently used, but it turns out that the pattern is the same as a half-wavelength rod reflector.

2. Dual Dipole

The H-plane beamwidth can be narrowed by adding a second parallel dipole over a plane reflector, such as the EIA (sometimes erroneously called NBS) reference antenna.[9] This is a reasonably good feed with good symmetry for reflectors with *f/D* around 0.55 and has been used with good success for 432-MHz EME.

3. Penny-Splasher

The penny-splasher feed is equivalent to a dual dipole with reflector—the slots in the waveguide act as dipoles.[10]

4. Rectangular Horn

The beamwidth of a horn antenna is controlled by the horn aperture dimension, but a square horn has different E- and H-plane beamwidths. We can make it rectangular with the aperture dimensions adjusted so that the E- and H-plane patterns and beamwidths are similar. G3RPE described this technique and showed that at 10 GHz it can only illuminate dishes with f/D greater than 0.48 if the horn is driven by common WR-90 waveguide.[11] However, the smaller WR-75 waveguide is also suitable for 10 GHz and could

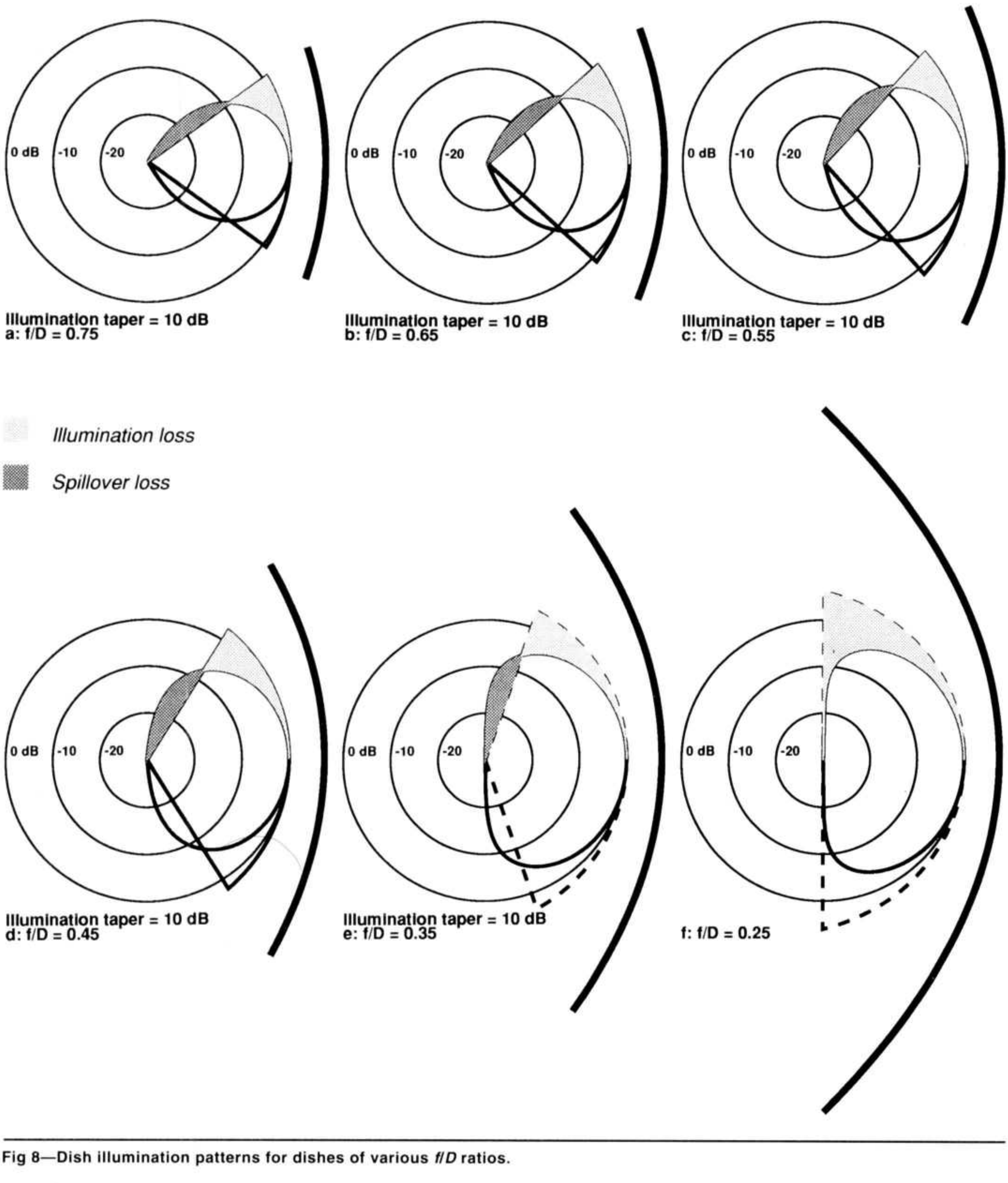

Fig 8—Dish illumination patterns for dishes of various *f/D* ratios.

drive horns which would illuminate an *f/D* as small as 0.43.

With a rectangular horn, it is difficult to achieve both a common phase center for the E- and H-planes and similar patterns in both planes. The horn section of the *HDL_ANT* computer program calculates the phase centers and allows adjustment of dimensions to change them. Kraus shows a series of patterns for horns with different flare angles, and some of them approximate the desirable feed pattern of Fig 3.[12] However, no phase information is given; W2IMU once told me they were terrible, and I accept his authority.

5. Circular Horn

A circular horn antenna, since it is symmetrical, might be expected to provide a fairly symmetrical pattern. Unfortunately, it doesn't, and the phase centers are different for the E- and H-planes. The beamwidth is controlled by the diameter of the horn—for wide beamwidths, the horn may have no flare, like the coffee-can feed, or cylindrical horn, often used at 1296 MHz.[13]

Some improvement in the pattern may be provided by adding a choke flange to a cylindrical horn.[14] Further improvement is possible by adding slots in the flange, though radiation patterns are shown in only one plane.[15]

All of the above feeds have $\cos^q(\theta)$ *patterns similar to Fig 4. Many of these were developed for radar applications, where feed inefficiency may be compensated by increased power. More recently, satellite communication has prompted research into more efficient feed antennas, particularly for deep dishes (small* f/D*) with reduced sidelobes and better* G/T. *Here are a few of the many variations that have been described, chosen for their potential for construction without elaborate machining:*

6. Clavin Feed

The Clavin feed is a cavity antenna fed by a resonant slot, with probes that excite a second waveguide mode to broaden the pattern in the H-plane to match the E-plane.[16] Radiation patterns approximate our desired feed pattern, Fig 3, while maintaining a good phase center. Fig 10 is a sketch of one I made from a 1-inch copper plumbing pipe cap. It is best for deep dishes with *f/D* in the 0.35 to 0.4 range. The resonant slot makes it more narrowband than the others (not a problem for amateur use), and the smaller size would have less feed blockage than the "Chaparral" or Kumar feeds, so it might provide better performance on smaller dishes.

A scalar feed is one that has no inherent polarization; the word "scalar" means that the electric field distribution is independent of the axis in which you look at the distribution. The result is that scalar horns have equal

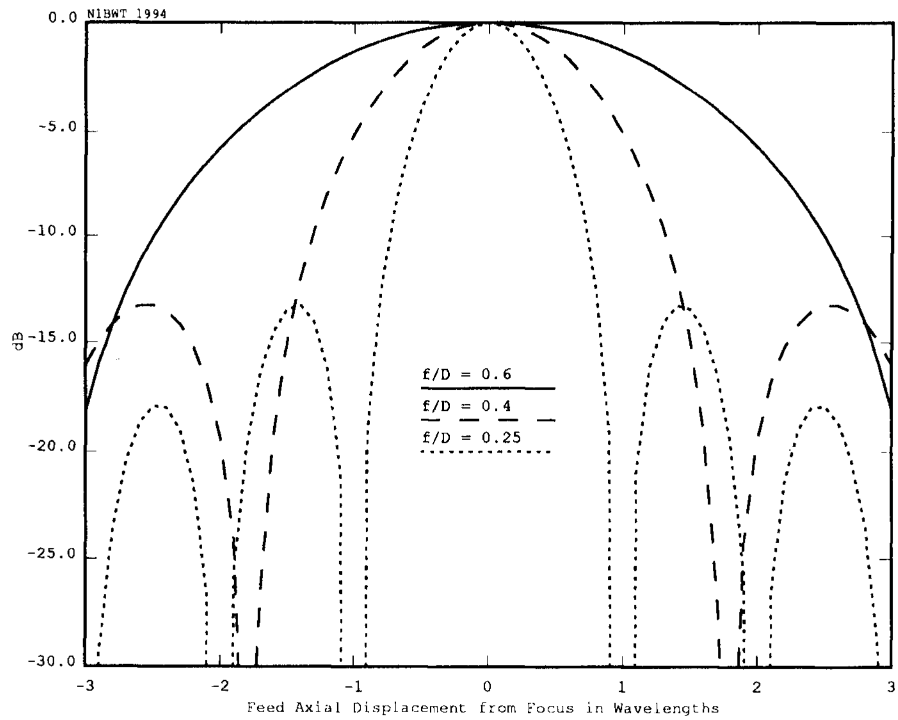

Fig 9—The loss due to axial displacement of the feed from the focus point is highly dependent on the *f/D* ratio.

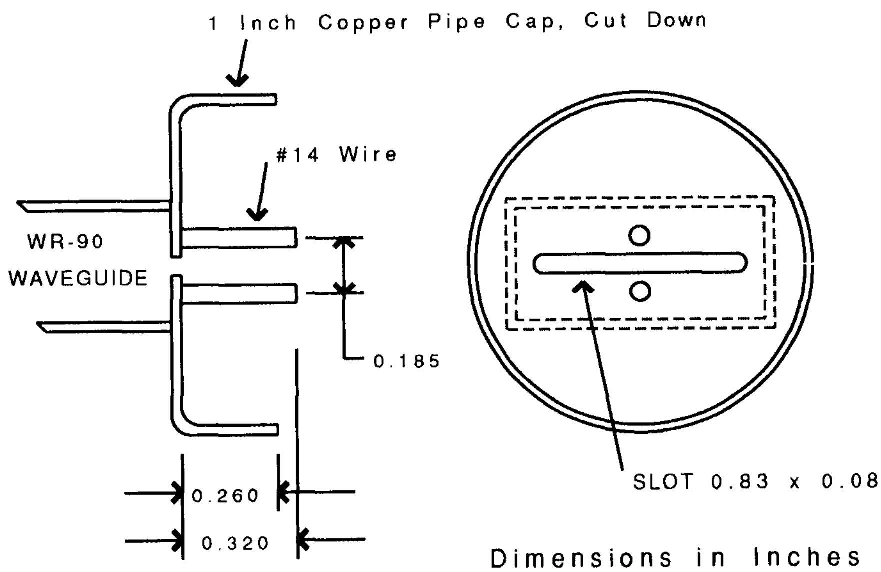

Fig 10—A Clavin feed for 10 GHz, made from a 1-inch copper pipe cap.

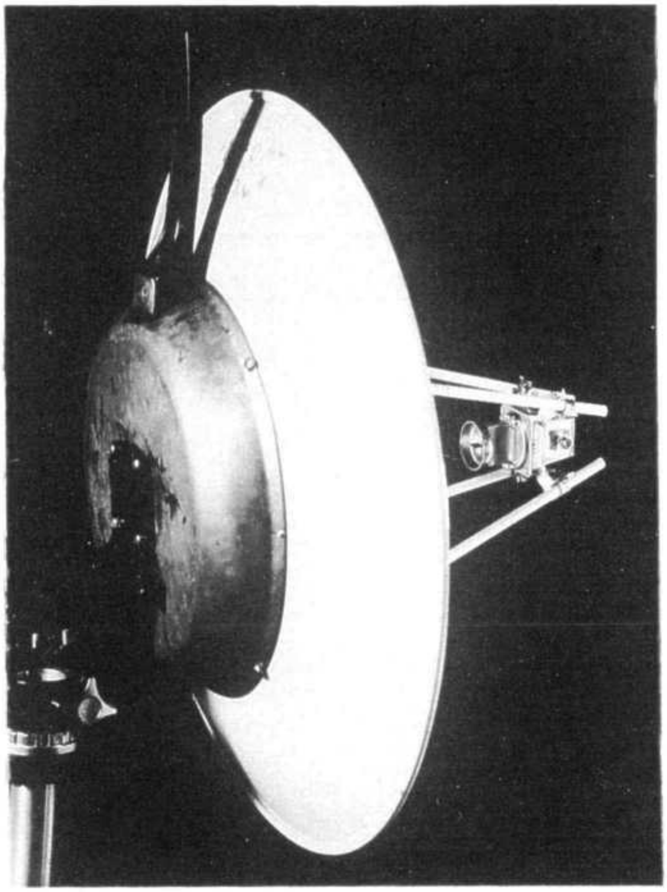

Fig 11—This photo shows the technique of mounting a dish using a frying pan with a rolled edge. Also note the Clavin feed used with this dish.

beamwidths and sidelobes in both azimuth and elevation. This can't be achieved with a standard flared horn, so scalar horns are usually preferred for dish feeds. The symmetry also makes them suitable for both linear and circular polarization. The W2IMU dual-mode horn and the "Chaparral" and Kumar feeds below are scalar feeds.

7. W2IMU Dual-Mode Horn

Diffraction from the edge of a horn causes sidelobes that reduce efficiency. In the W2IMU dual-mode horn design, there is a flare from a small section, which only supports the lowest waveguide mode, to a larger section that supports two waveguide modes.[5,17,18] The size of the flare controls the relative amplitude of the two modes, and the length of the large section is chosen so that the two modes cancel at the edge of the horn because they travel at different phase velocities in the waveguide. The cancellation eliminates the sidelobes and thus puts more energy onto the reflector. The requirement for a larger horn makes this feed optimum for larger *f/D* reflectors, in the 0.5 to 0.6 range.

8. Chaparral Feed

The "Chaparral" feed is a type of scalar feedhorn often found on TVRO dishes, with a series of cavity rings surrounding a circular waveguide.[19,5,4] The rings modify the pattern to approximate our desired feed pattern, Fig 3, while maintaining a good phase center. This feed is best for deep dishes, with *f/D* in the 0.35 to 0.45 range. Fine adjustment of the pattern is possible by changing the protrusion of the central waveguide in relation to the surrounding rings.

Note: I have not seen any mention of the location of the phase center, but my experiments show that it is controlled by the location of the outer rings, not the central waveguide.

9. Kumar Feed

The Kumar feed is a scalar feedhorn similar to the Chaparral feed, but with a single larger outer ring, so construction is somewhat simpler.[20] Radiation patterns approximate our desired feed pattern, Fig 3, while maintaining a good phase center. Ham-band versions of this feed have been described by VE4MA for 1296, 2304 and 3456 MHz.[2,21] Like the Chaparral feed, it is best for deep dishes, with *f/D* in the 0.35 to 0.45 range, with similar fine adjustment.

Complete Dish Antennas

Many of the papers describing feed horns show great detail of the horn performance, but very few even mention what happens when a reflector is added. The reflector may add too many uncertainties for good research, but our goal is to make a good working antenna. We want high efficiency because a dish has the same size, wind loading, and narrow beamwidth regardless of efficiency—we should get as much performance as possible for these operational difficulties. In other words, if I am going to struggle with a one-meter diameter dish on a windy mountain top, I certainly want one meter worth of performance!

In order to compare the different feeds, I wanted to measure the gain of several of them with the same reflector, to find their performance as complete antennas. I made a mechanism from an old slotted-line carriage and some photographic hardware that allows the feed to be moved in three dimensions with fine control of adjustment, so the feed position can be adjusted for maximum gain.

The emphasis here is on smaller dishes intended for mountaintopping and other portable operation, so maximum gain with minimum size and weight is a definite consideration. For other applications, there would be other considerations; EME, for instance, would mandate maximum performance.

Parabolic Reflector

I have managed to collect a half-dozen parabolic reflectors of various sizes and origins, and I wanted to know if they were useful at 10 GHz. First, for each dish I measured the diameter and depth in the center of the dish in order to calculate the focal length and f/D ratio. This can only be an approximation for some dishes, due to holes or flat areas in the center. The focal length is calculated as:

$$f = \frac{D^2}{16 \cdot depth}$$

The *HDL_ANT* computer program does the calculation and then generates a Postscript plot of a parabolic curve for the specified diameter and *f/D* ratio. For each reflector, I made a series of plots on a laser printer for a range of f/D values for antennas in general near the calculated value, cut out templates, then fitted them to the surface to find the closest fit. For 10 GHz, the surface must be within ±1 mm of a true parabola for optimum performance, although errors up to ±3 mm result in only 1 dB degradation.[22] I selected several reflectors with good surfaces and discarded one that wasn't even close.

Given a choice, a reflector with a large f/D (0.5 to 0.6) would be preferable. As described earlier, dishes with small f/D are hard to illuminate efficiently and are more sensitive to focal length errors. On the other hand, a dish that is available for the right price is always a good starting point!

Parabolic reflectors can come from many sources, not just antenna manufacturers. Some aluminum snow coasters (now unfortunately replaced by plastic, but aluminum foil glued to the surface might make them usable) are good, and hams in Great Britain have put dustbin lids into service as effective parabolic reflectors for years.

Homebrewing a parabolic reflector is possible, but great difficulty is implied by the surface accuracy cited above. The surface accuracy requirement scales with wavelength, so the task is easier at lower frequencies. Of course, hams are always resourceful—N1IOL found that the cover from his 100-pound propane tank was an excellent 14-inch parabolic surface and has used it to mold a number of fiberglass reflectors. K1LPS then borrowed a larger cover from a different type of propane tank and found it to be nowhere near a parabola!

Recommended Feed Systems

Since no single feed system is optimum for all dishes, a good feed recommendation depends on the f/D of the particular dish. For shallow dishes (f/D of 0.5 to 0.6), I'd recommend the W2IMU dual-mode horn or a pyramidal horn designed for the exact f/D.[5,11] The horn section of the computer program will design the horn and plot a construction template. For deeper dishes (f/D of 0.3 to 0.45), I'd recommend the Chaparral, Kumar or Clavin feeds.[5,20,16] For 10 GHz, a Chaparral horn designed for 11-GHz TVRO use works well; your local satellite TV dealer might be persuaded to order it as an "11 GHz Superfeed."

Mechanical Support

There are two critical mechanical problems: mounting the feedhorn to the dish and mounting the dish to the tripod. Most small dishes have no backing structure, so the thin aluminum surface is easily deformed. K1LPS discovered that some cast-aluminum frying pans have a rolled edge that sits nicely on the back of a dish; Mirro is one suitable brand. This is a good use for that old frying pan with the worn-out Teflon coating, so buy a new one for the kitchen. Tap a few holes in the edge of the old pan, screw the dish to it, and you have a solid backing. A solid piece of angle iron or aluminum attaches the bottom of the frying pan to the top of a tripod. The photograph in Fig 11 shows a dish mounted using a frying pan. WA1MBA uses this technique for a 24-inch dish at his home and reports that it stands up well to New England winters.

The mounting structure for the feedhorn is in the RF field, so we must minimize the blockage it causes. We do this by using insulating materials and by mounting the support struts diagonally, so they aren't in the plane of the polarization. Fiberglass is a good material; plant stakes or bicycle flags are good sources, and WA5VJB recommends cheap target arrows. Use of four rather than three struts is recommended—if they are all the same length, then the feed is centered. The base of the struts should be attached to the backing structure or edge of the frying pan; the thin dish surface is not mechanically strong.

Aiming

A quality compass and a way of accurately aligning the antenna to it are essential for successful operation. Narrow beamwidth and frequency uncertainty can make searching for weak signals frustrating and time-consuming. A heavy tripod with setting circles is a good start; hang your battery from the center of the tripod and it won't blow over as often. Calibrate your headings by locating a station with a known beam heading rather than by eyeballing the dish heading; small mechanical tolerances can easily shift the beam a few degrees from the apparent boresight. As W1AIM can testify, having the wind blow a dish over can distort it enough to move the beam to an entirely different heading.

Alternatives

The narrow beamwidth may actually make contacts more difficult, particularly in windy conditions. I have worked six grids from Mt. Wachusett in central Massachusetts using a small Gunnplexer horn. The longest path, 203 km, required a 12-inch lens for additional gain to make the contact on wideband FM; it would have been easy with narrowband SSB or CW.[23]

For a rover station, a reasonable size horn might be a good compromise, with adequate gain and moderate beamwidth for easy aiming. I often use the 17.5-dBi Gunnplexer horn, with a 12-inch lens ready to place in front of it when signals are marginal.

Conclusions

A parabolic dish antenna can provide very high gain at microwave frequencies, but only with very sharp beamwidths. To achieve optimum gain, careful attention to detail is required: checking the parabolic surface accuracy with a template, matching the feedhorn to the f/D of the dish, and, most importantly, accurately locating the phase center of the feedhorn at the focus.

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Inexpensive PC A-to-D

Use your PC's game port to interface to the analog world.

by Gary C. Sutcliffe, W9XT

The PC's standard I/O ports can be used by the resourceful amateur for uses other than they were intended. In an earlier article I described how the LPT printer port could be used as a general purpose parallel port.[1] At the request of several readers, this article continues in that direction by discussing the game port and how it can be used for analog to digital (A/D) conversion.

Admittedly, the game port is a crude A/D converter. Even so, the game port is good enough for game applications, and may also be good enough for some of your nongame applications as well. This article shows how to use the game port and describes a simple $2 method of measuring temperature.

In its normal use the game port measures the resistance of potentiometers on a joystick's X and Y axes to determine the stick position. It also has digital inputs to detect switch closures. The game port supports two joysticks as well as four switch inputs.

The conversion of resistance to a digital format is done with one-shot circuits (sometimes called monostable multivibrators, or just monostables). When a one-shot is triggered its output goes high for a period of time determined by a resistor and capacitor. The larger the values of R and C, the longer the period the output will be high. The game port has four such one-shots. The timing capacitors are fixed in value and are located on the game port board. The resistors are on the joystick, and their resistancea vary with the movement of the stick.

The game port's monostables work with resistances ranging from under 1 kΩ to around 100 kΩ. The resulting time periods will range from a few tens of microseconds to over a millisecond. The monostable's output period is measured by software with timing loops or in conjunction with the PC's hardware timers.

The Game Port Register

The game port has one 8-bit register located at I/O address 201 (hex). Table 1 shows the bit functions of this register when it is read.

The first four bits are the outputs of the one-shots. Each bit reads as 0 until the one-shot is triggered. It then reads as 1 until the one-shot times out. The one-shots are

Table 1—Game port register bits. The port is at I/O address 201h.

Bit	*Signal*
0	Monostable output, pins 1 & 3
1	Monostable output, pins 6 & 8
2	Monostable output, pins 9 & 11
3	Monostable output, pins 13 & 15
4	Switch, pin 2
5	Switch, pin 7
6	Switch, pin 10
7	Switch, pin 14

[1]Sutcliffe, Gary C., "Simple and Inexpensive PC Interfacing," *QEX*, November 1993.

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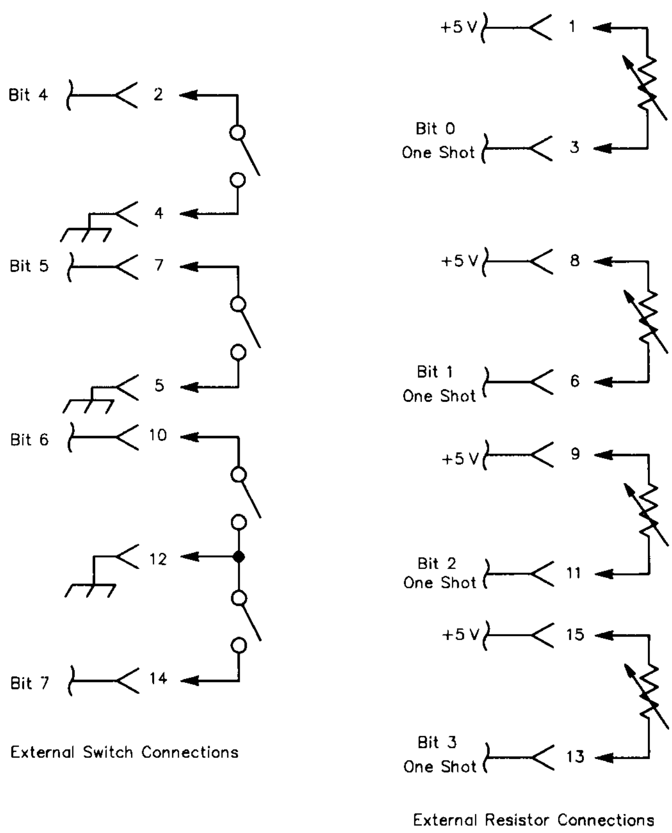

Fig 1—Game port interfacing.

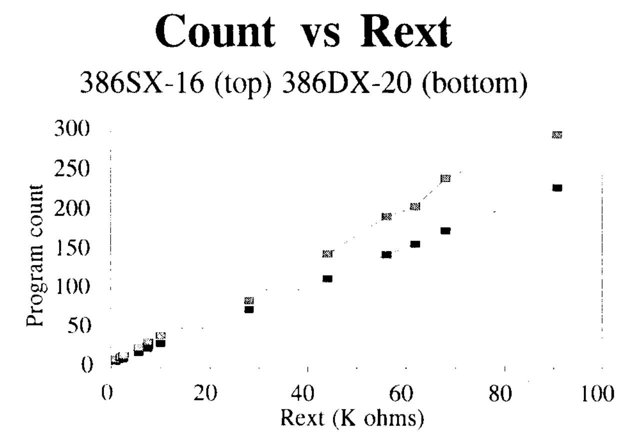

Fig 2

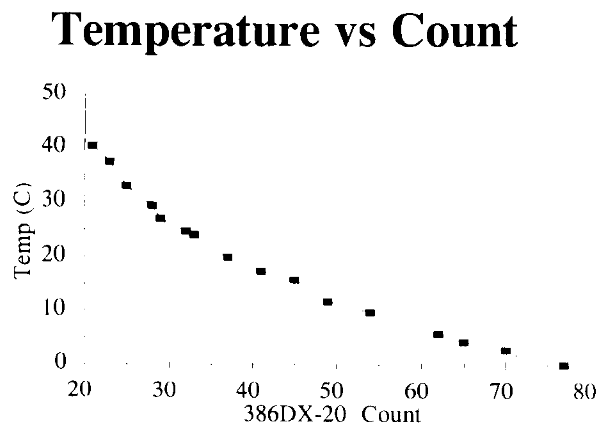

Fig 3

triggered by a write to the game port register. The data value written is irrelevant; it is the act of writing that triggers the mono-stables. Note that all four of the one-shots are triggered at once with a single write operation.

The upper four bits in the game port register are used to detect switch closures, usually in the form of firing buttons on the joystick unit. Of course you can use them to detect other switch status as needed for your particular application.

Using the Switch Inputs

Fig 1 shows how to connect external switches to the game port. You can use a common ground for the switches if needed.

The switch bits read as logical ones if the switch is open and as logical zeros if the switch is closed. Programmers using Borland's *Turbo C* can test the status of the game port register by using the inportb() function. This function reads an 8-bit byte from a PC I/O device register. Other *C* dialects will have a similar function, but the name may be different.

Microsoft's *QBasic* has a special function for testing the switch status, STRIG(). STRIG returns a 0 if the switch indicated by the argument is open and a –1 if it is closed. Listing 1 shows examples for using it. Note that the odd-numbered arguments are used to check if the switch is currently closed. The even-numbered arguments check if the switch has been closed since the last time STRIG was called with the same argument. Of course, if you don't want to use the STRIG functions, you can use the INP() function.

Take switch bounce into account when using mechanical switches with the game port. Many switches make and break contact several times quickly when they are opened or closed. A computer is fast enough to sense each closure, which could cause problems. For example, suppose you wanted to count the number of times the switch was pressed. You could easily get a much higher count than you should. Another situation that can cause a problem is if you want to initiate an activity when the switch is pressed and continue until it is released.

An easy way to handle switch bounce is to use a delay. After you sense a change in state on the switch, wait a period of time before checking the state again. Delays of 10 to 50 milliseconds are sufficient for most switches.

```
Listing 1.  C and QBasic programs to monitor game port switches

'This program shows how to use the QBasic STRIG function to monitor
' state of switches connected to the game port
'
' STRIG arguments:
' 1 = current status of switch on pin 2
' 3 = current status of switch on pin 10
' 5 = current status of switch on pin 7
' 7 = current status of switch on pin 14
'
' 0 = checks if switch on pin 2 closed since last STRIG(0)
' 2 = checks if switch on pin 10 closed since last STRIG(2)
' 4 = checks if switch on pin 7 closed since last STRIG(4)
' 6 = checks if switch on pin 14 closed since last STRIG(6)
'
' STRIG returns 0 if the switch is/was open, and 1 if it is/was closed

DO
  IF STRIG(1) = 0 THEN PRINT "Pin 2 switch is open" ELSE PRINT "Pin2 switch closed"
  IF STRIG(4) = 0 THEN PRINT "Pin 7 SW not closed since last pass" ELSE PRINT "Pin 7
                                                                          SW was closed"
  PRINT
  FOR i = 1 TO 1000      'just a delay
  NEXT i
LOOP     'forever
END
                              *******

/*****************************************************************/
/* GPSW.C    This program checks the status of the switches connected */
/* the game port.  The inportb() and delay() statements are Borland   */
/* TURBO C functions for PC support.  Other C dialects may have       */
/* different function names.                                          */
/*****************************************************************/

#include <stdio.h>
#include <dos.h>

#define GPA 0x201      /*Game port I/O address*/

#define MASK2 0x10   /*mask to blank bits other than switch on pin 2*/
#define MASK7 0x20   /*mask to blank bits other than switch on pin 7*/
#define MASK10 0x40   /*mask to blank bits other than switch on pin 10*/
#define MASK14 0x80   /*mask to blank bits other than switch on pin 14*/

main()
{
unsigned char x,y;
delay(0);        /* calibrate delay function*/
for(;;)          /*do forever*/
       {
       x = inportb(GPA);                /*read the port*/
       printf("\n\n");
       if(x & MASK2)printf("Pin 2 Switch open\n");
       if(x & MASK7)printf("Pin 7 Switch open\n");
       if(!(x & MASK14))printf("Pin 14 Switch closed\n");
       if(inportb(GPA) & MASK10)
              {
              printf("Pin 10 Switch open\n");
              }else{
              printf("PIN 10 Switch closed\n");
              }
  delay(500);    /* wait 500 msec so the screen does not fill up too fast */
                 /* A shorter delay might be used here to debounce switches */
       }
}      /* end of GPSW.C */
```

Listing 1

```
Listing 2.  QBasic and C programs to demonstrate game port A/D

' This program demonstrates the QBasic function STICK for use with the
' game port's analog inputs.
' STICK(0) returns the count for resistor on pins 1 & 3
' STICK(1) returns the count for resistor on pins 6 & 8
' STICK(2) returns the count for resistor on pins 9 & 11
' STICK(3) returns the count for resistor on pins 13 & 15

DO
   r1 = STICK(0) 'trigger all the one shots & return value for Rext pin 1 & 3
   r2 = STICK(1) 'return value for Rext pins 6 & 8 measured in last statement
   r3 = STICK(2) 'value for Rext pins 9 & 11
   r4 = STICK(3) 'value for Rext pins 13 & 15
   PRINT "R1: "; r1; "    R2: "; r2; "    R3: "; r3; "    R4: "; r4
   FOR i = 1 TO 1000       'just a delay before starting again
   NEXT i
LOOP
END

                         ********************

/*****************************************************************/
/* GP1.C  This simple program shows how to use the monostables as a   */
/* simple A/D converter.  This program can also be used for calibration */
/* purposes.                                                          */
/*                                                                    */
/* This program was written in Turbo C.  Some changes may be required */
/* with other compilers.                                              */
/* July 1994  - Gary Sutcliffe W9XT                                   */
/*****************************************************************/

#include <stdio.h>
#include <dos.h>
#define GPADR 0x201    /*Game port register address*/

/* use the proper mask for the the A/D channel you are using */
#define MMASK1 0x01   /* mask for monostable 1, pins 1 & 3*/
#define MMASK2 0x02   /* mask for monostable 2, pins 6 & 8*/
#define MMASK3 0x04   /* mask for monostable 3, pins 9 & 11*/
#define MMASK4 0x08   /* mask for monostable 4, pins 13 & 15*/


main()
{
int count;
delay(0);                       /* calibrate delay */

while(1)
      {
      count =0;                /* zero loop counter*/
      disable();            /*turn off interrupts*/
      outportb(GPADR,0);       /*trigger one shots*/
      while(inportb(GPADR) & MMASK1)count++; /*increment count until it times out*/
      enable();             /*turn interrupts back on*/
      printf(" %d\n",count);
      delay(500);                /*wait 1/2 second before doing it again */
      }
}   /* end GP1.C */
```

Listing 2

Using the A/D Converter

Fig 1 shows how to connect resistors to the game port's DB-15 connector. A sensor that varies resistance in response to some stimulus can be used in place of the resistors.

Variable resistors might be mechanically hooked up as position sensors, as is done for the joystick. For example, many rotators use a resistor to indicate direction. Some home-brew satellite and EME arrays use a pot that has its body fixed with respect to the antennas. A rigid pendulum arrangement is attached to the shaft. The pendulum keeps the shaft fixed with respect to the ground, but the pot body rotates as the array is elevated. Other variable resistance devices are candidates for use, too. Thermistors are obvious choices for temperature sensing, and some humidity sensors change resistance with changes in relative humidity.

Keep a few things in mind whenselecting your sensor. For best resolution the sensor should have a large change of resistance for a small change in position, temperature or whatever is being measured. The sensor should also remain between 1 kΩ and 100 kΩ over the range you wish to cover.

The typical method of determining the on-times of the monostables is to trigger them and then use a software loop to count up until the output goes low. Listing 2 shows sample routines in Borland's *Turbo C* and Microsoft *QBasic*.

QBasic has a built in function, STICK, which can be used to time the game port monostables. To use STICK, you must supply the number of the monostable you wish the value of. You must first call STICK(0) though. STICK(0) does the actual timing for all four one-shots and returns the value of the first one. STICK(1), STICK(2), and STICK(3) will return the respective values measured by the last STICK(0).

Note that the *C* version of the program turns off the interrupts before triggering the monostables and doing the counting. This is done because if the computer's real-time clock or other hardware interrupts while it is in the timing loop, the resulting count will be lower than it should be.

The actual count value returned will depend on a number of factors. First, of course, is the value of the resistor. Other factors include the speed of the computer. A faster computer will give a higher count (it executes the loop more times in a given period). Timing loops written in assembly language or a compiled high-level language will also give higher counts than those in an interpreted one.

The *QBasic* STICK function apparently uses the PC's hardware timers to measure the one-shots. Reducing the PC's clock speed by pressing the Turbo switch does not

```
Listing 3.  C program to display temperature

/*****************************************************************/
/* GEMP.C  - This program times the monostable on the game port.   */
/* A Radio Shack thermistor has been connected to pins 1 & 3.  It  */
/* calibrated, using the program GP1.C.  The calibration data is   */
/* put in the table, and a linear interpolation is done to convert */
/* the gameport count to temperature.  Every computer/thermistor   */
/* must be calibrated for accurate results.                        */
/*                                                                 */
/* This program was written in Borland's Turbo C.  Some changes may */
/* be needed if other C compilers are used.                        */
/* July 1994 - Gary C. Sutcliffe  W9XT                             */
/*****************************************************************/


#include <stdio.h>
#include <dos.h>

#define TSIZE  16    /* number of table entries */
main()
{
float table[TSIZE][2] = { 21, 40.3,  23, 37.4,  25, 33,  28, 29.4,  29, 27,
                    32, 24.7, 33, 24.1,  37, 20,  41, 17.4,  45, 15.8,
                    49, 11.8,  54, 9.8,  62, 5.8,  65, 4.3,  70, 2.8,
                    77, 0.0 };  /*This is the count to temperature
                    table.  First number is count, second is temperature.
                    This data would need to be changed for other systems &
                    thermistors. */

int count;            /* game port count value */
int i;                /* general purpose*/
float temp;           /* temperature */
float fcount;         /* count converted to floating point*
float dc;             /*delta count - used for interpolation*/

count = 0;

disable();            /* disable interrupts so clock wont interfere */
outportb(0x201,0);      /* trigger game port monostable */
while(inportb(0x201) & 0x01)count++;  /* time the monostable */
enable();               /* turn interrupts back on */


fcount = (float)count;
temp = 0.0;
if(fcount < table[0][0])temp = 999;  /*out of range - report it*/
if(fcount > table[TSIZE 1][0])temp = -999;  /*out of range*/
if(temp == 0.0)
      {      /* do a linear interpolation */
      i = 1;
      while(fcount >  table[i][0] && i <  TSIZE-1)i++;
      dc = (fcount - table[i-1][0])/(table[i][0] - table[i-1][0]);
      temp = ((table[i-1] - table[i]) * dc) + table[i-1][1];
      }
if(fcount - (float)((int)fcount) >= .05)fcount = fcount + 1;  /*round it off*/
printf("\n\nTemperature = %dC\n",(int)temp);

}  /* end of GTEMP.C */
```

Listing 3

change the values returned by STICK. I prefer the *C* function however. The values returned by STICK vary by plus or minus a few counts, even when a fixed resistor is used, but the *C* function shows smaller short-term variations in count.

Board-to-board differences will also affect the count value. Fig 2 shows the count value vs resistance for the game ports on two different computers with the *C* program in Listing 2. A number of precision 1% resistances were used and the count value from the program was recorded. Note that the 16-MHz 386SX computer actually gave a higher count for the same resistors than the 20-MHz 386DX. This is due to differences in the game port circuits of two different multifunction I/O cards.

The hardware differences and the varying response of the game port to different resistance value means you will need to calibrate each system and then convert the count value into engineering units. You may want to make a look-up table of readings and then do a linear interpolation between the two nearest entries. Another method is to analyze the calibration data with regression analysis to generate a simple equation that relates count to the unit you wish to measure. Many spreadsheet programs and scientific calculators have such functions built in, which lets you do the analysis without dusting off your old statistics text book.

Long wires from your computer to your sensor can pick up noise that could affect the accuracy. Keep in mind that long wires will also add resistance. So, don't calibrate your antenna elevation resistor with a short cable and then run a long cable out to the tower.

Also remember to use the same language for taking your readings for calibration that you use for the final program. Using a BASIC program for calibration and an assembly program for the final product won't give very accurate results!

Simple Temperature Measurement

You can let your computer measure temperature with a thermistor. Radio Shack sells one for about $2 (P/N 271- 110). I built a temperature sensor by making up a cable with a DB-15 on one end and a Radio Shack thermistor on the other. The thermistor is connected to pins 1 and 3 of the DB-15. I wanted it to be waterproof, so I covered the pins and solder connection with heat-shrink tubing. Then I covered the tubing with silicone glue. While the glue was still wet I slipped another piece of tubing over that.

To calibrate the system you can start with a styrofoam cup of crushed ice with enough water to form a slush. The temperature of this mixture is 0° C. Use a program such as those shown in Listing 2 to get the count reading.

Next remove most of the ice and fill the cup with cold water. Once the rest of the ice melts, the temperature will gradually rise towards room temperature. Slowly stir the water. Use an accurate thermometer and record the temperature and corresponding count from the computer. (You can get a certified thermometer at most drug stores or photography shops.) When the water has reached room temperature, fill the glass with warm water and repeat the process as it cools back towards room temperature.

Fig 3 shows temperature vs count for my temperature probe and the 20-MHz 386DX. You should get a similar looking curve, but it will vary depending on your type of computer, which program you use and the characteristics of your thermistor.

Fig 3 shows several interesting things. First, notice that between counts of 20 and 30 the temperature changes by about 15° C, or about 1.5°/count. Between counts 65 and 75 the temperature only changes by about 5°, or about 0.5°/count. For a given count, the resulting temperature can be determined more accurately at lower temperatures.

If I were concerned with getting more accuracy at warmer temperatures there are a couple of things I could do. First I could try a different thermistor with different characteristics. Perhaps one with a larger temperature coefficient. Another thing I could do is use a faster computer. All else being the same, a faster program would reduce the time for one pass through the timing loop, increasing the resolution.

Listing 3 shows a simple *C* program I wrote to display the temperature. This program includes an array holding the calibration data, which it uses to do a linear interpolation to convert from counts to temperature. Because the whole system is only accurate to about 1 to 2° C, the resulting temperature value is rounded off to the nearest whole number before displaying it.

As mentioned at the beginning of this article, the game port is a crude A/D converter. It may be adequate for your application though and may save you the $100 or so that a low-end PC A/D plug-in board would cost.

The programs described here can be downloaded from the ARRL BBS (203-666-0578) for via Internet from ftp.cs.buffalo.edu in directory /pub/ham-radio. The file is named QEXGPORT.ZIP.

Digital Communications

Harold E. Price, NK6K

Misery Loves Company

In my previous column, I'd lost the will to live. I'd asked for reader input, in hopes that I'd find something to cheer me up. I said: "If you disagree (or agree) with any of the above, please write to my email or US mail addresses. If you have an interesting application that is running on the packet network, let me know. Aside from some character-based white pages, I haven't seen much. We seem to be working hard, but I'm not sure we're having any fun."

I had several responses, all more depressing than mine. I've included the most depressing in the "Mailbox" section at the end of this column. I had no response from anyone with an interesting application.

Real World

In the July/August Issue of *Wireless*, a trade magazine for cellular and other wireless data communications, I found more cause for depression. First, in an article by Ira Brodsky, "Cellular's Multi-Pronged Data Strategy," the Amateur Radio service gets a plug. Actually, it is more plug than we deserve. Mr. Brodsky states: "First adapted to radio by amateur radio operators, packet radio differs from landline packet networks..." We weren't the first users of packet radio, of course, others get that credit. We *were* the first to take existing technology, cheap VHF radios, cheap modems, cheap single board computers, and produce viable systems for a few hundred dollars. This was at a much lower cost than ever before. More than 100,000 were sold, probably making us the largest user, in sheer numbers, of packet radio, at least for a while.

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71635,1174 (Compuserve)

It is nice to get a small plug now and then. Unfortunately, the same magazine gives an indication of how hopelessly out of touch we are now. Those of you not in the industry may not have a chance to see the ads for current wireless data systems. In "The Missing Links," by David Toll, a list of leading wireless data link products is given. The systems range from a low rate of 512 kbit/s and a range of 30 miles to 8 Mbit/s and a range of one mile. All of these systems use spread spectrum, mostly in the 902 to 928-MHz band or the 2.4-GHz band and do not require an FCC license. Many are directly Ethernet or token-ring compatible.

The only thing keeping hams from running right out and buying some for the local net is the price: $10,000 to $20,000 per pair. Remember, this is the commercial price, though. Some of these 900-MHz systems come with $700 price tags for omnidirectional antennas and $5000 for directional antennas. The electronics have a similar burden of real-world engineering and marketing overhead. We need an initiative like TAPR's TNC project in the early 1980s to repackage the chips that make up these systems and build a ham-priced box. For the next column, I'll try and find out what can be added to a 900-MHz antenna to make it cost $5000. Even with standard markups, this is an amazing price. We only need a 10-to-1 reduction to bring the price into the range of local groups, 20-to-1 for the individual. We've done it before.

Interestingly enough, I caught a hint of a possible project that fits this category as I was preparing this article. See "Fear of Success" in the "Mailbox" section.

Computer Networking Conference Proceedings

For some reason, the *Proceedings of the 9th ARRL Computer Networking Conference* have been very popular. In addition to carrying the definition of the Pacsat data protocols, other articles are also continually referenced. The 9th has been out of print for a while. TAPR has arranged to reprint and distribute the 9th, as well as the proceedings from other years. The agreement was described in a recent announcement by TAPR: "TAPR has arranged with the ARRL to be the agent for past *Proceedings of the ARRL Computer Networking Conferences* and *Digital Communications Conference Proceedings*. This will apply to proceedings that are more than two years old. For example, the ARRL will continue to distribute the 12th (1993) and 13th (1994) proceedings this year, then next year TAPR will begin to distribute the 12th (1993). In this way, TAPR and the ARRL hope that these proceedings will be fully available to digitally interested hams. *Proceedings* are available currently from the 1st through the 11th. The 9th CNC was out of print, but as part of the agreement we have reprinted the 9th and it is again available. TAPR would like the thank Mark Wilson, AA2Z, and Jon Bloom, KE3Z, for their help at the ARRL with this arrangement."

Contact TAPR at: 8987-309 E Tanque Verde Rd #337, Tucson, AZ 85749-9399, tel: 817-383-0000, fax: 817-566-2544, for more information.

Mailbox

Lyle Johnson, WA7GXD, was heavily involved in the TAPR TNC-1 and TNC-2, the Microsat CPU, various DSP projects, and is currently working on the Phase 3D spacecraft GPS system and CPUs. He writes in response to my August "Depression" column:

> "I just returned from an EDA forum (Electronic Design Automation, the software tools used to design ICs, PCB, etc). I had an opportunity to speak with some

manufacturers who spend many tens of millions of dollars per year on UNIX software R&D. I asked them when I'd see lower-end UNIX tools so I could bail out of *Windows* and use a 'real' operating system. I was quite surprised at their responses, which were largely along the following lines:

'Which UNIX? We spend, on average, 30% of our software development budget porting our applications amongst the various incompatible UNIX platforms and implementations. It isn't a simple matter of retargetting the compiler from SPARC to PA-RISC or i486. A binary compiled for Solaris 1.0 isn't compatible with Solaris 2.0. And next year, when Sun jumps to OpenStep (their version of NextStep), that won't be compatible either. And AIX isn't the same, and HP-UX isn't the same and...

'On top of that, the GUIs are all different. Last year it was News, then it was Motif, then it was X and OpenLook. Frankly, we're pumping resources into porting to Daytona (the next Microsoft NT system). We'll still support our UNIX customers, but we're planning on migrating to NT because:

- *Microsoft allows backwards compatibility. You can still run the application you bought for your 8088 DOS 2.0 PC back in 1984 and run it on your Pentium DOS 6.22 PC you bought in 1994. I can't do that with UNIX. I have to continuously redevelop, and my clients have to continuously upgrade. Some customers won't or can't upgrade to the latest version of "UNIX." Support is a nightmare!*
- *NT is written by a monolith—Microsoft—for better or worse. This means the APIs will be the same for the Power-PC as they are for the 486. I can literally just retarget the compiler and have my application run on Intel; or Alpha, or MIPS, or...*
- *With NT, we'll be able to effectively add nearly 50% more staff devoted to making our products better with the same budget.*
- *We don't care that NT hasn't taken off yet. It will, because the tools people buy from us now will be better and cheaper under NT. Microsoft stuck to Windows from 1985 until it finally took hold with the release of Windows 3.0 in 1992. Bill Gates has the ego and money to make it happen, and it'll happen.*
- *We aren't interested in NextStep, or OS/2. We're going to NT!'*

"The point is that, among people who make a living at manufacturing and supporting products for UNIX (as opposed to products for Sun, or other specific platforms), UNIX is a nightmare.

"Linux is just another flavor of UNIX. Sure, it is available for $40 on a CD-ROM, but I can't run it on my PC at home Why? My video card is made by Diamond, and there are no Linux drivers for it. And, I can't run the applications I need under Linux (MS Word 6.0a, for example, or the interface program to my multi-mode controller, or my callbook database, or...) So, I have to reboot every time I want to do something else. Not practical, even (especially?) for home use in the ham shack.

"We use Linux at work for our commercial application—DISPATCH. We love it. Yet, when we try and buy a laptop PC for our sales staff, it turns out that we have to buy expensive high-end color machines (rather than cheap $2 k WinBook machines) because Linux doesn't support the Western Digital Rocket Chip video ICs (yet). We have to spend an extra $2.5 k per laptop to run a 'free' operating system.

"If you are buying a new computer from Zeos, or Compaq or Gateway, try asking the sales person if the video cards, and CD-ROM, and (fill in the blank) will run under Linux. You get silence. If you ask what the cards are so you can research it yourself, you usually get inaccurate information (they may ship a different video card, for example, because that was the one they could buy that day, perhaps due to chip manufacturer's long lead times). We bought a lot of PCs from Gateway for the office and Linux ran just fine. The next lot, purchased just a few weeks later, didn't. Turned out they had changed the video card. The new one ran great under *Windows*, but Linux couldn't talk to it. We wound up having to replace the video cards.

"Bottom line? I contend that UNIX is completely impractical for the average ham, and maybe even for a reasonable fraction of the techies. I think the question is MSDOS/Windows on Intel platforms versus everything else. It's too bad, because UNIX offers a lot of resources and operating flexibility that hams could use. But, using UNIX in the ham shack is like trying to run MOSAIC and the Word Wide Web over a 1200-bit/s packet link—it just isn't practical."

Fear of Success

Lyle is touching on one aspect of the problem I call "haves and have-nots." Just after the start of the "packet revolution" everyone was at the same starting point. Most of us had an IBM PC clone and most of us had a TNC-2 clone. The required software was a terminal emulator; wiring was PTT, mic in, speaker out and ground. Now we've exploded again into a plethora of computers, operating systems, TNCs, modems, radios and protocols. The technology "haves" know what they are doing, have read the RFCs, write their own device drivers and design their own interface hardware. The "have nots," while functionally literate in other walks of life, don't know why min(++a,b) didn't do what they expected, can't figure out how to cohost Linux, Windows, and OS/2 on a double-spaced disk, and don't know how to add CTS to their data cable.

As the gap between solution providers and users grows, we end up with potential providers who are unwilling to throw themselves into the bottomless pit of needs. I can attest to the problems that come with publishing solutions. I still get the occasional letter, based on two articles I wrote for *QST* almost ten years ago, asking for help. Usually from some remote island, sometimes from students who have pooled their resources to buy IRCs, I'm asked to provide an interface from computer X to TNC Y to radio Z, when I've never heard of X, Y or Z. It is painful to ignore these requests, but even more painful to try to answer each one. Glenn Elmore, N6GN, says it very well:

> "My recent posting has generated some questions about the availability of information, boards or kits of the 200-400 kbit/s radios we are using with PI2 cards here in northern California. I think I may need to explain the situation.
>
> "With regard to construction articles or other detailed documentation: I haven't yet published any. The fact is that I'm afraid to. The 2-Mbit/s, 10-GHz link published first in *Ham Radio* and then in the *ARRL Handbook* has netted around 500 telephone calls, letters and

pieces of email. My phone still rings about once a week about it. I think I'm nearing DXCC with countries of requests. This even though the original article is almost five years old and was complete enough to be a construction article (as proven by those who constructed it). Needless to say, my XYL isn't too happy about it.

"The 904-MHz radios, interfaces and antennas have more parts, are more complex and require working with 10 watts at 1 GHz, which is arguably more difficult than dealing with a 'canned' low-power 10-GHz transceiver. All this is simply not easy to 'kit.' I'm afraid that if I publish the plans (which are quite complete and include circuit boards and assembly drawings) I would generate requests for assistance that make the 10-GHz link experiences pale in comparison. I'm sure I'd have to change my telephone number.

"Still, I really do want to help as much as I can. The information on col.hp.com under ~/hamradio/packet/n6gn was put there to help alleviate this problem. It includes 10-GHz info, 904 antennas, ARRL CNC papers and more. I try to add to it as I find things that might be of interest.

"I hope this doesn't sound too stuffy. I'd really like for folks to be able to copy this hardware and have fun improving amateur radio networks, but I simply don't have the resources to support what I'm sure would ensue if I published a 'how to' article. Sometimes I wish that hardware were as easy to clone and codevelop as software.

"My approach is now, instead, to work on a 'layer 3 TNC' which is to be an entirely printed board, a very inexpensive combined direct sequence, direct conversion, spread spectrum 1200-MHz RF assembly married to a 68302-based digital controller. The whole works is intended to be in the cost range of current multimode TNCs but have an antenna connector on one side and host connector (nominally a parallel port but optionally Ethernet) on the other. It is to provide layer 3 connectivity to the associated host. Clearly, it needs to be an open environment and documented well enough that many can work on it together.

"This requires considerable software to allow economy of design. I'm attempting to perform many previously hardware functions with s/w. Additionally there needs to be a mechanism to provide user RF path measurement and qualification. Fortunately the same hardware that can recover the ~16 Mcps DS SS signal can also serve to measure path loss and multipath component. 'Just a simple matter of programming.'

"This is something I'd hope would be generally available, once there is a mechanism set up to support folks so they can reasonably expect to be successful. However, this is an enormous project and may not be finished in my lifetime.

"I wish I knew how to partition a project like this so that it could effectively be pursued in a physically distributed manner. But my experience to date is that close physical proximity of the contributors is extremely important. It also requires a breadth of skills: RF, CS/networking and organizational. All contributors need to cooperate well since they each have vital contributions to bring to the table and there are often solutions in one 'domain' which are much more expensive or obscure in one of the others. There's no question that I need help, I just don't know how to appropriate it."

I'll be talking to Glenn about getting him some development help. If anyone else has a depression buster, please email.

Upcoming Technical Conferences

The 1994 AMSAT-NA Annual Meeting and Space Symposium

October 7-9, 1994, Holiday Inn, Orlando International Airport, Orlando, Florida.

Contact: Steve Park, WB9OEP, 12122 99th Ave N, Seminole, FL 34642, tel: 813-391-7515. Internet: SKPA@QMGATE.ECI-ESYST.COM; Am Pkt Radio: WB9OEP @ W4DPH.

Events: Friday, October 7, 1994 11 AM-5 PM, open registration, International Ballroom entrance; 1-5 PM, papers and presentations—Session A—Mission Status Briefings, International Ballroom.

Saturday, October 8, 1994, 7:30-8 AM, open registration, International Ballroom entrance; 8-11:45 AM—Session B—Orbital Science and Systems, International Ballroom; 8-8:10 AM, Welcome and Opening Remarks; 8:10-11:45, AM papers and presentations; 11:45-1 PM, Lunch. Session C—P3D Design Review, International Ballroom; 1-5 PM, papers and presentations. 7-10 PM Banquet and Annual Meeting, International Ballroom (cost: $23 per person). Guest speaker, Dr. Paul Shuch, N6TX, "The Search for Dark Matter."

Sunday, October 9, 1994, 9 AM-1:30 PM, Session D—Satellite Operating, International Ballroom; 9 AM-2 PM, Bus Tours to the P3D Integration Facility;10 AM, Open Board of Directors Meeting, Continental Room.

Registration: Cost is $25.00 per person (includes Proceedings). Registration can be made by phone to AMSAT, 850 Sligo Ave, #600, Silver Spring, MD 20910, tel: 301-589-6062, fax: 301-608-3410.

Hotel, etc: Special hotel rates are available: $58.00 per night (one or two persons per room). Call the Holiday Inn at 407-851-6400 (9 AM to 6 PM EDT). 5% discount on US Air—Use Gold File Number 14210074.

ARRL Continuing Education Workshop: Computer-Aided Design of HF Antennas

October 21, 1994, Pacificon—ARRL Pacific Division Convention, Hilton Hotel, Concord, California.

Topics include: Propagation, Antenna selection, transmission line selection and putting it together.

Speakers: R. Dean Straw, N6BV, ARRL Senior Assistant Technical Editor and Roy Lewallen, W7EL, ARRL Technical Advisor.

Cost: For ARRL members is $20, $25 for nonmembers, and includes a workbook, convention ticket and six hours of practical information that will allow you to design your own antennas.

Checks should be made payable to Pacificon, and sent to Ray Gaschk, KD6BLS, 2680 Cherry Lane, Walnut Creek, CA 94596, tel: 510-933-3243. Registration cut-off is October 17, 1994. To preregister and for more information, contact the ARRL Educational Activities Department at tel: 203-666-1541 or fax: 203-665-7531.

(*Have an upcoming technical event? Drop us a note with the all the details and we'll include it in* Upcoming Technical Conferences.)

Proceedings

The following *Proceedings* are available from ARRL:

Central States VHF Society Proceedings, ISBN: 0-87259-482-3; ARRL Order Number: 4823, cost: $12 plus s/h.

Titles include:

Subharmonic IF Receivers, Rick Campbell, KK7B

1296 MHz Amplifier Notes, Dave Meier, N4MW

Shielding No-Tune Printed Filters, Rick Campbell, KK7B

PROPLOG—Automated Propagation Monitor, Robert J. Carpenter, W3OTC

Fifteen Years of EME, Francis Shepard, W7HAH

Algoquin EME Expedition, Peter Shilton, VE3VD

Modulating a "Brick" Microwave Source or Beacon Service, Dave Meier, N4MW

Simple Antennas—How to make 'em Cheap!, Kent Britian, WA5VJB

How I Integrated a Siemans TWT into My Station, Dave Meier, N4MW

Our Early Heritage—A History of VHF, Bill Tynan, W3XO

A Simple 5 Band CW Transceiver, Rick Campbell, KK7B

Equidistant Azimuthal Projection Program for Amateur Use, Joseph Mack, NA3T

Transient Suppression and Protection, Charlie Chennault, WA5YOU

Notes on Down East Microwave 5760 MHz Transverter Assembly, Dave Meier, N4MW

Practical Electric Motors 101, Charlie Chennault, WA5YOU

The Console Station, Dave Meier, N4MW

North American VHF and Above DX Records, A.J. Ward, WB5LUA

27th Central States VHF Society Conference, Preamp Measurement Results

2 Meter EME Standings, John Carter, KØIFL

23 cm EME Standings, Tommy Henderson, WD5AGO

13 cm EME Standings, Tommy Henderson, WD5AGO

1994 North American Beacon Register—144 MHz & Higher, Emil Pocock, W3EP

Central States VHF Society Proceedings—Cumulative Index, Dave Meier, N4MW

Proceedings of the Twentieth Eastern VHF/UHF Conference of the Eastern VHF/UHF Society; ISBN:0-87259-485-8; ARRL Order Number: 4858; cost: $12 plus s/h.

Titles include:

The Eastern VHF/UHF Conference Experience, Stan Hilinski, KA1ZE, Chairman

The Eastern VHF/UHF Conference Trivia Quiz, Stan Hilinski, KA1ZE, Chairman

OPERATING RELATED TOPICS

New Rover Rules, ARRL Contest News Release, Billy Lunt, KR1R

Maritime Mobile Operation Aboard the S.S. Chemical Pioneer, Clint Walker, N1KTM

50 MHz. Beacon List, O. B. Madsen, OZ1IZB

The 1994 Summer Sporadic-E Season, A Summary of Six Meter Observations, Ken Neubeck, WB2AMU

Microwave Operating—making succesful contacts, Zack Lau, KH6CP/1

Getting on 10 GHz, Dale P. Clement, AF1T

10 GHz POTPOURRI—Meeting of the North East Weak Signal Group, Photos by N2LIV and WB2DVK

AZ_PROJ.PS—A Program to Create Azimuthal Equidistant Map Projections, Joseph Mack, NA3T, and Michael Katzmann, NV3Z

ANTENNAS

Secrets of Parabolic Dish Antennas, Paul Wade, N1BWT

Accurate Dimensions for Horn Antenna Design, Jeff Millar, WA1HCO

CUSHCRAFT 4218XL Modifications Revisited, Ron Klimas, WZ1V

EQUIPMENT DESIGN

Digital Receiver Design, Peter Traneus Anderson, KC1HR

RF Power Amplifiers—Demystified, David M. Upton, WB1CMG

A Collection of VHF Filters, Zack Lau, KH6CP/1

A Digital Sequencer for Tower Mounted Preamps, Ron Klimas, WZ1V

A Look at Tower Mounting SHF Transverters: Why it Makes Good Dollars and Sense, Fred Stefanik, N1DPM

Up, Up & Away to 10-GHz or 10-GHz Semi-Commercial Style, Bruce Wood, N2LIV

Interfacing HF Tranceivers to Transverters, a Compendium, Introductory Notes and General Information, David Olean, K1WHS

ICOM IC-735 Transverter Wiring, Ron Klimas, WZ1V

Kenwood TS-830S, TS-820S Transverter Wiring, David Olean, K1WHS

Kenwood TS-850S, Drake TR-7 Transverter Wiring, Steve Powlishen, K1FO

Drake C Line, Yaesu FT757 Transverter Wiring, Fred Stefanik, N1DPM

Yaesu FT101ZD Transverter Wiring, Franks Potts, NC1I

Uniden HR-2510 Modifications for Transverter Applications, Chris Fagas, WB2VVV

Uniden HR-2510 Modification Notes, John Balboni, AC1T

Radio Shack HTX-100 Transverter Wiring, Al Carpin, NA1W

Amp Supply TR-110 Transverter Wiring, Ernie Gray, WA1NIE

TESTING

Antenna Gain Measurements Nineteenth Eastern VHF/UHF Conference

Noise Figure Measurements Nineteenth Eastern VHF/UHF Conference

The Hottest GaAsFETs, Emil Pocock, W3EP

13th ARRL Digital Communications Conference Proceedings; ISBN: 0-87259-483-1; ARRL Order Number: 4831; cost $12 plus s/h.

Titles include:

A Proposal for a Standard Digital Radio Interface, Jeffrey Austen, K9JA

Automatic Packet Reporting System (APRS), Bob Bruninga, WB4APR

Boadcast, UI and unconnected protocols—the future of Amateur Packet Radio?, Paul Evans,W4/G4BKI

Packet, GPS, APRS and the Future, Paul Evans, W4/G4BKI

Computer Networks in Africa: From Utopian Discourse to Working Reality, Iain Cook

A Low Cost DSP Modem for HF Digital Experimentation, Johan Forrer, KC7WW

G-TOR: The Protocol, Mike Huslig, Phil Anderson, Karl Medcalf and Glenn Prescott

GMON—a G-TOR Monitoring Program for PC Compatibles, Richard Huslig and Phil Anderson, WØXI

A Theoretical Evaluation of the G-TOR Hybrid ARQ Protocol, Glenn E. Prescott, WBØSKX, and Phil Anderson, WØXI

On Fractal Compression of Images for Narrowband Channels and Storage, W. Kinsner, VE4WK

Fast CELP Algorithm and Implementation for Speech Compression, A. Langi, VE4ARM, W. Grieder, VE4WSG, and W. Kinsner, VE4WK

Wavelet Compression for Image Transmission Through Bandlimited Channels, A. Langi, VE4ARM, and W. Kinsner, VE4WK

ROSE X.25 Packet Switch Status Update, Thomas A. Moulton, W2VY

A Primer on Reliability as Applied to Amateur Radio Packet Networks, T.C. McDermott, N5EG

FSK Modem with Scalable Baud Rate, Wolf-Henning Rech, N1EOW, and Gunter Jost, KD7WJ

MacAPRS: *Mac* Automatic *P*acket *R*eporting *S*ystem—A Macintosh Version of APRS, Keith Sproul, WU2Z, and Mark Sproul, KB2ICI

Formation of the TAPR Bulletin Board System Special Interest Group, David A. Wolf, WO5H

How Amateur Radio Operators Can Emulate an HF ALE Radio, David R. Wortendyke, NØWGC

A Preview of HF Packet Radio Modem Protocol Performance, Teresa Young, Stephen Rieman and David Wortendyke, NØWGC

Microwave Update 1994; ISBN: 0-87259-486-6; ARRL Order Number: 4866; cost: $12 plus s/h.

Titles include:

Chicago-Area Amateur Microwave Activity, Gary Hess, K3SIW/9

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A Broadband Dish Feed for Amateur Radio SHF, Tom Williams, WA1MBA

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Estimating Sun Noise at Various Frequencies, Based on the 10.5 cm Flux Reported by WWV, Melvin B. Graves, WRØI

Power Amplifiers for 903 Through 3456 MHz, Jim Davey, WA8NLC,

Bibliography of 33-cm (902 to 928 MHz) Articles (Neyens, NØCIH); 3456 MHz IMFET Amplifier (Ward, WB5LUA, and Fogle, WA5TNY); 2304 MHz Power Amplifier Using 7289 or Similar Tube (Malowanchuk, VE4MA); 2 GHz to 6 GHz Power Amplifiers (Hilliard, WØPW); 2.3 GHz Power Amplifiers (Campbell, KK7B); 2304 and 3456 MHz Power Amplifiers (McIntire, AA5C); A 7289 Amplifier for 3456 MHz (Malowanchuk, VE4MA); Update on a 7289 Amplifier for 3456 MHz (Malowanchuk, VE4MA); Bias Circuit for 7289 Tubes (Malowanchuk, VE4MA); Using Impedance Matched Power GaAsFETs at Other Frequencies (Malowanchuk, VE4MA)

Control Unit for RWN 220/221 TWT Power Supply, Rick Beatty, NU7Z

Viariable Bias Supply for Grounded Grid Power Triodes, Al Ward, WB5LUA

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10 GHz *No*Tune Single Board

Transverter, Toshihiko Takamizawa, JE1AAH

HEMT has broken through a barrier of 2.0 dB or NF for 24 GHz, Toahihiko Takamizawa, JE1AAH

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A Sweep Generator Using a YIG Oscillator, John Petrich, W7HQJ

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3456 MHz No-Tune Transverter Mod, Dave Mascaro, WA3JUF

Shielding Printed No-Tune Filters, Rick Campbell, KK7B

Subharmonic IF Receivers, Rick Campbell, KK7B

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A Simple High Performance Microwave Transverter, Philip Gabriel, AAØBR

Up, Up & Away to 10-GHz or 10-GHz Semi-Commercial Style, Bruce Wood, N2LIV

Programming Surplus Synthesizers for Use on 10 GHz, Bruce Wood, N2LIV, and Bob Schoenfeld, WA2AQQ

Fiber Optics: Waveguide of the Future, Dr. H. Paul Shuch, N6TX

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Feedback

A Practical Note On Designing FIR Filters

It is recommended that both positive and negative values of k be used when designing FIR filters using the method described in the sidebar, "Description of FIR Filters," in our article, "DSP Voice Frequency Compandor for use in RF Communications," from the July, 1994, *QEX* (p 8). Doing this provides much better convergence in the truncated Fourier series (Eq B2) for most ordinary desired filter responses, $H(f)$. Adding these coefficients is equivalent to doing $n/2$ time-advance operations with their associated multiply-accumulates before the delays shown in Fig A. This might bother you, because it requires knowing $x(t)$ for the future ($t > 0$). Don't let it; just define $t = 0$ to be a few seconds ago, $n/(2f)$ seconds to be exact. With these modifications, all the equations are the same, with the exception that the summations now run from $k = -n/2$ to $n/2$ instead of from 0 to n. We actually used negative k values in the filters designed for the compandor; we simply forgot to note it when writing the article.—*Rob Frohne, KL7NA*

I just noticed a small mistake in the artwork for Fig 4 of my article, "5760-MHz from the Junkbox," May 1994 *QEX.* Choke L2 should be connected to the junction of C5 and FB in the drain Circuit of the MRF-966. See the figure below.—*Robert Cook, N2SB*