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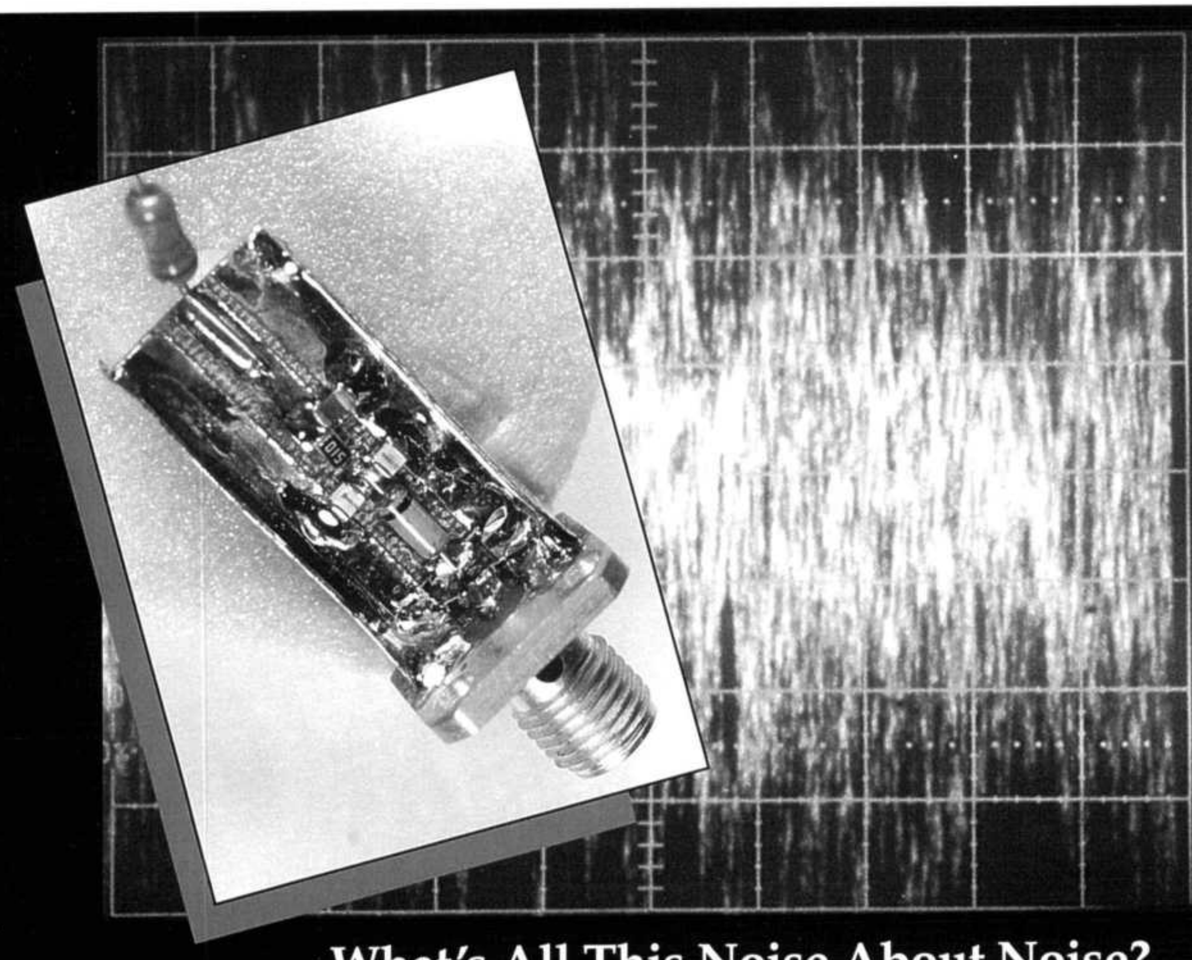
ARRL Experimenter's Exchange

November 1996



## What's All This Noise About Noise?

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QEX

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## About the Cover

**N1BWT explains how noise affects our receiving and describes circuits and techniques for measuring receiver noise performance.**

# Features



# Columns



# November 1996 QEX Advertising Index

# *Empirically Speaking*

## RF Exposure and the Experimenter

As you may know, the FCC has recently promulgated rules that will require most transmitting stations—including amateurs—to evaluate their transmission systems to ensure that they meet RF exposure limits. This is of particular interest to VHF-and-up experimenters, where station configurations and equipment can lead to large field intensities near the antenna. (Which is not to say that it's not of interest to HF users—particularly those who run significant power.) The following synopsis of the matter is provided by Laboratory Supervisor Ed Hare, KA1CV:

Regulations on RF exposure have been on the books since the mid 1980s. These regulations were based on the IEEE/ANSI C95.1 RF-exposure standards. The regulations set limits on the maximum permissible exposure (MPE) permitted from operation of transmitters in all radio services, including the Amateur Radio Service, and required some radio services to conduct station evaluations to ensure that they were in compliance. Amateurs were categorically exempt from the requirement to evaluate their stations because the relatively low power and typical operating duty cycles used by hams made it unlikely that the MPE levels would be exceeded.

On August 1, 1996, the FCC announced major changes to their RF-exposure regulations. Under the new regulations, most of the categorical exemptions for the Amateur Radio Service have been eliminated. As they always have been, hams are required to meet the limits for maximum permissible exposure (MPE), but now many operators will also be required to conduct a "routine station evaluation" of their station's operation. Stations using less than 50 W PEP and mobile or portable stations using push-to-talk do not need to be evaluated. The regulations also required that five additional questions be added to the Novice, Technician and General Class examinations.

The overall impact of these regulations is not devastating. Amateurs can easily develop the required questions for the pools. Most hams are already in compliance with the MPE requirements and some will need to conduct a simple evaluation, using look-up tables to determine if their antennas are far enough away from areas of exposure for the mode and power they are using.

The ARRL has filed a number of Petitions for Partial Reconsideration. To date, we have been successful at having the original implementation date extended from January 1, 1996 to January 1, 1997. The FCC is also permitting the question pools to be updated in their normal cycle. ARRL offered considerable assistance to the Question Pool Committee (QPC) to release revised question pools for the Novice and Technician Class question pools on schedule December 1, 1996. The FCC has not yet acted on our request to extend the federal preemption clauses in the regulations to Amateur Radio. They are also still deciding whether to add an exclusion for 150-W HF stations whose antennas are located more than 10 meters from areas of exposure.

January 1997 *QST* carries an article, "The FCC's New RF-Exposure Regulations," describing the current state of this continuing story, and up-to-date details will be available on the ARRL Web site at **http://www.arrl.org/news/rfsafety/**. Check there frequently.—*KA1CV, email:* ***ehare@arrl.org***.

## This Month in *QEX*

If you want to optimize your station for weak-signal performance, you'll need to understand and measure system noise performance. This month, Paul Wade, N1BWT, explains how "Noise Measurement and Generation" is done by the amateur.

In the May 1996 *QEX*, Mark Mandelkern, K5AM (then KN5S), presented "A Luxury Linear Amplifier" for 2 meters. This month, Mark is back with detailed design notes, construction hints and some updated circuitry.

Constructing projects is probably the most time-consuming part of the experimenter's activity. Getting some good hints on how to do it well helps, and Zack Lau, W1VT, provides details on how to mount RF connectors—one of the trickier parts of construction—in this month's "RF" column.—*KE3Z, email:* ***jbloom@arrl.org***.

# Noise Measurement and Generation

*Quality weak-signal reception requires a low-noise system. Here's how to calculate and measure the noise performance of your system.*

By Paul Wade, N1BWT

As anyone who has listened to a receiver suspects, everything in the universe generates noise. In communications, the goal is to maximize the desired signal in relation to the undesired noise we hear. To accomplish this goal, it would be helpful to understand where noise originates, how much our own receiver adds to the noise we hear, and how to minimize it.

It's difficult to improve something unless we can measure it. Measurement of noise in receivers does not seem to be clearly understood by many amateurs, so I will attempt to explain the concepts and clarify the techniques, and to describe the standard "measure of merit" for receiver noise performance: *noise figure*. Most important, I will describe how to build your own noise generator for noise-figure measurements.

A number of equations are included, but only a few are needed to perform noise-figure measurements. The rest are included as an aid to understanding, with, I hope, enough explanatory text for everyone.

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### Noise

The most pervasive source of noise is thermal noise, which arises from the motion of thermally agitated free electrons in a conductor. Since everything in the universe is at some temperature above absolute zero, every conductor must generate noise.

Every resistor (and all conductors have resistance) generates an rms noise voltage:

$$e = \sqrt{4kTRB} \qquad \text{Eq 1}$$

where $R$ is the resistance, $T$ is the absolute temperature in kelvins (K), $B$ is the bandwidth in hertz, and $k$ is Boltzmann's constant, $1.38 \times 10^{-23}$ joules/K.

Converting to power, $P=e^2/R$, and adjusting for the Gaussian distribution of noise voltage, the noise power generated by the resistor, in watts, is:

$$P_n = kTB \qquad \text{Eq 2}$$

which is independent of the resistance. Thus, all resistors at the same temperature generate the same noise power. The noise is white noise, meaning that the power density does not vary with frequency, but always has a power density of $kT$ watts/Hz. The noise power is directly proportional to absolute temperature $T$, since $k$ is a constant. At the nominal ambient temperature of 290 K, we can calculate this power; converted to dBm, we get the familiar –174 dBm/Hz. Just multiply by the bandwidth in hertz to get the available noise power at ambient temperature. The choice of 290 K for ambient might seem a bit cool, since the equivalent 17° C or 62° F would be a rather cool room temperature, but 290 K makes all the calculations come out to even numbers.

The *instantaneous* noise voltage has a Gaussian probability distribution around the rms value. The Gaussian distribution has no limit on the peak amplitude, so at any instant the noise voltage may have any value from $-\infty$ to $+\infty$. For design purposes, we can use a value that will not be exceeded more than 0.01% of the time. This voltage is four times the rms value, or 12 dB higher, so our system must be able to handle peak powers 12 dB higher than the average noise power if we are to measure noise without errors.[1]

*Signal-to-Noise Ratio*

Now that we know the noise power in a given bandwidth, we can easily calculate how much signal is required to achieve a desired signal-to-noise ratio (S/N). For SSB, perhaps 10 dB of S/N is required for good communication; the ambient thermal noise in a 2.5-kHz bandwidth is –140 dBm, calculated as follows:

$$P_n = kTB = 1.38\times10^{-23}\times290\times2500 = 1.0\times10^{-17}\,\mathrm{W}$$

$$P_{dbm} = 10\log(P_n\times1000) = -140\ \mathrm{dBm}$$

(The factor of 1000 converts watts to milliwatts.) The signal power must be 10 dB greater than the noise power, so a minimum signal level of –130 dBm is required for a 10 dB S/N. This represents the noise and signal power levels at the antenna. We are then faced with the task of amplifying the signal without degrading the signal-to-noise ratio.

[1]Notes appear on page 12.

*Noise Temperature*

Any amplifier will add additional noise. The input noise $N_i$ per unit bandwidth, $kT_g$, is amplified by gain $G$ to produce an output noise of $kT_gG$. The additional noise added by the amplifier, $kT_n$ is added to the input noise to produce a total noise output power $N_o$:

$$N_o = kT_gG + kT_n \qquad \text{Eq 3}$$

To simplify future calculations, we pretend that the amplifier is noise-free but has an additional noise-generating resistor of temperature $T_e$ at the input, so that all sources of noise are inputs to the amplifier. Then the output noise is:

$$N_o = kG\left(T_g + T_e\right) \qquad \text{Eq 4}$$

and $T_e$ is the *noise temperature* of the excess noise contributed by the amplifier. The noise added by an amplifier is then $kGT_e$, which is the fictitious noise generator at the input amplified by the amplifier gain.

*Cascaded Amplifiers*

If several amplifiers are cascaded, the output noise $N_o$ of each becomes the input noise $T_g$ to the next amplifier. We can create a large equation for the total. After removing the original input noise term, we are left with the added noise:

$$N_{added} = \left(kT_{e1}G_1G_2\ldots G_N\right) + \left(kT_{e2}G_2\ldots G_N\right) + \ldots + \left(kT_{eN}G_N\right) \qquad \text{Eq 5}$$

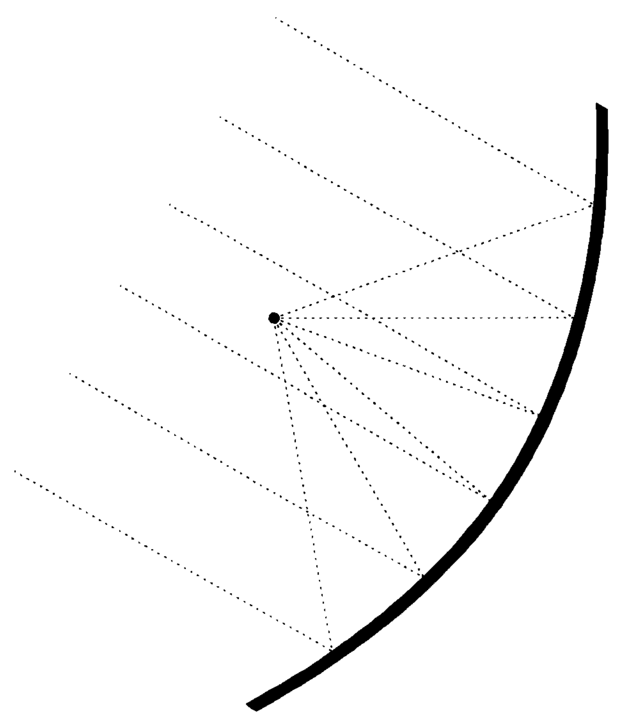

**Fig 1—A parabolic dish antenna aimed at a satellite.**

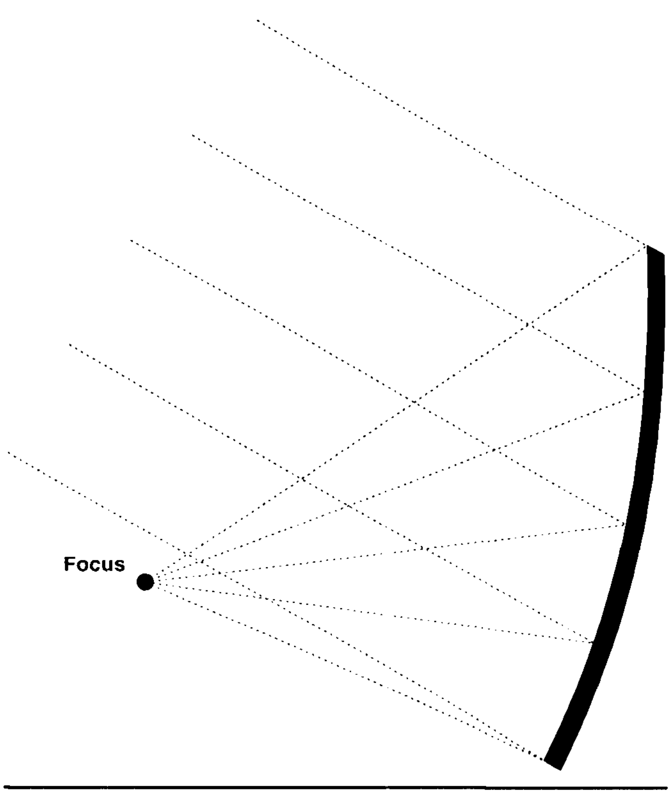


**Fig 2—An offset-fed parabolic dish antenna aimed at a satellite.**

Substituting in the total gain $G_T = (G_1 G_2 \dots G_N)$ results in the total excess noise:

$$G_T = (G_1 G_2 \dots G_N) \qquad \text{Eq 6}$$

with the noise of each succeeding stage reduced by the gain of all preceding stages.

Clearly, if the gain of the first stage, $G_1$, is large, then the noise contributions of the succeeding stages are not significant. This is why we concentrate our efforts on improving the first amplifier or preamplifier.

*Noise Figure*

The noise figure (NF) of an amplifier is the logarithm of the ratio (so we can express it in dB) of the total noise output of an amplifier with an input $T_g$ of 290 K to the noise output of an equivalent noise-free amplifier. A more useful definition is to calculate it from the excess temperature $T_e$:

$$\text{NF} = 10\log\left(1+\frac{T_e}{T_0}\right)\text{dB at } T_0 = 290 \text{ K} \qquad \text{Eq 7}$$

If the NF is known, $T_e$ may be calculated after converting the NF to a ratio, $F$:

$$T_e = (F-1)T_0 = \left[10^{(\text{NF}/10)} - 1\right]T_0 \qquad \text{Eq 8}$$

Typically, $T_e$ is specified for very low-noise amplifiers, where the NF would be fraction of a dB, and NF is used when it seems a more manageable number than a $T_e$ of thousands of kelvins.

*Losses*

We know that any loss or attenuation in a system reduces the signal level. If attenuation also reduced the noise level, we could suppress thermal noise by adding attenuation.

Intuitively, this can't be true. The reason is that the attenuator—or any lossy element—has a physical temperature, $T_x$, that contributes noise to the system while the input noise is being attenuated. The output noise after a loss $L$ (expressed as a ratio, not in dB) is:

$$T_{g'} = \frac{T_g}{L} + \left(\frac{L-1}{L}\right)T_x \qquad \text{Eq 9}$$

If the source temperature $T_g$ is higher than the attenuator temperature $T_x$, the noise contribution is the familiar result found by simply adding the loss in dB to the NF. However, for low source temperatures the degradation can be much more dramatic. If we do a calculation for the effect of 1 dB of loss ($L$ = 1.26) on a $T_g$ of 25 K:

$$T_{g'} = \frac{25}{1.26} + \left(\frac{0.26}{1.26}\right) \times 290 = 80 \text{ K}$$

The resultant $T_{g'}$ is 80 K, *a 5 dB increase in noise power* (or a 5-dB degradation of signal-to-noise ratio for 1 dB of feed line loss). Since noise power = $kT$ and $k$ is a constant, the increase is the ratio of the two temperatures 80/25, or in dB, 10 log (80/25)=5 dB.

*Antenna Temperature*

How can we have a source temperature much lower than ambient? If an antenna, assumed to be lossless, is receiving signals from space, rather than the warm earth, then the background noise is much lower. The background temperature of the universe has been measured as about 3.2 K. An empirical number for a 10-GHz antenna pointing into clear sky is about 6 K, since we must always look through the attenuation and temperature of the atmosphere.[2] The figure will vary with frequency, but a good EME antenna might have a $T_g$ of around 20 K at UHF and higher frequencies.

A couple of examples of actual antennas might bring all of this together.[3]

1. A 30-inch conventional dish at 10 GHz, with a measured gain of 36.4 dBi and efficiency of 64%. The estimated spillover efficiency is 87% for a 10-dB illumination taper. With the dish pointing at a high elevation, as shown in Fig 1, perhaps half of the spillover is illuminating earth at 290 K, which adds an estimated 19 K to the 6 K of sky noise, for a total of 25 K. In a 500-Hz bandwidth, the noise output is −157.6 dBm.

2. An 18-inch DSS offset-fed dish at 10 GHz, with measured gain of 32.0 dB and efficiency of 63%. The spillover efficiency should be comparable, but with the offset dish pointing at a high elevation, as shown in Fig 2, far less of the spillover is illuminating warm earth. If we estimate 20%, then 8 K is added to the 6 K of sky noise, for a total of 14 K. In a 500-Hz bandwidth, the noise output is −160 dBm.

The larger conventional dish has 2.4 dB higher noise output but 4.4 dB higher gain, so it should have 2.0 dB better signal-to-noise ratio than the smaller offset dish when both are pointing at high elevations.

However, while the offset dish is easy to feed with low loss, it is more convenient to feed the conventional dish through a cable with 1 dB of loss. Referring back to our loss example above, the noise temperature after this cable loss is 80 K. In a 500-Hz bandwidth, the noise output is now −152.6 dBm, 7.4 dB worse than the offset dish. The convenience of the cable reduces the signal-to-noise ratio by5 dB, making the larger conventional dish 3 dB worse than the smaller offset dish. Is it any wonder that the DSS dishes sprouting on rooftops everywhere are offset-fed?

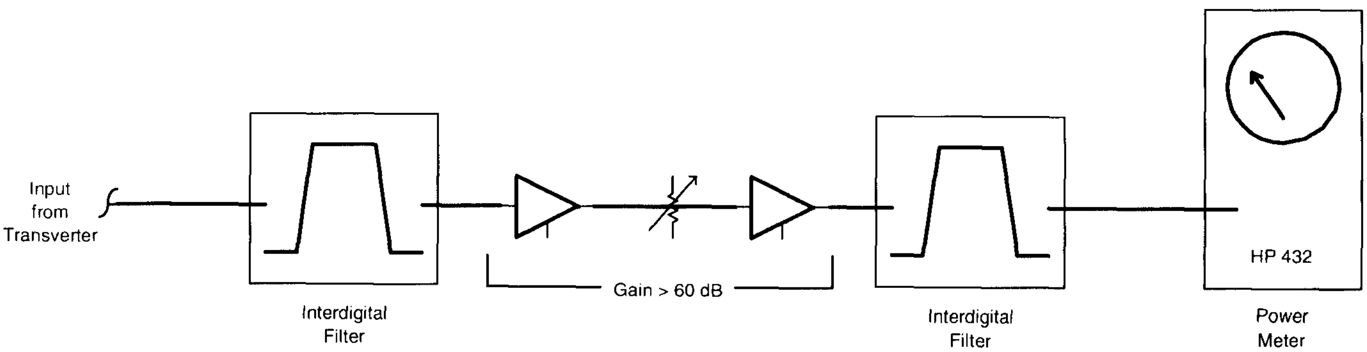


**Fig 3—System for measuring sun noise.**

If the dishes are pointed at the horizon for terrestrial operation, the situation is much different. At least half of each antenna pattern is illuminating warm earth, so we should expect the noise temperature to be at least half of 290 K, or about 150 K. Adding 1 dB of loss increases the noise temperature to 179 K, a 1 dB increase. At higher noise temperatures, losses do not have a dramatic effect on signal-to-noise ratio. In practice, the antenna temperature on the horizon may be even higher since the upper half of the pattern must take a much longer path through the warm atmosphere, which adds noise just like any other loss.

*Image Response*

Most receiving systems use at least one frequency-converting mixer that has two responses, the desired frequency and an image frequency on the other side of the local oscillator. If the image response is not filtered out, it will add additional noise to the mixer output. Since most preamps are broadband enough to have significant gain (and thus, noise output) at the image frequency, the filter must be placed between the preamp and the mixer. The total NF including image response is calculated:

$$NF = 10\log\left[\left(\frac{1+T_e}{T_0}\right)\left(1+\frac{G_{image}}{G_{desired}}\right)\right] \qquad \text{Eq 10}$$

assuming equal noise bandwidth for desired and image responses. Without any filtering, $G_{image} = G_{desired}$ so $1+(G_{image}/G_{desired}) = 2$, doubling the noise figure, which is the same as adding 3 dB. Thus, without any image rejection, the overall noise figure is at least 3 dB *regardless of the NF of the preamp*. For the image to add less than 0.1 dB to the overall NF, a quick calculation shows that the gain at the image frequency must be at least 16 dB lower than at the operating frequency. A filter is a convenient way to reduce gain at the image frequency, but it must be between the preamp and mixer.

## Noise Figure Measurement

So far we have discussed the sources of noise and a figure of merit for evaluating the receiving system's response to noise. How can we measure an actual receiver?

The noise figure of a receiver is determined by measuring its output with two different noise levels, $T_{hot}$ and $T_{cold}$, applied to its input. The ratio of the two output levels is referred to as the *Y-factor*. Usually, the ratio is determined from the difference in dB between the two output levels, $Y_{dB}$:

$$Y_{(ratio)} = 10^{(Y_{dB}/10)} \qquad \text{Eq 11}$$

Then the receiver $T_e$ may be calculated using $Y_{(ratio)}$:

$$T_e = \frac{T_{hot} - Y \times T_{cold}}{Y-1} \qquad \text{Eq 12}$$

and converted to noise figure:

$$NF = 10\log\left(1+\frac{T_e}{T_0}\right)\text{ dB} \qquad \text{Eq 13}$$

where $T_0$ = 290 K

The two different noise levels may be generated separately, for instance by connecting resistors at two different temperatures. Alternatively, we could use a device that can generate a calibrated amount of noise when it is turned on. When such a device is turned off, it still generates noise from its internal resistance at $T_{cold}$, the ambient temperature (290 K); usually this resistance is 50 Ω, to properly terminate the transmission line that connects it to the receiver. When the noise generator is turned on, it produces excess noise equivalent to a resistor at some higher temperature at $T_{hot}$. The noise produced by a noise source may be specified as the Excess Noise Ratio ($ENR_{dB}$), the dB difference between the cold and the equivalent hot temperature, or as the equivalent temperature of the excess noise, $T_{ex}$, which is used in place of $T_{hot}$ in Eq 12. If the ENR is specified, then the calculation is:

$$NF_{dB} = ENR_{dB} - 10\log\left(Y_{(ratio)} - 1\right) \qquad \text{Eq 14}$$

The terms $T_{ex}$ and ENR are used rather loosely; assume that a noise source specified in dB refers to $ENR_{dB}$, while a specification in "degrees" or kelvins refers to $T_{ex}$.

An automatic noise-figure meter, sometimes called a PANFI (*precision automatic noise-figure meter*), turns the noise source on and off at a rate of about 400 Hz and performs the above calculation electronically.[4] A wide bandwidth is required to detect enough noise to operate at this rate; a manual measurement using a narrowband communications receiver would require the switching rate to be less than 1 Hz, with some kind of electronic integration to properly average the Gaussian noise.

Noise-figure meters seem to be fairly common surplus items. The only one in current production, the HP 8970, measures both noise figure and gain but commands a stiff price.

AIL (later AILTECH or Eaton) made several models; the model 2075 measures both NF and gain, while other models are NF only. The model 75 (a whole series whose model numbers start with 75) shows up frequently for anywhere from $7 to $400, typically

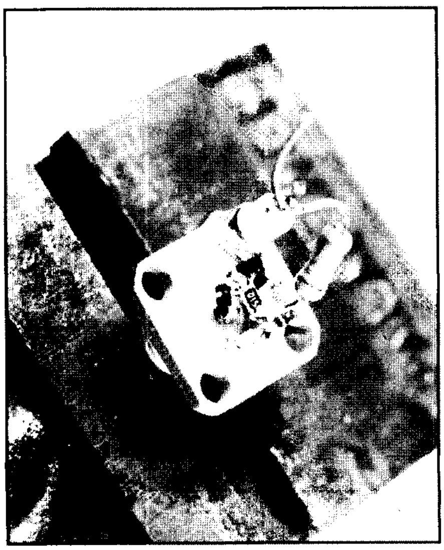

**Fig 4—A noise source built on an SMA connector.**

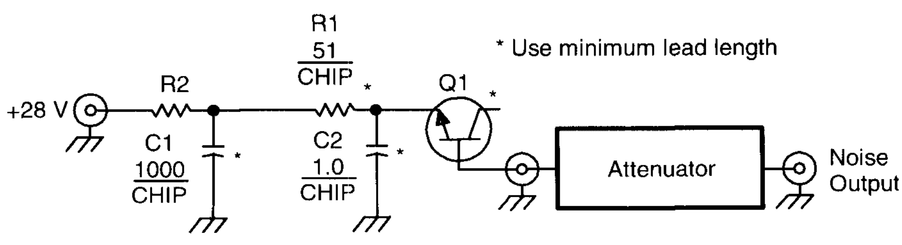


**Fig 5—Schematic diagram of a noise source built on an SMA connector.**

**Q1—Tiny silicon NPN RF transistor such as NEC 68119.**

**R2—Select to set current (see text). 1 kΩ minimum, ¼ W.**

$25 to $50, and performs well. Every VHFer I know has one, with most of them waiting for a noise source to be usable. Earlier tube models, like the AIL 74 and the HP 340 and 342, have problems with drift and heat, but they can also do the job.

Another alternative is to build a noise figure meter.[5]

*Using the Noise-Figure Meter*

I'll describe the basic procedure using the model 75; others are similar, but the more complex instruments will require studying the instruction manual.

Input to almost all noise-figure meters is at 30 MHz, so a frequency converter is required (some instruments have internal frequency converters; except for the HP 8970, I'd avoid using this feature). Most ham converters with a 28-MHz IF work fine, unless the preamp being measured is so narrowband that a megahertz or two changes the NF. The input is fairly broadband, so LO leakage or any other stray signals can upset the measurement—this has been a source of frustration for many users. There are two solutions: a filter (30-MHz low-pass TVI filters are often sufficient) or a tuned amplifier at 30 MHz. Since a fair amount of gain is required in front of the noise figure meter, an amplifier is usually required anyway.

A noise source (which we will discuss in detail later) is connected to the rear of the instrument by a coax cable: a BNC connector marked DIODE GATE provides +28 V for a solid-state noise source, and high-voltage leads for a gas-tube noise source are also available on many versions. The noise-figure meter switches the noise source on and off. The noise output coax connector of the noise source is connected to the receiver input.

The model 75 has four function pushbuttons: OFF, ON, AUTO, and CAL. The OFF and ON positions are for manual measurements: OFF displays the detector output with the noise source turned off, and ON displays the detector output with the noise source turned on. If all is working, there should be more output in the ON position, and a step attenuator in the IF line may be used to determine the change in output, or Y-factor, to sanity-check our results. The knob marked GAIN is used to get the meter reading to a desirable part of the scale in the OFF and ON positions only; it has no effect on automatic measurements.

The AUTO position causes the instrument to turn the noise source on and off at about a 400-Hz rate and to calculate the NF from the detected change in noise. The model 75 has a large green light near the meter which indicates that the input level is high enough for proper operation—add gain until the light comes on. Then the meter should indicate a noise figure, but not a meaningful one, since we must first set the $ENR_{dB}$ using the CAL position. The lower scale on the meter is marked for from 14.5 to 16.5 dB of ENR; adjust the CAL ADJ knob until the reading in the CAL position matches the ENR of the noise source.

If the ENR of your noise source is outside the marked range, read the section below on homebrew noise sources.

Now that we have calibrated the meter for the ENR of the noise source, we may read the noise figure directly in the AUTO position. Before we believe it, a few sanity checks are in order:

- Manually measure the Y-factor and calculate noise figure.
- Insert a known attenuator between the noise source and preamp—the NF should increase by exactly the attenuation added.
- Measure something with a known noise figure (known means measured elsewhere; a manufacturers claim is not necessarily enough).

Finally, too much gain in the system may also cause trouble if the total noise power exceeds the level that an amplifier stage can handle without gain compression. Gain compression will be greater in the on state, so the detected Y-factor will be reduced, resulting in erroneously high indicated NF. The Gaussian distribution of the noise means that an amplifier must be able to handle 12 dB more than the average noise level without compression. One case where this is a problem is with a microwave transverter to a VHF or UHF IF followed by another converter to the 30-MHz noise-figure meter, for too much total gain. I always place a step attenuator between the transverter and the converter which I adjust until I can both add and subtract attenuation without changing the indicated noise figure.

One final precaution: noise-figure meters have a very slow time constant, as long as 10 seconds for some of the older models, to smooth out the ran-

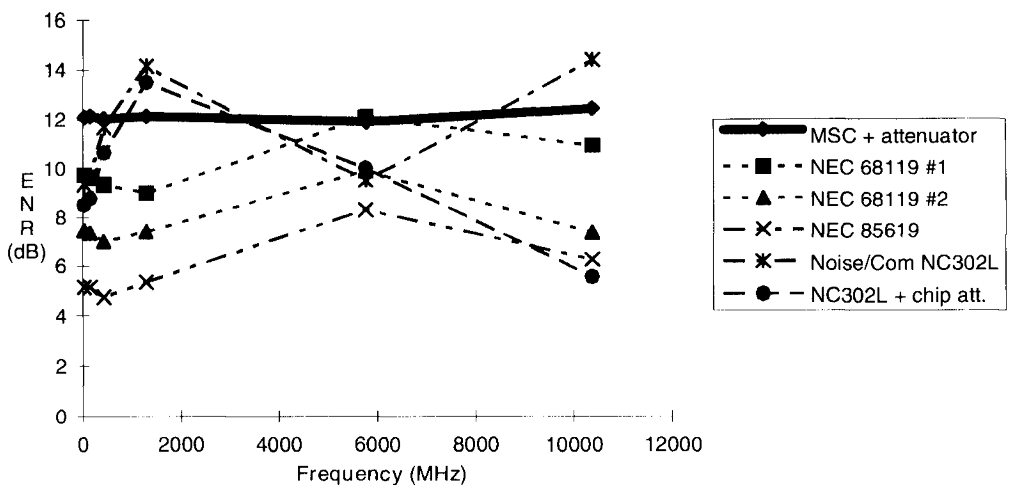


Fig 6—ENR of several versions of the noise source of Fig 5.

dom nature of noise. If you are using the noise-figure meter to "tweak" a receiver, *tune very slowly!*

*Sky-Noise Measurement*

Another way to measure noise figure at microwave frequencies is by measurement of sky noise and ground noise.[3,6] Sky noise is very low, around 6 K at 10 GHz, for instance, and ground noise is due to the ground temperature, around 290 K, so the difference is nearly 290 K. At microwave frequencies we can use a manageable antenna that is sharp enough that almost no ground noise is received, even in sidelobes, when the antenna is pointed at a high elevation. A long horn would be a good antenna choice.

The antenna is pointed alternately at clear sky overhead, away from the sun or any obstruction, and at the ground. The difference in noise output is the Y-factor; since we know both noise temperatures, the receiver noise temperature is calculated using the $Y_{(ratio)}$ and Eq 12.

The latest version of my microwave antenna program, *HDLANT21*, will make this calculation.[3] Since the measured Y-factor will be relatively small, this measurement will only be accurate for relatively low noise figures. On the other hand, they are the most difficult to measure accurately using other techniques.

A system for measuring sun noise was described by Charlie, G3WDG, that also works well for measuring noise figure from sky noise.[7] He built a 144-MHz amplifier with moderate bandwidth using MMICs and helical filters that amplifies the transverter output to drive a surplus RF power meter. The newer solid-state power meters are stable enough to detect and display small changes in noise level, and the response is slow enough to smooth out flicker. Since my 10-GHz system has an IF output at 432 MHz, duplicating Charlie's amplifier would not work. In the junk box I found some surplus broadband amplifiers and a couple of interdigital filters, which I combined to provide high gain with bandwidth of a few megahertz, arranged as shown in Fig 3. I found that roughly 60 dB of gain after the transverter was required to get a reasonable level on the power meter, while the G3WDG system has somewhat narrower bandwidth so more gain is required.

Several precautions are necessary:

• Peak noise power must not exceed the level that any amplifier stage can handle without gain compression. Amplifiers with broadband noise output suffer gain compression at levels lower than found with signals, so be sure the amplifier compression point is at least 12 dB higher than the indicated average noise power.

• Make sure no stray signals appear within the filter passband.

• Foliage and other obstructions add thermal noise that obscures the cold sky reading.

• Low-noise amplifiers are typically very sensitive to input mismatch, so the antenna must present a low VSWR to the preamp.

A noise-figure meter could also be used as the indicator for the sky-noise measurement, but a calibrated attenuator would be needed to determine the Y-factor. Using different equipment gives us an independent check of noise figure so we may have more confidence in our measurements.

W2IMU suggested that the same technique could be used for a large dish at lower frequencies.[8] With the dish pointing at clear sky, the feedhorn is pointing at the reflector, which shields it from the ground noise so it only sees the sky noise. If the feedhorn is then removed and pointed at the ground, it will see the ground noise.

Noise-figure meters are convenient, but if you don't have one, the equipment for measuring sun and sky noise could also be used indoors with a noise source. The only complication is that the Y-factor could be much larger, pushing the limits of amplifier and power meter dynamic range.

## Noise Sources

The simplest noise source is simply a heated resistor—if we know the temperature of the resistor, we can calculate exactly how much noise it is generating. If we then change the temperature, the noise output will change by a known amount. This would work if we could find a resistor with good RF properties whose value does not change with temperature, an unlikely combination. There are commercial units, called *hot-cold noise sources*, with two calibrated resistors at different temperatures and low VSWR. Typically, one resistor is cooled by liquid nitrogen to 77.3 K (the boiling point of nitrogen), while the other is heated by boiling water to 100° C, or 373.2 K. The preamp is connected to first one resistor, then the other; the difference in noise output is the Y-factor.

Since the boiling point of pure liquids is accurately known, this type of noise generator can provide very accurate measurements. However, they are inconvenient to use, since the receiver must be connected directly to alternate resistors (the loss in an RF switch would significantly reduce the

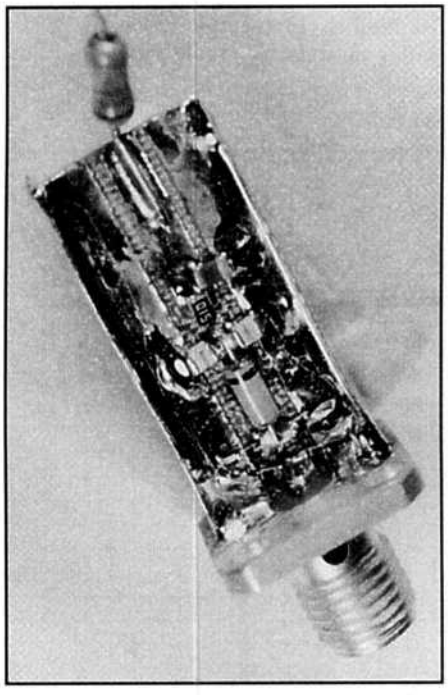

Fig 7—The homebrew noise source of Fig 8.

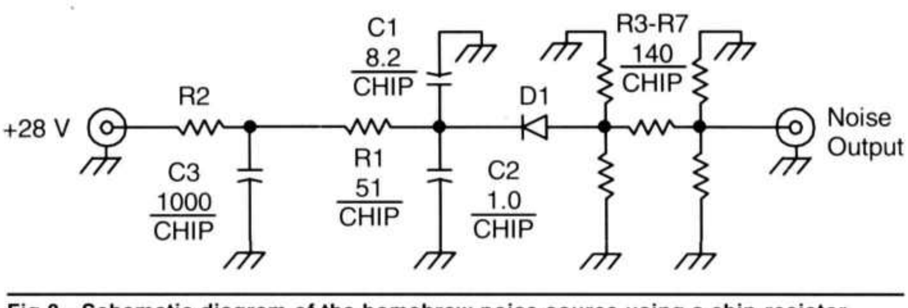


Fig 8—Schematic diagram of the homebrew noise source using a chip-resistor attenuator. C1 and C2 are microwave chip capacitors, ATC or equivalent.

noise output and accuracy). Also, few amateurs have a convenient source of liquid nitrogen.

Three types of noise sources are commonly available and convenient to use:

1. Temperature-limited vacuum tube diode. The noise output is controlled by the diode current but is only accurate up to around 300 MHz due to limitations of the vacuum tube. These units generate around 5 dB of excess noise.

2. Gas tube sources. The noise is generated by an ionized gas in the tube, similar to a fluorescent light—homebrew units have been built using small fluorescent tubes. The noise tubes use a pure gas, typically argon, to control the noise level. These units typically generate about 15 dB of excess noise.

Coaxial gas tube sources work up to around 2.5 GHz, and waveguide units to much higher frequencies. One problem using these is that a high voltage pulse is used to start the ionization (like the starter in a fluorescent light) which is coupled to the output in the coaxial units and is large enough to damage low-noise transistors. Since a noise-figure meter turns the noise source on and off continuously, pulses are generated at the same rate.

Since waveguide acts as a high-pass filter, the starting pulses are not propagated to the output, so waveguide gas-tube noise sources are safe to use, though bulky and inconvenient. However, they could be used to calibrate a solid-state noise source.

Another problem with all gas tubes is that the VSWR of the noise source changes between the on and off states. If the source VSWR changes the noise figure of an amplifier, as is almost always the case, the accuracy of the measurement is reduced.

3. Solid-state noise sources. Reverse breakdown of a silicon diode PN junction causes an avalanche of current in the junction that would rise to destructively high levels if not limited by an external resistance. Since current is "electrons in motion," a large amount of noise is generated. If the current density of the diode is constant, the average noise output should also be constant; the instantaneous current is still random with a Gaussian distribution, so the generated noise is identical to thermal noise at a high temperature. Commercial units use special diodes designed for avalanche operation with very small capacitance for high-frequency operation, but it is possible to make a very good noise source using the emitter-base junction of a small microwave transistor.

Typical noise output from an avalanche noise diode is 25 dB or more, so the output must be reduced to a usable level, frequently 15 dB of excess noise to be compatible with gas tubes or 5 dB of excess noise for more modern equipment. If the noise level is reduced by a good RF attenuator of 10 dB or more, the source VSWR (seen by the receiver) is dominated by the attenuator, since the minimum return loss is twice the attenuation. Thus, the change in VSWR as the noise diode is turned on and off is minuscule. Commercial noise sources consist of a noise diode assembly and a selected coaxial attenuator permanently joined in a metal housing, calibrated as a single unit.

*Homebrew Noise Sources*

There are three components of a noise source: a noise generator, an attenuator and the calibration data—the ENR at each frequency. The most critical component is the attenuator; it is very important that the noise source present a very low VSWR to the preamp or whatever is being measured since low-noise amplifiers are sensitive to input impedance. Even more important is that the VSWR does not change significantly when the noise source is turned on and off since a change causes error in the measurement. Because an attenuator provides twice as many dB of isolation as loss (reflections pass through a second time), 10 dB or more of attenuation will reduce any change in VSWR to a very small value.

Commercial solid-state noise sources occasionally appear in surplus sources, usually at high prices but occasionally very cheap if no one knows what it is. I have found two of the latter, and one of them works! It produces about 25 dB of excess noise, which is too much to be usable. I went through my box of hamfest attenuators and found one that has excellent VSWR up to 10 GHz and 13 dB of attenuation. Mated with the noise source, the combination produces about 12 dB of excess noise—a very usable amount. Finally, I calibrated it against a calibrated noise source for all ham bands between 50 MHz and 10 GHz; not exactly NTIS traceable, but pretty good for amateur work.

While noise sources are hard to locate, noise-figure meters are frequent finds. If we could come up with some noise sources, all the VHFers who have one gathering dust could be measuring and optimizing their noise figure.

Several articles have described construction of homebrew noise sources that work well at VHF and UHF but

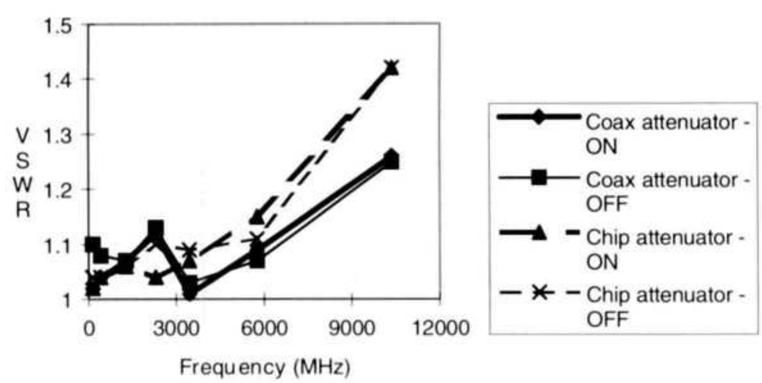


**Fig 9—Measured VSWR of homebrew noise sources of Figs 4 and 7.**

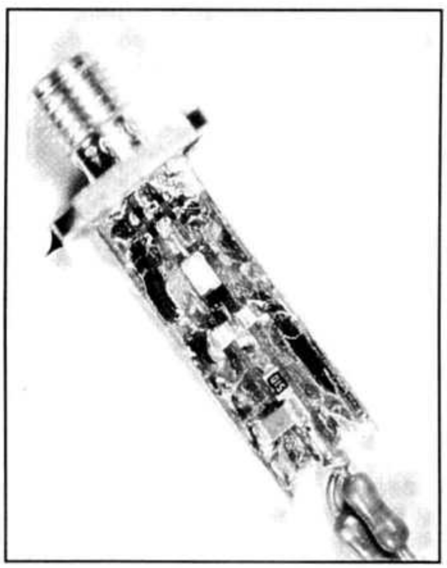

**Fig 10—The noise source of Fig 8, constructed on a photographically printed circuit board.**

not as well at 10 GHz.[9,10,11] All of them have the diode in a shunt configuration, with one end of the diode grounded. When I disassembled my defective commercial noise source (even the attenuator was bad), I found a bare chip diode in a series configuration—diode current flows into the output attenuator. Obviously I could not repair a chip diode, but I could try the series diode configuration. I found the smallest packaged microwave transistor available, some small chip resistors and capacitors, and soldered them directly on the gold-plated flange of an SMA connector with zero lead length, as shown in the photograph, Fig 4. We've all soldered components directly together in "dead-bug" construction; this is more like "fly-speck" construction. The schematic is shown in Fig 5, and it works at 10 GHz! I built several versions to evaluate reproducibility and measured them at several ham bands from 30 MHz to 10 GHz, with the results shown in Fig 6. All units were measured with the same 14-dB attenuator, so the diode noise generator output is 14 dB higher.

(Later I found that the MIT Radiation Laboratory had described a noise source with a series diode 50 years ago, so we aren't giving away anyone's trade secrets.)[12]

I then remembered that I had a commercial noise diode, a Noise/Com NC302L, which was used in a noise source described in *QST*, with the diode in the shunt configuration.[11] The diode is rated as working to 3 GHz, so, in the amateur tradition, I wanted to see if I could push it higher, using the series configuration. Since I didn't expect to reach 10 GHz, I increased the value of the bypass capacitor, but otherwise, it looks like the units in Fig 5. When I measured this unit, it not only worked at 10 GHz, but had more excess noise output than at lower frequencies, probably due to an unexpected resonance. The performance is shown in Fig 6 along with the other units.

Also shown in Fig 6 is the output of my pseudo-commercial noise source; even with the external attenuator, the excess noise output is pretty flat with frequency. Commercial units are typically specified at ±0.5 dB flatness. In Fig 6, none of the homebrew ones are that flat, but there is no need for it; as long as we know the excess noise output for a particular ham band, it is perfectly usable for that band.

All the above noise sources rely on a coaxial microwave attenuator to control the VSWR of the noise source. Attenuators are fairly frequent hamfest finds, but ones that themselves have good VSWR to 10 GHz are less common, and it's hard to tell how good they are without test equipment. An alternative might be to build an attenuator from small chip resistors. I used my PAD.EXE program to review possible resistor values, and found that I could make a 15.3-dB π attenuator using only 140-Ω resistors if the shunt lugs were formed by two resistors in parallel, a good idea to reduce stray inductance.[13] I ordered some 0402-size (truly tiny) chip resistors from DigiKey and more NC302L diodes from Noise/Com, and built the noise source shown in Fig 7 on a bit of Teflon PC board, cutting out the 50-Ω transmission line with a hobby knife. The schematic of the complete noise source is shown in Fig 8.

The chip-resistor attenuator works nearly as well as an expensive coaxial one. The measured VSWR of two noise sources, one with the chip attenuator and the other with a coaxial attenuator, is shown in Fig 9. Curves are shown in both the off and on states, showing how little the VSWR changes. The VSWR of the chip attenuator unit is 1.42 at 10 GHz, slightly over the 1.35 maximum specified for commercial noise sources, but still fine for amateur use.

Still, I wondered if I could do even better. The hand-cut board used is 0.031-inch-thick Teflon material, which is a bit thick at 10 GHz. I obtained some 0.015-inch-thick material and made a photo mask to print an accurate 50-Ω line.[14] Then I carefully assembled the components under a microscope (see sidebar: Enhance Miniature Construction with Optical Feedback). Fig 10 shows the construction: the thin PC board is supported by a thin brass strip soldered along each side to create miniature I-beam, a much sturdier structure. The brass strips also connect the top and bottom ground areas of the board.

I built four units like the one shown in Fig 10, with encouraging results. The ENR of these units, shown in Fig

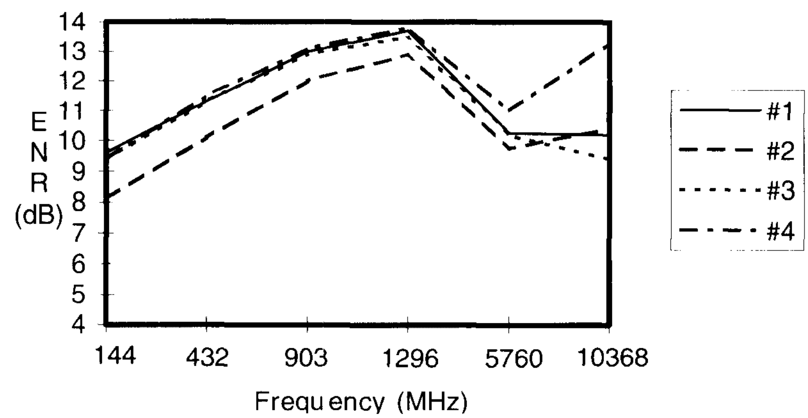


**Fig 11—ENR of the PC-board noise sources (Fig 10).**

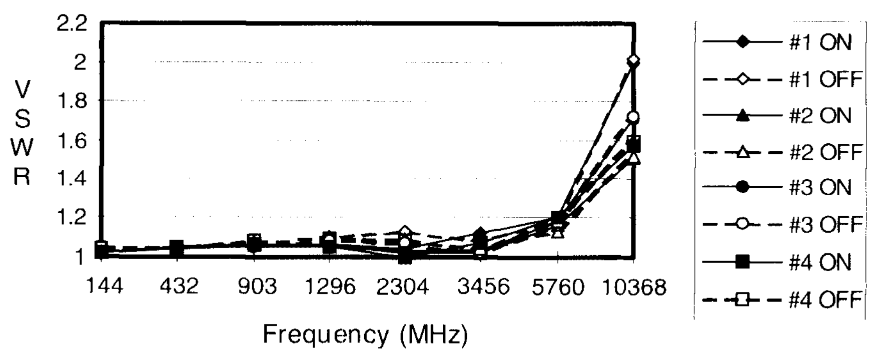


**Fig 12—Measured VSWR of the PC-board noise sources (Fig 10).**

## Enhance Miniature Construction with Optical Feedback

Many microwave construction projects, and most modern equipment, use extremely small surface-mount components. Working with these parts requires steady hands and good vision. And as we get older, our vision usually deteriorates—I got my first bifocals last year.

I'm convinced that the key to working with tiny parts is to see them well. When I built the first noise source with the tiny 0402-size (1 x 0.5 mm) chip resistors, it was frustrating trying to get the resistors soldered where I wanted them. After that experience, I was on the lookout for a surplus stereo microscope and finally located one at a reasonable price. These microscopes are commonly used for microelectronics assembly work, providing moderate magnification at a long working distance.

After I set up the microscope on my workbench with adequate illumination, I was ready to build some more noise sources. Now the tiny chip resistors were clearly visible, and I was able to hold them in place with tweezers while soldering them exactly where I wanted them. On another project, I wanted a clearance hole in the ground plane around a hole drilled though a Teflon PC board. The hole is 0.025 inches in diameter, and I was able to cut an octagon around it with a hobby knife; the length of each side of the octagon is about the same as the hole diameter. Then I lifted the unwanted copper with the point of the knife. Magnification makes miniature work feel precise and easy instead of clumsy and frustrating.

What the microscope does is add gain to the feedback loop from the eyes to the hand. Our hands are never perfectly steady, but adding this feedback steadies them under the microscope, as the brain takes input from the eyes and automatically compensates (after a bit of practice).

A microscope is an elegant solution for very small parts, but any optical magnification helps. I have also used magnifiers, jewelers loupes and "drugstore" reading glasses. If the reading glasses are stronger than you need, they will provide additional magnification; just don't try walking around wearing them.

Other aids to miniature work are tweezers, fine-point soldering irons and lots of light. When an object is magnified, proportionally more light is required for the same apparent brightness. Tweezers help in holding small objects—I prefer the curved #7 style Swiss tweezers, of stainless steel so solder won't stick. Finally, a temperature-controlled soldering iron prevents overheating, which can destroy the solder pads on surface-mount components; 700° F tips are hot enough. All the tools I've mentioned came from hamfests, surplus places or flea markets, at reasonable prices.

So, even if you think that microwave project with tiny parts is beyond your capability, use a magnifier and give it a try. I'll bet you surprise yourself.—*N1BWT*

11, was higher at 10 GHz than the hand-cut one and reasonably flat with frequency—and consistent from unit to unit. The VSWR, however, was still high at 10 GHz, as shown in Fig 12. It is difficult to make a really good coax-to-microstrip transition at 10 GHz! Ordinarily, in an amplifier, we simply tune out the slight mismatch as part of the tuning procedure, but broadband tuning is much more difficult. As a final improvement, I dug up some 5-dB SMA attenuators from the swap session at Microwave Update last year. Adding one of these to the worst unit in Fig 12 reduced the VSWR to 1.18 at 10 GHz (below 1.10 at lower frequencies) and the ENR to 5.0 dB. This performance is every bit as good as a very expensive commercial noise source, lacking only NTIS-traceable calibration.

*Noise-Source Alignment*

The only alignment requirement for a solid-state noise source is to set the diode current; the current is always set at the highest frequency of interest. A noise figure meter must be set up with converters, etc, for the highest frequency at which the noise source might be used and set to display the detector output (OFF position on a model 75). Then voltage from a variable dc power supply is applied to the noise diode through the 1-kΩ current-limiting resistor. The detector output should increase as the voltage (diode current) increases, reach a peak, then decrease slightly. The optimum current is the one that produces peak output at the highest frequency (I set mine at 10 GHz). The optimum current for the NC302L noise sources was between 7 and 12 mA. Then additional resistance must be added in series with the current-limiting resistor so that the peak output occurs with 28 volts applied, so that the noise source may be driven by the noise-figure meter. Once the proper resistor is determined and added, the dc end of the noise source is connected to the diode output of the noise-figure meter and the meter function is set to ON. This should produce the same detector output as the power supply.

Then the meter function is set to AUTO and the meter should produce some noise-figure indication, but not yet a calibrated one. However, it is good enough to tune up preamps—a lower noise figure is always better, even if you don't know how low it is.

*Noise-Source Calibration*

Much of the high price of commercial noise sources pays for the NTIS-traceable calibration. Building a noise source only solves part of the problem—now we need to calibrate it.

The basic calibration technique is to measure something with a known noise figure using the new noise source, then calculate what ENR would produce the indicated noise figure.

Fortunately, the calculation is a simple one involving only addition and subtraction; no fancy computer program required. Simply subtract the indicated noise figure, $NF_{indicated}$, from the known noise figure, $NF_{actual}$, and add the difference to the ENR for which the meter was calibrated, $ENR_{cal}$:

$$ENR_{(noise\ source)} = ENR_{cal} + (NF_{actual} - NF_{indicated}) \quad \text{Eq 15}$$

This procedure must be repeated at each frequency of interest; at least once for each ham band should be fine for amateur use.

The known noise figure is best found by making the measurement with a calibrated noise source, then substituting the new noise source so there is little opportunity for anything to change. Next best would be a sky noise measurement on a preamp. Least accurate would be to measure a preamp at a VHF conference or other remote location, then bring it home and measure it, hoping that nothing rattled loose on the way. If you can't borrow a calibrated noise source, it would be better to take your noise source elsewhere and calibrate it. Perhaps we could measure noise sources as well as preamps at some of these events.

*Using the Noise Source*

Now that the ENR of the noise source has been calibrated, the noise-figure meter calibration must be adjusted to match. However, the model 75 in the CAL position has only 2 dB of adjustment range marked on the meter scale. Older instruments have no adjustment at all. However, we can just turn around the equation we used to calculate the ENR and calculate the NF instead:

$$NF_{actual} = NF_{indicated} + \left(ENR_{(noise\ source)} - ENR_{cal}\right) \qquad \text{Eq 16}$$

There is a short cut. My noise source has an ENR around 12 dB, so I set the CAL ADJ in the CAL position as if the ENR were exactly 3 dB higher, then subtract 3 dB from the reading. Even easier, the meter has a +3DB position on the ADD TO NOISE FIGURE switch. Using that position, I can read the meter starting at 0 dB. Any ENR difference from 15 dB that matches one of the meter scales would also work—rather than an involved explanation, I'd urge you to do the noise figure calculations, then try the switch positions and see what works best for quick readout.

Reminder: Noise-figure meters have a very slow time constant, as long as 10 seconds for some of the older models, to smooth out the random nature of noise. *Tune slowly!*

Don't despair if the ENR of your noise source is much less than 15 dB. The optimum ENR is about 1.5 dB higher than the noise figure being measured.[1] The fact that today's solid-state noise sources have an ENR around 5 dB rather than the 15 dB of 20 years ago shows how much receivers have improved.

## Conclusion

The value of noise-figure measurement capability is to help us all to "hear" better. A good noise source is an essential part of this capability. Accurate calibration is not necessary but helps us to know whether our receivers are as good as they could be.

### Notes


[1] Pettai, R., *Noise in Receiving Systems*, Wiley, 1984.
[2] Graves, M. B., WRØI, "Computerized Radio Star Calibration Program," *Proceedings of the 27th Conference of the Central States VHF Society*, ARRL, 1993, pp 19-25.
[3] Wade, P., N1BWT, "More on Parabolic Dish Antennas," *QEX*, December 1995, pp 14-22. The *HDLANT21* program may be downloaded from **http://www.arrl.org/qexfiles**.
[4] Pastori, W. E., "Direct-Reading Measurement of Receiver-Noise Parameters," *Microwave Journal*, April 1973, pp 17-22.
[5] Bertelsmeier, R., DJ9BV, and Fischer, H., DF7VX, "Construction of a Precision Noise Figure Measuring System," *DUBUS Technik 3*, DUBUS, 1992, pp 106-144.
[6] Fasching, H., OE5JFL, "Noise Figure Measurement using Standard Antennas," *DUBUS Technik 4*, DUBUS, 1995, pp 23-25.
[7] Suckling, C., G3WDG, "144 MHz wideband noise amplifier," *DUBUS 2/1995*, DUBUS, 1995 pp 5-8.
[8] Turrin, R., W2IMU, "Method for Estimating Receiver Noise Temperature," Crawford Hill Technical Note #20, September 1986.
[9] Britain, K., WA5VJB, "10 GHz Noise Source," *Microwave Update '87*, ARRL, 1987, p 63.
[10] Wade, P., N1BWT, and Horsefield, S., NR1E, "Homebrew Solid State Noise Sources," *Proceedings of the 1992 (18th) Eastern VHF/UHF Conference*, ARRL, 1992.
[11] Sabin, W. E., WØIYH, "A Calibrated Noise Source for Amateur Radio," *QST*, May 1994, pp 37-40.
[12] Valley, G. E., Jr., and Wallman, H., *Vacuum Tube Amplifiers*, MIT Radiation Laboratory Series, McGraw-Hill, 1948.
[13] Wade, P., N1BWT, "Building VHF Power Attenuators," *QEX*, April 1994, pp. 28-29. The PAD.EXE program may be downloaded from **http://www.arrl.org/qexfiles**.
[14] PC boards are available from Down East Microwave.

# *Design Notes for "A Luxury Linear" Amplifier*

*Here are some of the design considerations that went into a recent amplifier project, along with an additional experimental circuit and some interface suggestions.*

By Mark Mandelkern, K5AM

This article is a companion to the article "A Luxury Linear," which included a general description of the circuits, performance specifications and all the schematics for a 1500-W 2-m linear amplifier.[21] But it gave few particulars explaining the functioning of the circuits or the special roles of the various components. This article will fill in most of these details. One circuit in the amp, of a more experimental nature, was omitted from the previous article and will be described here. This is the heater idle circuit, which drops the heater voltage 12% during very long standby periods—aiming to dramatically increase tube life. Finally, some problems of interface with other station components will be discussed, including driver IMD, coax relay sequencing and ALC.

[1]Notes appear on page 20.

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### RF Circuits

Most of the details for the RF circuits are given in the caption to Fig 4.[22] Here are a few additional suggestions. There can be quite a bit of RF voltage in the plate tank. The swinging link with its Teflon sleeving should stay, throughout its travel, at least 6 mm away from the tank coil. For avoiding excessive RF voltage and arcing, it is most important to keep the loading sufficiently heavy, especially on initial tune-up. The dc-grounded link is a major safety feature. There is no sliding connection at any point of the output circuit; this is important for efficiency and to prevent erratic tuning.

The only difficult item in the output circuit is the link tuning capacitor. There was arcing in the first capacitor I tried, which had less spacing. If you find a capacitor as specified, except with wider spacing, so much the better. But you don't want less capacitance—that would mean more RF voltage. Too much capacitance would limit the effectiveness of the link tuning capacitor as a loading control. To minimize the RF voltage across the plates, I try to keep the link tuning capacitor near maximum capacitance. I use the swinging link for coarse-loading adjustment and the link tuning capacitor for fine-loading adjustment. This works well. Use the link tuning capacitor for loading; if you approach maximum, nudge the link in a bit, and if you find you are down towards half capacitance, ease the link out a bit. Sounds complicated, but I haven't touched the swinging link in the past six months. Another advantage of this loading procedure is that adjusting the link tuning capacitor has very little detuning effect on the tank coil, whereas swinging the link has a noticeable effect.

The voltage divider for HV metering provides a 10,000 to 1 sampling ratio at the HV2 point. The 20-MΩ resistor in Fig 4 consists of two 10-MΩ special glass HV types, with neat spiral resistive traces wound inside the glass tubes. These types are found now and then in the surplus catalogs. A string of twenty 1-MΩ, 1/4-W, 1% metal-oxide resistors would be suitable (Digi-Key #1.00MXBK).[20] The resistors are specified as 1% types not because accuracy is required here, since there is a calibrating adjustment in the metering circuit (Fig 10), but because the precision types will be more stable, thermally and over time.

Don't miss the note on C5 in Fig 4, about not connecting any additional bypass capacitor at the screen terminal. The 100-Ω resistor has an important decoupling function.

It may be possible to reduce input drive requirements below 30 W. Owners of 25-W transceivers might be especially interested. A heavy, wide silver-plated strap for the T-match coil would be the first thing to try. I did try a heavier wire in a hairpin loop, with no improvement. That's when I concluded that most of the loss was inside the tube. My driver is capable of 200-W output. So, except for trying to hit the magic 25-W input spec, I had little motivation to work further on the input circuit. The T-match tuning capacitors might be one place to try for improvement. Glass piston trimmers might be better. Trying to read RF voltages at each of the three terminals was wild! One high, one middling (the one fed), one almost nothing. Adding a heavy wire connecting the three grid terminals didn't help. Feeding just one grid terminal doesn't seem to make the other two jealous. The geometry of the socket and the tube base indicates that it may be pointless to worry about this asymmetry; the solid grid ring built into the tube base has less inductance than anything that could be built around the socket. This amp shows a noticeable improvement over the old 1000-W amp; I heard my EME echoes with a horizontal single Yagi at moonset—this is not a first, but it might be for an all-homebrew station.

## Power-Supply Circuit

Most of the power supply is straightforward and routine. The heater regulator uses the ubiquitous 723 IC. A minor complication arises from having both sides of the heater above ground, due to the cathode current shunt and the construction of the tube, with the cathode internally connected to one side of the heater. The heater voltage is applied to the A+/A– points, while the S+/S– points provide heater voltage sensing at the socket. Thus the entire regulator circuit is referenced to the S– point. Current for the 723 does flow through the cathode metering shunt, but it hardly moves the plate meter. The heater power supply is returned to the A– point. As noted in "A Luxury Linear" (Note 21), the heater supply did not reach the 90- to 130-V ac operating goal, dropping out at 98 V. I was content with that and did not try to improve it, but it should be easy to do so. I used #18 wire for the A+/A– leads. Since the regulator senses voltage at the socket, drop in these leads is of no concern; but heavier wire could be used to improve low-line-voltage performance. I think, though, that the greatest drop may be in the heater fuse holder, an ordinary 3AG type—it gets pretty hot! An automotive type blade fuse and socket will probably have less voltage drop.

The choice of heater transformer is the main factor determining line regulation, however. I used a Signal #36-6. Rated 18 VCT at 12 A, it is much heavier than needed in the FWCT circuit used. But on the surplus market (searching 20 catalogs) it cost less than the next smaller size available, which was a bit too small.

## Screen Regulator

Not so commonly discussed, and of special interest (judging from inquiries) is the screen voltage shunt regulator. Power tetrodes, and especially the 4CX1000A, commonly exhibit negative screen current flow due to secondary emission. Feisty electrons from the cathode hit the screen and knock off more electrons, even more than those arriving. Like, throw one ping-pong ball, forcefully, into a bucket of ping-pong balls, and see a dozen bounce out. To the amp builder (who, in my case at least, disavows knowing anything about what's going on inside the tube) it seems as if current is coming out of the tube. (Current in my shack flows from positive to negative.) This negative screen current flows into the screen supply and tends to increase the screen voltage, which if unchecked will destroy the tube. (The actual source of the negative screen current is the plate supply.)

A series regulator will not suffice to deal with this negative screen current problem; only a shunt regulator will do. VR tubes are traditional, and Zener diodes have been recently in favor. I wanted a fully adjustable supply to enable experimentation with different operating parameters. I also felt that power transistors are inherently more reliable than Zeners. And the transistors give you more watts per dollar. Also, the transistors used operate very

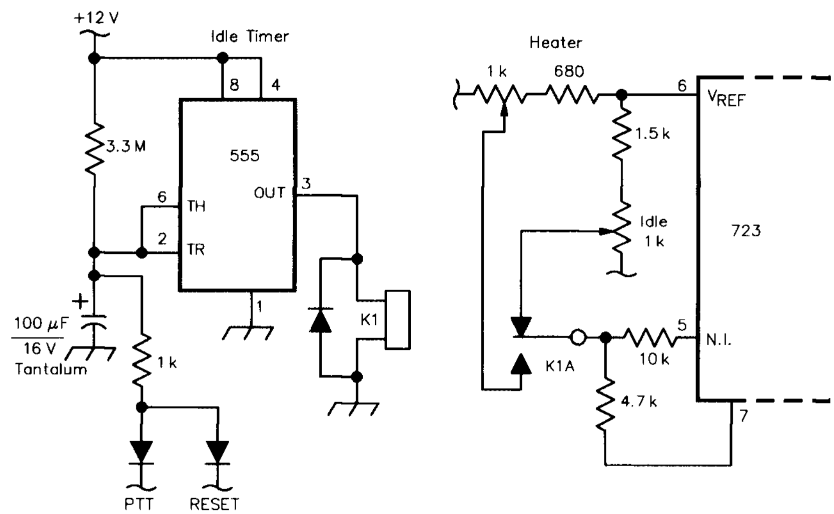


**Fig 11—Schematic diagram of the heater-idle circuit. This is a modification of a portion of the heater-regulator circuit in Fig 5. Refer also to the caption for Fig 6.**

far from their maximum ratings while Zeners, in this application, typically operate very near their ratings. The shunt transistor regulator was only a bit of extra trouble and has worked flawlessly.

Well, in fact, the first shunt regulator tried, using a 723, didn't work at all! It worked fine on the bench, in all positive and negative current tests, but the shunt transistor shorted when I put the amp in the rack and fired it up. It's best to write about what works, rather than what didn't work, but in this case the story includes a few warnings and explains the reason for some components in the final circuit. I made a half dozen changes, and the second regulator, as shown in Fig 5, performed perfectly from the start. It's hard to say exactly what went wrong with the first circuit, but here are the guesses: oscillation in the 723, possibly triggered by the very jumpy screen current variations during CW, SSB or pulse tuning operation. RF getting into the sensitive 723? (That would explain erratic behavior, but would not be enough by itself to cause the transistor to short out.) Excessive current in the shunt transistor. (In a 5-A transistor?—Note the energy stored in the 22-µF electrolytic capacitor before the 3-kΩ resistor was added.) Puncturing of the mica under the shunt transistor—even two micas. (At 300 V?—What's going on here?—I wish I understood transients better.) Dynatron oscillation? (See page 54 in the reference in Note 3.) I never took a course in electronics, so I can only guess what true failure analysis in an industrial environment would involve: digital storage scopes, chart recorders and a bucket full of transistors (paid for by the company). So, after losing three of the four transistors I had on hand, all I could do was go on hunches and try something a bit different.

The screen regulator that works doesn't use a 723, but merely a single, fairly high-gain transistor as driver for the shunt transistor. A certain amount of gain is necessary, but more than is needed might bring in stability problems. The Zener to the left of the 2N2222A in Fig 5 is merely to provide collector voltage within the transistor ratings; using a high-voltage transistor here would have meant lower gain. The other 15-V Zener is the regulator reference, along with two $V_{be}$ drops in the transistors. Say the reference is then 16.2 V. The output voltage will be this reference multiplied by the ratio set up by the voltage divider at the output, including the screen voltage adjustment pot on the front panel. This ratio having the range 12 to 22, the nominal output voltage is about 200 to 360 V.

There are two details that may be crucial. I used no micas, but mounted the TO220 shunt transistor directly onto a 2×3×1/8-inch aluminum plate, which in turn is mounted on the side wall using ceramic standoff insulators. Second, I added the 3-kΩ, 20-W resistor in the shunt transistor collector lead. This limits the peak (transient?) shunt-transistor current to about 100 mA (at 300 V) and prevents a possible momentary saturated crowbar short to ground. The resistor could be lower valued, and sink more current, but the amp includes an overload circuit set to 30-mA negative screen current, so a regulating range of 100 mA is adequate. Normal operation results in about 20-mA negative screen current peaks during SSB operation. The 0.1-µF capacitor on the collector and the 22-µF capacitor at the output are intended to soften transient pulses which were thought to make the transistor unhappy.

The regulation obtained is only about 1%. This is intentional, as I was most concerned with stability, which is inversely related to regulation. The degree of regulation depends on the gain of the transistors and the resistance of the voltage divider at the output since base current for the first transistor flows through the 120-kΩ resistor. Lowering the divider resistance would improve the regulation, but 1% regulation is more than adequate. It's sometimes difficult, in these cyberdays, to remember that the super-precision available is not always needed—or desired.

From the E = 400-V supply, the R = 5-kΩ (10-W) series resistor limits the forward screen power to $P = E^2/4R$ = 8 W, well within the 12-W rating. Thus no forward screen current overload circuit is needed.

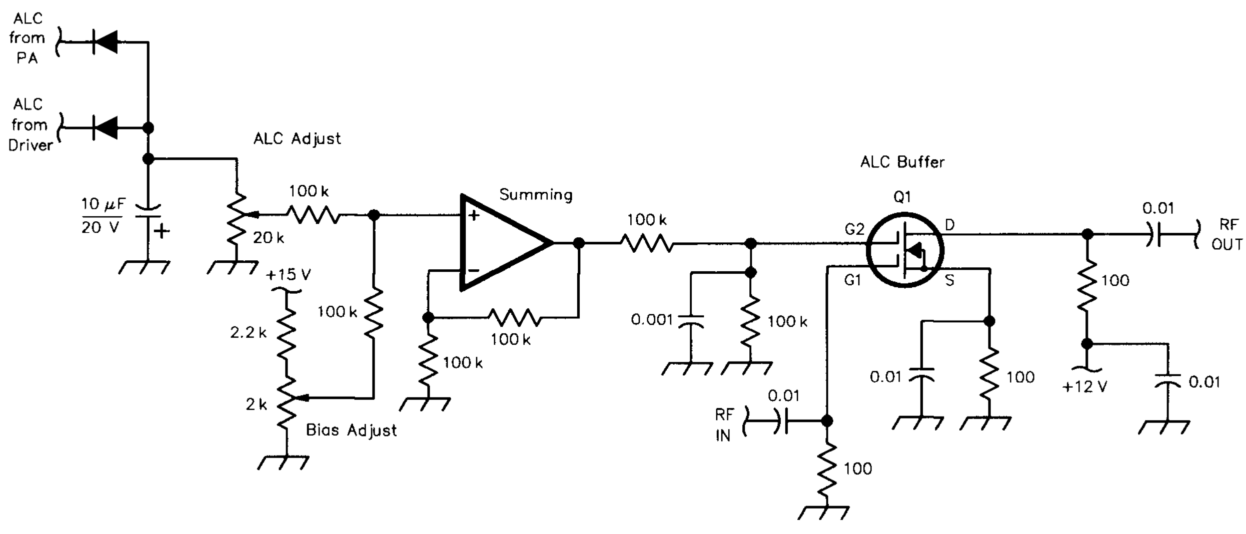


Fig 12—Schematic diagram of an ALC circuit that may be added to a transceiver or transverter. The MOSFET Q1 may be any one of dozens of small-signal RF types (40673, 3N211, ...) often found on the surplus market. The NTE 454 replacement type is available from Hosfelt.[19]

## Control Circuits

The warm-up timer in Fig 6 is powered by the +12H rail, only after the air-flow switch, at terminal AIR, is actuated. Several seconds are required for the blower to get up to speed after the line switch is thrown; then the blue light comes on and the timer starts. But only after voltage appears at the heater. This is sensed by the op amp connected to the S+ terminal. So, for example, if the fuse is blown, the timer will not start until after you replace the fuse.

The cool-down latch in Fig 6 is a bit tricky. The basis of operation is the fact that when the line switch is thrown off, the +12-V rail drops more quickly than the +24-V rail (which has minimal load except while transmitting). This throws the latch high and starts the cool-down timer. Since the timer starts in the low state, the discharge pin at the COOL terminal is low, and the cool-down relay K3 in Fig 5 is energized. After about three minutes the timing capacitor charges enough to bring the trigger terminal low and set the 555 high, opening the discharge terminal and allowing the cool-down relay to drop out.

An unexpected anomaly showed up when this simple timer was first tried. Not a very serious defect, but perhaps an amusing incident in the quest for perfection. Here is the situation. After working a few choice DX stations, the band seems dead and you throw the line switch off. Before the cool-down circuit times-out, you hear another DX station, change your mind, and throw the line switch ON again. After allowing the tube to warm-up (see page 8 in "A Luxury Linear"), you work the station. Congratulations! Now you throw the line switch off and are surprised to hear the blower immediately stop! What happened? The cool-down circuit was in COOL mode and timed-out while you were working the new one. That's the reason for the connection to the +24RLY line. What does the cool-down circuit have to do with the coax relays? The +24RLY line resets the cool-down timer (if latched high) the next time you transmit; this was the simplest band-aid solution.

The regulated heater supply, with 10-V at the unregulated point, made over voltage protection (OVP) absolutely essential, to protect against a short-mode failure in the pass transistor Q1 in Fig 5. After hearing of troubles with fold-back and SCR methods of OVP, and considering the various failure modes, an op-amp circuit and a relay seemed to afford the highest reliability. At the bottom of Fig 6, the first op amp acts as a comparator. When the heater voltage exceeds the preset limit, the op-amp goes low and pulls the OVP latch low, removing the ground return for the heater relay coil at point HR in Fig 5. The 10-μF capacitor in the OVP circuit ensures that the OVP latch goes high on start-up when the line switch is thrown ON.

The 2.2-μF capacitor in the overload circuit (Fig 9) forces the overload latch low (normal) on power-up. When low, the latch holds the flash oscillator high and the red light off. I suppose the inverter could have been avoided if, when I got to that point, surprised to find myself standing on my head, I had gone back and redesigned the entire control board. But a low parts count was not a major design goal.

The air-flow switch protects the tube in three ways. Not only does it control the supply voltage (+12H) to the 723 in the heater regulator, and by the same line enable the operate circuit in Fig 7, but it also controls the coil voltage to the heater relay in Fig 5.

### Coax Relay Sequencing

VHF operators are especially interested in antenna-relay sequencing devices. In my station (all homebrew), sequencing is included internally in the transceiver, each transverter, each driver amp and each kW amp. This provides the highest level of protection. An MGF-1402 GaAsFET preamp is still working perfectly after ten years with the original transistor. The sequencing circuits used here may be incorporated in other transceivers and transverters for fool-proof coax relay and preamp protection. The sequencing circuits are very simple and much more reliable than many excessively complex systems.

The two op amps shown in the sequencing circuit in Fig 7 use very simple R-C networks to obtain switching delays. Notice that the diodes are reversed. The PTT op amp shifts to an output of about +11 V in transmit. At the coax relay op amp, the diode allows this to charge the capacitor almost instantly, so the coax relay can begin to close. But at the screen relay op amp, the positive control signal must pass through the 68-kΩ resistor, so the screen relay is delayed. At the end of the transmission, the PTT op amp shifts back to an output of about −11 V, and the procedure is reversed.

Keying the amplifier on CW requires special care to avoid any modification of the keying waveform produced by the transceiver. Assuming that the transceiver has the desired keying waveform (my homebrew transceiver has separate internal adjustments for make and break), we want the amplifier to pass this along to the antenna without alteration. The keying circuit at the bottom of Fig 7

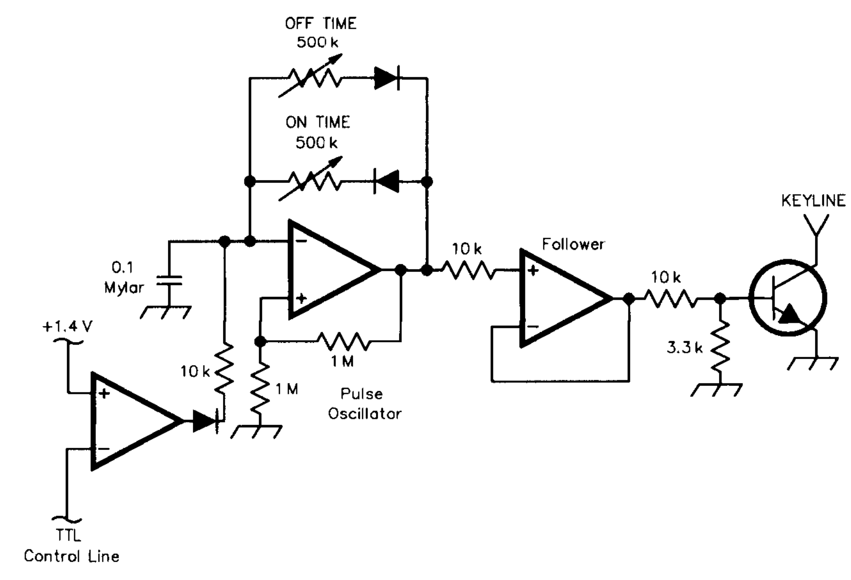


**Fig 13—Schematic diagram of the pulse-tuning circuit used for low-duty-cycle tune-up and adjustment. Refer to the caption for Fig 6.**

provides for this. The diode allows the key closure signal, a negative output from the first op amp, to almost immediately key the bias circuit (at the KEY2 point), so the amp is up and ready for the dit when the transceiver finally gets around to sending it (about 2 ms later). When the key opens, however, the positive output from the op amp must force its way through the 6.8-kΩ resistor to charge the cap. This allows sufficient time to keep the amp keyed while the dit is decaying. The second op amp, with hysteresis provided by the 1-MΩ feedback resistor, switches sharply when the capacitor crosses the zero potential point. This is a fast-switching delay circuit; it does not produce a keying waveform. It has a square-wave output which encloses the transceiver keying waveform, allowing the transceiver alone to shape the transmitted CW element.

The key sequencing also has a connection to the screen-relay sequencing circuit. This makes the keying dependent on the operate, PTT, sequencing and overload circuits, for an extra bit of protection.

In connecting the amplifier key line to the station, the easiest way to avoid unwanted interaction would be to build an interface circuit that allows the station keyer to separately, but simultaneously, key the transceiver and amplifier. However, for many transceivers it may be possible to simply tie the key lines together. The transceiver schematic must be checked; can it handle 12 V on the keyline? Can an isolating diode be added?

### Metering Circuits

This was the really fun part! Imagine: training op amps to do arithmetic! The op amps here are used in three ways: the noninverting amplifier, the inverting summing amplifier with virtual ground and the differential amplifier. These amplifier circuits are described fully in Chapter 8 of the *Handbook*.[7] In the following, refer to Figs 8.44 and 8.46 to 8.49 in the *Handbook*, along with the associated formulas and notation.

The simplest noninverting amplifier used here is the one for HV metering. The meter is to read 5000-V full-scale, and the divider inside the plate compartment (Fig 4) yields a 10,000 to 1 ratio, so 5000 V would produce 0.5 V at the HV2 point in Fig 10. The 10 kΩ resistor in the noninverting (+) input, and the 10 kΩ resistor inside the plate compartment (which is for filtering, in conjunction with feed-through and bypass capacitors not shown on the diagram) have a negligible effect on the circuit, as the op amp input is of very high impedance. Our op amps on the +12/–12 rails have maximum outputs of about +11/–11 V, so we ask for +10 V output at full-scale, using a 10-kΩ meter multiplier resistor and a 1-mA meter. We'll need a gain of 10/0.5 = 20. Arbitrarily choosing 10 kΩ as input resistor $R_i$, we calculate 190 kΩ for the feedback resistor $R_f$. I usually choose $R_i$ = 10 kΩ because I have a whole box of 10-kΩ resistors. For $R_f$, I choose a trimmer at about half the nominal value, and then a fixed resistor, chosen from standard values, to obtain the best range. In this case, a 100-kΩ trimmer and a 150-kΩ fixed resistor yield a range of 150 kΩ to 250 kΩ. That's a fairly good bracketing of the 190-kΩ goal; it doesn't usually work out so well. A 200-kΩ trimpot would not permit sufficient adjustment range, allowing for the tolerances of the circuit components. So omitting the fixed resistor would mean using a 500-kΩ trimpot and having a very touchy adjustment situation.

Plate-current metering is similar. The cathode shunt at the socket (Fig 4), with 1 A of plate current, will produce 0.33 V at point S– in Fig 10.[23] A gain of 10/0.33 = 30 will yield the 10 V we want for the meter with its multiplier. With $R_i$ = 10 kΩ again, we'll need $R_f$ = 290 kΩ. "Shunt" is the traditional term, as if we were to connect a meter directly, but here the shunt merely converts current to voltage, which the op amp amplifies, and the meter multiplier then converts back to current. The 1-µF capacitor provides some meter damping, giving a less wild indication on SSB.

Screen-voltage metering is also similar to the above. So are forward- and reverse-power metering, although the amplification needed for these may depend on the particular construction of the directional coupler. The forward-power meter is peak-indicating. Although very simple, it is extremely useful in conjunction with pulse tuning (see pp 11-12 in "A Luxury Linear" and the "Operation" section below). The diode enables fast attack with a positive signal from the forward power coupler, while the 1-MΩ/1-µF network provides the hold function. The time constant is chosen so that the meter holds long enough for full-power indications under pulse tuning and CW operation, but reacts fast enough to follow tuning adjustments.

### Zero-Center Screen Metering

Perhaps more interesting are the zero-center and expanded-scale metering circuits. Zero-center screen-current metering is essential with the 4CX1000A. I could find no zero-center meter on the surplus market to match the others on hand. Finding five similar-looking meters on the surplus market was the hardest part of the whole project. The alert reader will have already noticed that there are actually three different Triplett types (with identical cases) in the cover photo for "A Luxury Linear".

The inverting-summing amplifier with virtual ground is one of the most useful of the op-amp-arithmetic connections; we use it for zero-center screen current metering. The secret is the virtual ground. The noninverting input (+) is really ground, while the op amp with inverse feedback tends to keep the inverting input (–) at the same voltage level as the noninverting, namely zero! So the inverting input is always at 0 V; it acts like a ground. The main advantage of this circuit, compared to the noninverting summing amplifier, is that there is no interaction between the several inputs.

We want a 50-0-50 mA screen meter. The 100-Ω screen-current shunt in Fig 5 will produce –5 V at the –E2 input for 50-mA of positive screen current. Thus we need a gain of –1 to obtain the desired +5 V for half-scale deflection in this inverting circuit. Again with $R_i$ = 10 kΩ, $R_f$ will have a nominal value of 10 kΩ; we use a 5-kΩ trimpot and an 8.2-kΩ fixed resistor.

Now we center the meter. We need only fool the circuit into thinking there is 50 mA flowing. We want an op-amp output of +5 V with no screen current. Using the –12-V rail as an input for this purpose, we need a gain on this branch of +5/–12 = –0.417. With $R_f$ already chosen as nominally 10 kΩ, we calculate $R_i$ = 24 kΩ in this branch; the components shown give a range of 18 kΩ to 28 kΩ. The zero setting hasn't moved a hairline in the two years since the initial adjustment.

### Expanded-Scale Heater Metering

The differential amplifier is the circuit of choice for measuring a voltage between two points, neither of which is at ground. Alternatively, the virtual ground in the summing amplifier part of this circuit could have been made virtual S–. But while that would have worked for measuring the voltage, the expanded-scale part of the circuit uses the +12 rail, which is ground-referenced. (The regulation of these rails is essential to the stability of these circuits.) That means another regulator would be needed, referenced to S–. So the differential amplifier is really the simplest solution. The circuit used here in Fig 10 is particularly simple. The heater voltage sensing is at terminals S– and S+; let me call the voltages at these terminals $E_1$ and $E_2$, respectively. With all four resistors equal (the value doesn't matter), the output of the differential amplifier is $E_1–E_2$. The differential amplifier therefore simply measures the heater voltage and converts it to a ground-referenced voltage.

The summing amplifier is used to obtain the expanded-scale heater metering, 5.0 to 6.0 V. We want a 1-V change in heater voltage to produce a 10-V change at the op-amp output, for full-scale indication. Thus a gain of –10 will be needed. The differential amplifier has been arranged to invert the heater voltage once, so once more and we're on our feet again. With $R_i$ = 10 kΩ in the measuring branch, we want $R_f$ = 100 kΩ. For an expanded scale indication, we use the expanded scale branch to give the circuit a tendency to indicate –5 V (of course the left meter pin gets in the way); then it will take +5-V input at the measuring branch to indicate a composite zero. The meter has a 1-V full-scale range, so this means a hypothetical –50 V at the op-amp output. Using the +12-V rail for this, we want a branch gain of –50/+12 = –4.17, and we need $R_i$ = 24 kΩ. The individual calculated outputs due to the two input branches are superimposed—they add. Thus the output is $10(E_2–E_1)–50$. The output is zero when the heater voltage $E_2–E_1$ is 5.0, and the output is 10 V when the heater voltage is 6.0.

Adjustment of these circuits is easier than it might seem at first. All adjustments are made with the amp on the bench using small test power supplies to simulate the parameters. Not with high voltage while transmitting! For most tests and adjustments the tube is even cold; a switch turns off the blower, so there is no noise on the bench and the heater is off. In each circuit, first adjust the CAL trimpots (for correct deflection, no matter what part of the scale), and then the CENTER or LEFT trimpots. (I should have put the CAL trimpots at the input—then there would have been no interaction between the CAL and ZERO or LEFT adjustments.)

All the trimpots are at the front edge of the control board. For touch-up adjustments, the amplifier may be slid a few inches out of the rack, leaving all cables attached. The trimpots are then accessible through the top cover vent holes.

### Construction

Blower mounting was one of the toughest jobs. The low-noise requirement demands mounting on rubber. The suggestions here are not exactly the way I installed the blower—they are hopefully easier. A good source for material is your local auto parts supplier. Vacuum hose is very thick-walled, with a small ID. Slicing lengthwise through one wall yields a heavy piece of rubber that can be slipped over the edge of the blower outlet. Fig 3 shows that I supported the blower by three screws, with rubber grommets, to the rear panel, but my suggestion is to fabricate a bracket inside the amp and support it from above. The weight of the blower, and a bit of pressure from the bracket, will keep the rubber gasket snug and air-tight.

Inside the grid compartment, across the air inlet, is an RF shield made of copper screening. The air-flow switch is mounted inside the grid compartment; the steel actuating wire on the switch passes through a small hole in the copper screen to the air vane, which is fully inside the blower outlet tube. Alternatively, and more easily, the air flow switch may be mounted on the side of the outlet tube.

After a few clumsy attempts, a very simple method for assembling the input circuit T-match (with insulated above-ground rotors) was found. The key idea is that the two rotors connect together and to the coil (Fig 4 shows the rotor of C2 incorrectly). The two capacitors mount on a piece of one-sided copper circuit board (Fig 2), which is mounted on ceramic insulators. This results in a very low-inductance connection between the two capacitor rotors. The coil connects between the circuit board and a ground lug. The lug is copper; use a small magnet to reject the steel solder lugs in your junk box. The silver-mica blocking capacitor, seemingly redundant, is to protect the tube from loss of bias in the event of a short in C2. The shafts in both grid and plate boxes are of Delrin rod from Small Parts.[16] There might be a better material; there was a recent discussion on the rec.radio.homebrew Usenet Newsgroup about the best materials in RF environments, but I don't remember any conclusive consensus. I do know that there is a glob of melted nylon sitting on my desk as a reminder of the first attempt to connect to the input tuning capacitor shafts.

To wind the 1/4-inch tubing for the plate tank without kinking (I made a new coil after the photo!), seal one end in a vise, fill with sand, seal the other end, wind, cut off the ends, and then return the stolen sand to your kid's sandbox.

Construction of the control board entails a certain dilemma. Many projects are essentially experimental; the builder wants to try a few new methods. If one goes through all the trouble of etching a board based on the first draft of the circuit, the many subsequent modifications will shortly turn it into a nightmare. If a different construction method is used, then, when all the final changes have been made, the amplifier is finished and there is no longer a need for an etched board.

For the past seven years I've used wire-wrapping for control boards in all my gear. There are quite a few advantages. Quite dense packing is possible, much denser than ordinary perf-board construction. I even install all the resistors and other small parts in sockets, along with the ICs. This makes it a trivial matter to change a component value, facilitating circuit development. Although it does add a bit to the cost, it can save hours and hours of time. Making a wiring change cannot be said to be easy, working in the maze of wires under the board, but it is possible, clean and neat. It does require some concentration, a small price to pay for having no unsoldering to do. The board is mounted on hinges for easy access to the wire side. I've never seen a wire-wrap connection fail.

All the wire-wrap supplies are available from Digi-Key.[20] Every socket and pin is numbered using a simple matrix scheme, and the numbers are noted on the schematics in the notebooks. Point 237 is pin 7 of U23, the third IC in the second row. Resistor R456 is plugged into the fifth socket in the fourth row, with one lead at pin 6 (by convention, this will be the left or upper end in the schematic). Trimpots can also be fitted onto sockets, if the sockets are of the machined-pin type. Depending on the trimpot type, a touch of solder at each lead may be warranted. The dip relays naturally plug into sockets. A few larger components, such as the coax relay driver and its 5-W base resistor, are soldered to separate wire-wrap pins at the edge of the board (at the bottom in Fig 1). Each (stranded wire) lead to the board is filtered as noted in the caption to Fig 6; the rows of bypass capacitors and RF chokes consume a surprisingly large portion of the board. Each lead to the board passes through two holes before soldering to its wire-wrap pin. This is for strain-relief, so a wire cannot be bent at the point where it is soldered. At certain parts of the amplifier, tie-downs are used for this purpose; Hosfelt has handy stick-on types.[19] DX chasing and contest work require the utmost level of reliability; one broken wire can spoil your whole run.

### Operation

Providing suitable drive power is important. Problems can arise when a solid-state driver is used. These "bricks" often exhibit greater IMD (splatter) when operated at reduced output. This will be the case no matter how linear the final amplifier is; it amplifies whatever you feed it. It's just like the old computer cliché, "garbage in, garbage out." It sounds contradictory—a driver operating with less power should produce less splatter. But IMD is relative, and we are to amplify whatever comes out of the driver. One solution would be an attenuator, perhaps built around a dummy load, so the brick may be run at a more linear level. A better solution would be adjusting the bias in the driver. The bricks are usually rated for 100% duty cycle on FM, a steady carrier. SSB and CW operation is much gentler, so the driver idling power may be considerably increased, which should improve IMD performance.

My driver is capable of 200-W output; it's a homebrew conduction-cooled tetrode amplifier with neutralization and 26 dB of gain. As is typical for a class $AB_1$ tetrode, it is linear at any output level. The transverter maximum output is 2 W, running class A at the output transistor. About 14 dB of gain reduction is needed. This is done in the transverter at the milliwatt level by a resistive panel control in the RF path. The common practice of using high ALC levels to reduce gain often produces IMD (splatter).

### Heater Idle Circuit

There are several modes of station operation that call for very long periods—hours and hours—when an amplifier must be ready for near-instant operation, but is rarely used. In my case, with this 2-m amp, the main situation is VHF contest operation (in the sparsely settled Southwest). The contest starts at noon Saturday—hopefully with a big blast on a wide-open 6-m band. There is not much doing on 2 m until around sunset when tropo improves and operators in neighboring states start swinging antennas in all directions. But all afternoon we must be ready on 2; there may be sporadic-E at any moment. And then the same all day Sunday, with the 2-m tubes mostly just sitting and cooking away. A similar situation arises on any summer day when 6 m is open; if the skip is shortening we want to be ready on 2. A more common situation, which would apply to an HF amplifier, involves the DXer who spends almost all the time just listening (as the very best operators do), but must be ready for that new one.

The answer to these problems is the heater idle circuit. This is a simple timer that drops the heater voltage to a specified low level after no transmissions have been made for a specified time. I chose a 12% drop after 5 minutes. Because the heater voltage is regulated, this was easy to do with an IC timer, a DIP relay and a trimpot. In other amplifiers, an IC timer, a relay and a resistor in the filament transformer primary circuit would serve the same purpose.

The idle circuit schematic is shown in Fig 11. If there is no transmission for 5 minutes, the 555 timer switches the 723 regulator in Fig 5 from the panel pot, which sets operating heater voltage, to the trimpot inside that sets the idle voltage. The circuit resets to operating voltage automatically whenever the PTT line (mike button, foot switch or semi-QSK circuit) is keyed, or whenever the reset button on the amplifier front panel is pushed.

How long does it take for the heater to reach operating temperature after reset? I made a number of tests, aiming to determine whether I could forget about the idle circuit, and just grab the mike or key whenever I heard a new grid square, leaving the automatic recovery feature to restore operating heater voltage. The idea (unsubstantiated by any manufacturer or authority) is that the heater loses heat mainly through cathode current. Under this hypothesis, during receive periods the heater gets too hot, hotter than in continuous transmit operation. Thus, it should be possible to maintain operating temperature with less voltage when not transmitting, and full heater voltage would be needed only at the instant a transmission begins.

Does it work? Well, without asking the factory to saw open the tube to investigate, I can only conclude that since the tube has survived all the testing, it works fine. Still, I usually don't wait for the automatic reset. If I hear someone I want to call, I touch the foot switch for half a second to reset the heater voltage. But often I forget to reset manually and the automatic reset circuit works fine.

The tests involved going from the idle condition instantly to full-peak power (pulsed), or full-power CW (dits), to see if full power was instantly achieved. Yes, it was, as quickly as I could read the meter. This power-output test for adequate heater temperature is consistent with tube manufacturers' suggestions for heater voltage adjustment in VHF operation: reduce the heater voltage gradually until the power output drops slightly, then bring it back up a bit. In other words, if you're getting full output, the heater is hot enough. Any more heater voltage and you're just cooking the life out of the poor tube.

Using the heater voltage and current meters, the heater resistance may be calculated. From this the heater temperature may be inferred, relatively, although I don't know the exact relationship. The 4CX1000A is rated nominally at 6.0 V, with advice for lower voltage if adequate, especially on VHF. I operate at 5.8 V (where it draws 9.2 A) and idle at 5.1 V (where it draws 8.4 A). This is a 20% drop in heater power during idle periods, enough to expect a vastly increased tube life. The calculated heater resistance drops from 630 to 607 mΩ.

Since the heater has lower resistance during idle mode, the current will be slightly higher than normal immediately after reset. The time taken for the heater current to drop to normal after reset is another test of the idle-circuit idea. Although difficult to clock, because it happens so quickly, it takes less than 2 seconds.

Note that the heater drops to idle only after 5 minutes without a transmission, not between each transmission during normal operation. And, if you are still worried, you can always reset manually with the mike button, the foot switch or the reset button a few seconds before resuming operation.

### ALC for Transverters

For fighting splatter on the ham bands, ALC is the most powerful weapon. The ALC circuit in Fig 8 will produce a negative voltage for driving the ALC-controlled stages in a transceiver or transverter. The R-C network at the follower input prevents keying transients from generating ALC voltage. The pot on the front panel sets the ALC threshold at 0.1 mA of grid current. The panel adjustment may be used for experimentation. The 4CX1000A is rated for zero control-grid dissipation; that is, zero current. However, the spec sheet says a few milliamperes on peaks is okay. I wanted to see if on CW, where IMD is not a consideration, a few milliamps would result in higher efficiency. (I used a 5-mA grid meter initially.) The results: no, there is no improvement in efficiency with higher grid current.

Some rigs (eg, the FT-1000MP) do not have an ALC input line which controls the RF level at the transverter output jack. ALC may be added to a transverter with the circuit shown in Fig 12. The MOSFET buffer stage is designed for unity gain, but this may depend on the individual device. Increasing the drain resistor will increase the output, if necessary. The RF input level should be not more than –16 dBm. The gate 2 bias adjustment is set so that Q1 operates at the knee of its characteristic curve.[24] If the bias is set for maximum gain, the ALC voltage will be forced to rise to a high value before any significant gain reduction is obtained. My transverters have two or three ALC-controlled stages; this means that there is less gain reduction at each stage, and therefore less possibility of ALC-induced IMD. The ALC adjustment in Fig 12 is set so that –5 V applied at either ALC input will produce 20 dB of ALC compression. Excessive ALC may result in IMD. In operation, the transverter gain is adjusted for a maximum of 3 dB of ALC compression, metered on the transceiver panel.

### Pulsed Tune-Up

The dit tune-up procedure, with amplifier keying, was described in "A Luxury Linear." In fact, the method used in my shack goes even further. My home-brew transceiver has a built-in pulser, set for a 33% duty cycle. The TUNE switch on the panel automatically shifts the radio into CW, silences the sidetone, hits the PTT line and starts the pulser. So testing at 1500-W PEP output involves only 400 W of average plate dissipation.

The schematic for the pulser, which could be built into a transceiver or as a separate device, is shown in Fig 13. A scope with calibrated time-base is useful for precisely setting the desired pulse rate and duty cycle, but a simple RF monitor scope is sufficient. Merely setting the OFF trimpot at maximum and the ON trimpot at midrange will be adequate.

### Improvements

The main improvement I would suggest is in the swinging link mechanism. The simple shaft and insulator method described in the caption to Fig 4 works well enough. However, the link needs to swing only about 30°, and the heavy coax-braid connections make it a bit stiff. In operation, I use the swinging link only as a coarse loading adjustment, so there is no practical problem. Since I haven't touched the swinging link in the past six months, this is one of those situations where improvement is not worth the trouble. But for someone trying this circuit anew, I would suggest some sort of gear drive, for smoother control and finer adjustment.

I have only one other suggestion: build very carefully! Don't lay yourself open to a remark like I got from one of the locals here. All he could say when he saw the new amplifier was, "You got one of the meter labels crooked."

#### Notes

[21] "A Luxury Linear," *QEX*, May, 1996, p 3-12. (Photos also in *QST*, July 1996, p 19.)

[22] References to Figs 1-10 and Notes 1-20 are to the previous article (see Note 21 above). This article begins with Fig 11 and Note 21.

[23] Correction for "A Luxury Linear," Fig 10: Reverse the ± markings at the inputs of the plate current-metering op amp.

[24] A typical MOSFET characteristic curve is shown as Fig 10 in "A High-Performance AGC System for Home-Brew Transceivers," *QEX*, October 1995, p 12-22.

# RF

By Zack Lau, W1VT

### Mounting RF Connectors

I spend a lot of time mounting RF connectors—so much so that I've developed standard procedures for the popular ones I deal with. You might find these guidelines useful in designing or assembling your projects.

The first decision to make is the choice of connectors. Typically I use SMA, BNC, N and UHF connectors. Since I do a lot of microwave work, SMA connectors are my first choice for many applications. Not only do they work just fine on all the amateur bands through 10 GHz, but I like their small size. For hand assembly in the field, N connectors are a much better choice. BNC connectors are great for HF/VHF work, as well as lab experiments where the bayonet quick-disconnect feature comes in handy. Female BNC connectors will often mate with N males—a nice feature if you are looking for a push-on/pull-off connector pair, however, UHF connectors are still the best choice if you want your gear to be compatible with commercial HF equipment.

Fig 1 shows how I mount typical 4-flange SMA connectors. I tap four 2-56 holes on 0.340 inch centers. The center hole is drilled with a #19 drill. The two flange connectors merely have two of the holes lopped off, so the center-to-center spacing is typically 0.481 inch (481 mils). SMA connectors have $^1/_4$-36 threads on their mating surfaces. The 4-hole flanges are normally 0.50-inch squares, though you can find versions that are 0.70-inch squares that are interchangeable with BNC connectors. Sometimes they measure a little over 0.500 inches, which is important to keep in mind if you are working in a tight space.

I often solder the center pins of SMA connectors directly to 50-Ω microstrip traces. Lining up the pins can be a problem—here is my technique. First I trim the etched printed-circuit board to the desired size. Next, I measure the distance of the center pin to the edge of the board with a dial caliper. I then use the caliper to mark the enclosure material—usually brass strips 25 mils thick. The dial caliper works better than the new digital ones—it's pretty easy to dial in the 170-mil offsets for the screw holes. After drilling the #50 holes and tapping the 2-56 threads, I temporarily

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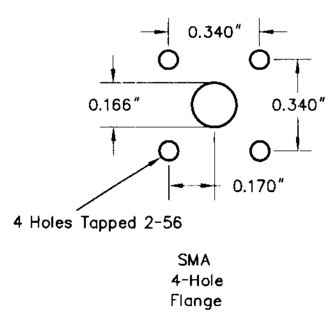


Fig 1—Template for drilling panels for SMA jacks.

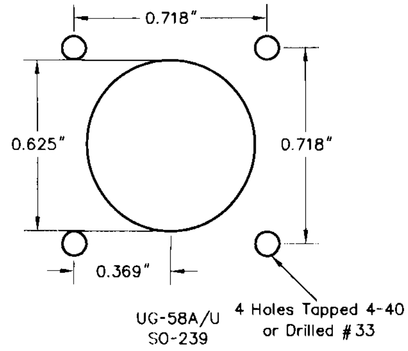


Fig 3—Template for drilling panels for N or UHF jacks.

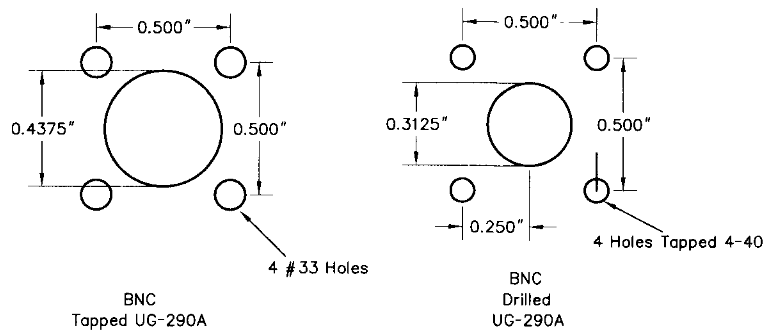


Fig 2—Template for drilling panels for BNC jacks.

mount the connectors to their respective panels and solder the center pins to the board. I then tack solder the board and brass strips together. Then I remove the connectors to solder everything together. Otherwise, the connector often gets in the way of the soldering iron. A 40-W iron works well for tacking stuff together, while a bigger 100-W iron usually does a better job for the final soldering, particularly with big enclosures. After making a lid out of aluminum or brass sheet, I mark and drill the cover mounting holes. Drilling the holes before installing the parts is a good idea. This makes it easier to remove the metal shavings and reduces the problem of destroying parts with the drill.

Fig 2 shows the drilling diagram for BNC connectors. (Full-scale images of these templates are available in *Microsoft Word* and Postscript format at **http://www.arrl.org/qexfiles/conntemp.zip**.) The 4-hole flange connectors are tapped for exotic 3-56 screws on 0.500-inch centers. I either drill the holes out with a #33 bit to pass 4-40 screws or tap the holes for a 4-40 thread. To pass the connector through the panel, I use a 7/16-inch drill bit. This is actually 16 mils smaller than specified in the Amphenol catalog (0.453 inches), but I've never had a problem getting the connector to fit. Otherwise, a 5/16-inch hole is big enough for the center pin. I find that I can usually just rethread the holes—it isn't necessary to enlarge the holes for the tap. The threaded BNC connectors have 3/8-32 threads. The UG-290 connectors have 0.687-inch-square flanges according to the Amphenol catalog, but I normally use a flange width of 0.75 inches when laying out a panel.

Fig 3 shows the drilling diagram for UHF and N connectors. The fact that they are the same is handy—you can often substitute one panel jack for another. I tap four 4-40 holes on 0.718-inch centers. Alternately, you could drill four #33 holes and use 4-40 screws with lockwashers and nuts. Both UHF and N connectors have 5/8-24 threads on the mating surfaces. The center hole is 5/8 inch. Again, the drawing in the Amphenol catalog shows a hole 16 mils wider, or 0.641 inches in diameter. I use a flange width of 1.0 inches when laying out a panel.

A disadvantage of using tapped holes is the necessity of accurate holes. You don't get to ream a tapped hole! I prefer to use a dial caliper for marking. If you don't mind buying something made in a third-world country, these can be remarkably inexpensive—under $20. They are cheap enough that I use the jaw tips as scribes. I've seen people avoid the problem of tapped holes by soldering gold-plated connectors directly to the chassis. But unless a big soldering iron is used to properly solder the connectors, the result can be rather ugly.

I've found that a Uni-bit can be quite useful for mounting connectors on thin sheet metal, such as that used to make tin cans. The Uni-bit will drill nice clean holes, where a normal bit might tear an ugly gash into the sheet metal. While I've soldered connectors directly to the sheet metal, it makes more sense to solder a brass mounting plate to the tin can. This makes it easy to remove the connector for microwave tasks like adjusting probe lengths. I've also found shim stock useful for adjusting probe lengths—just add a bit of shim stock to effectively shorten the probe. An assortment of connectors with different probe lengths and a pile of shim stock might be a wise investment if you do lots of microwave experimentation.