QEX

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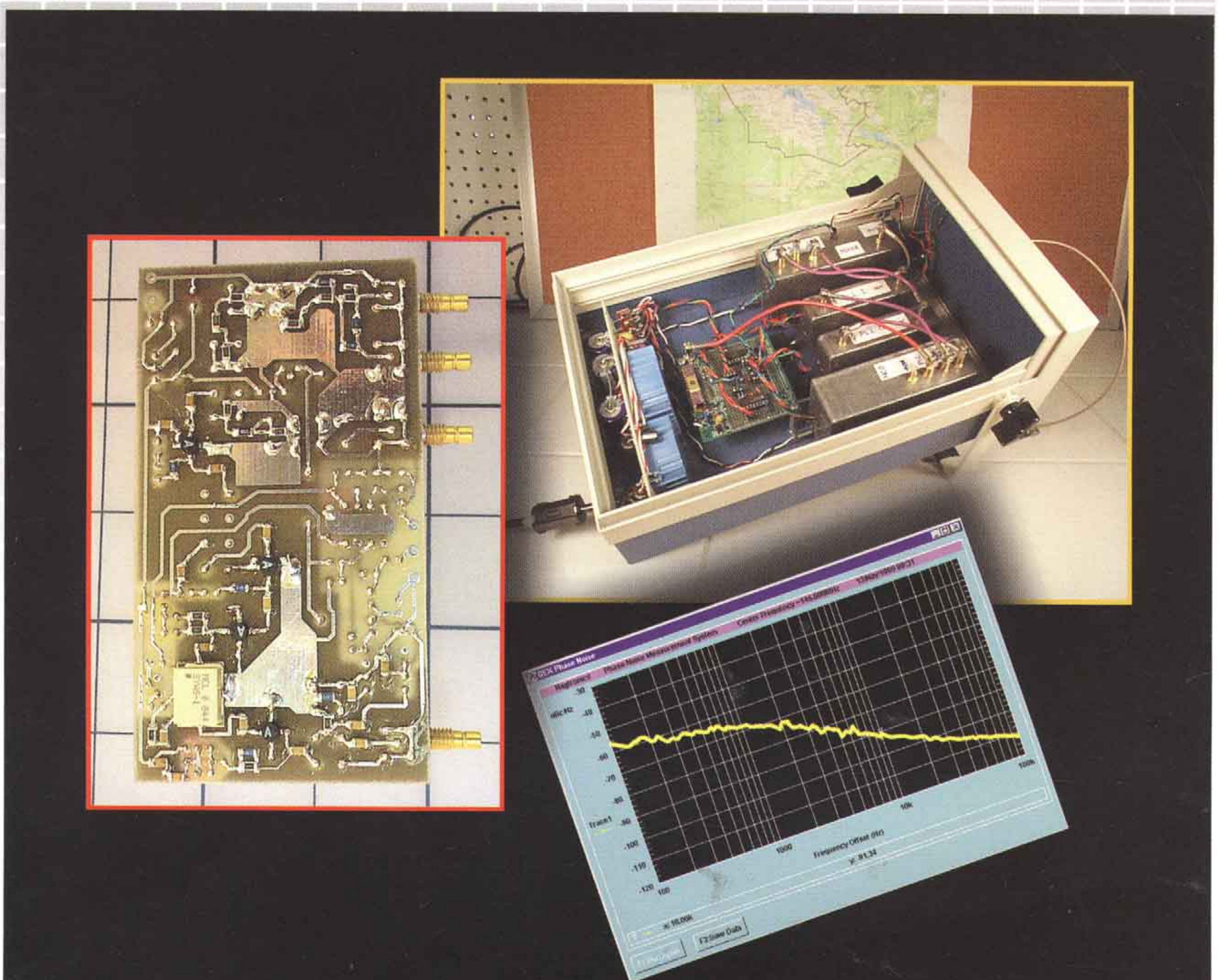
January/February 2000

Forum for Communications Experimenters



Build this 250-MHz Synthesized Signal Source

ARRL *The national association for AMATEUR RADIO*

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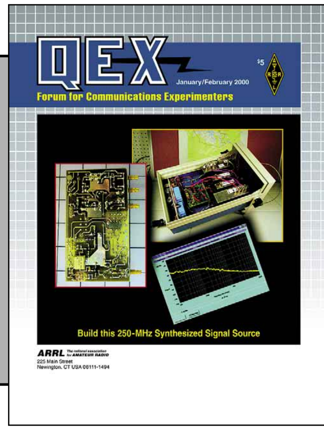

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THE AMERICAN RADIO RELAY LEAGUE

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1) provide a medium for the exchange of ideas and information among Amateur Radio experimenters,

2) document advanced technical work in the Amateur Radio field, and

3) support efforts to advance the state of the Amateur Radio art.

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Both theoretical and practical technical articles are welcomed. Manuscripts should be submitted on IBM or Mac format 3.5-inch diskette in word-processor format, if possible. We can redraw any figures as long as their content is clear. Photos should be glossy, color or black-and-white prints of at least the size they are to appear in *QEX*. Further information for authors can be found on the Web at **www.arrl.org/qex/** or by e-mail to **qex@arrl.org**.

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Empirically Speaking

By the end of this year, the press will have raised more hullabaloo over the new millennium than over all other topics combined. For experimenters, 2000 marks the passage of Amateur Radio into a period of unprecedented challenge and change.

Soon, we will celebrate with our 200th issue! Since I have a good excuse, I'm actively seeking updates on older articles. You authors may like this chance to give us a few paragraphs on what's changed in your designs, in your thinking or in your lives. I know work continues on many of your projects after publication. Too often, we don't get to see the subsequent refinements.

You may have ideas about how to improve circuits or software that you saw in *QEX*. Perhaps you don't think your idea justifies an entire article or letter to the editor. Maybe you have an update on the availability of a product, an overdue correction or just a note to say "Gee, look how things have changed!" I think we should make space for this kind of thing, especially because it gives us a chance to look back at where we've been.

Now is the time to ask "Where else are we going with *QEX*?" Well, we've seen numerous comments that you like what you're seeing thus far. Your thoughtful letters have sustained some interesting discussion. The flow of articles is up and we are covering many important themes in communications. We want to broaden our perspective by including material representing a growing range of interests and skill levels.

Technology's accelerating pace, however, forces experimenters to spend more time researching possibilities for designs, which leaves less time to write about what we actually built. Scarcity of good published information, in turn, makes it tougher for the next person to do his or her research. Major update cycles for reference works, such as *The ARRL Handbook*, are rapidly getting shorter.

Since *QEX* articles are usually published well within a year of their submission, we fill gaps by providing timely exposure for projects, some of which may migrate to the *Handbook*. We look forward to helping document some exciting new applications as we head into a new era.

In This *QEX*

Our examination of frequency synthesis techniques continues with Steve Hageman's 250-MHz generator. The availability of high-performance circuit building blocks has put this type of project within reach of amateurs. Steve employs computer control for flexibility and simplicity. Purity and stability issues are addressed in the design.

Guy Fletcher, VK2KU, has performed some fastidious measurements on VHF, UHF and microwave Yagis with some interesting results. He shows how the data support his formulation of "Guy's Rule" for the relation between boom and element diameters and element length. The model removes some of the surprises and much of the time involved in accurately designing antennas in those frequency ranges. I suspect the theory behind it also comes into play more often than not for other types of antenna structures. Thanks to our friends at *WIA* for their assistance.

Thomas Duncan, KG4CUY, shows how to build frequency-control functions around a VCO—almost any VCO. He illustrates important PLL and control-system features.

Stu Downs, WA6PDP, went to great lengths to eliminate *all* his automotive RFI on the HF bands. Armed with a laudable understanding of conducted and radiated emissions, he set about isolating and curing his problems, one by one, until his receiver was quiet. His revelations should help frustrated hams and automobile manufacturers alike.

As opposed to RF, digital circuits tend to behave in binary fashion: They either work or they don't! Troubleshooting them can involve observing a myriad of high-speed signals. Larry Cicchinelli, K3PTO, brings us a tool of significance in an increasingly digital world: a PC-driven logic analyzer. Again, the personal computer proves itself a matchless control system.

We get down to baseband with Mark Mandelkern, K5AM's home-brew transceiver. Mark shows special interest in noise limiting in this penultimate segment of his series. In *RF*, Zack Lau, W1VT, explores the use of ladder line at UHF—73, *Doug Smith, KF6DX,* ***kf6dx@arrl.org***

Build this 250-MHz Synthesized Signal Source

You can build this high-quality signal generator. PC control and prefab parts make it relatively easy to build and operate. An off-the-shelf doubler extends its range (fourth harmonic) to –16 dBm at 1 GHz!

By Steve Hageman

I find it extremely useful to have a general-purpose RF source available when I want to quickly breadboard a receiver. With a stable source, I can first concentrate on a new receiver's RF and IF portions and use the source as a temporary LO. Secondly, when the receiver's LO is finished, the source can be used as the RF input to verify performance. Finally, as a general-purpose troubleshooting tool, it sure is nice to have any frequency and amplitude you want right at your fingertips, when you want it.

The "wireless revolution" comes to our aid in this project. All the circuitry used in this project is the result of manufacturers' pushing the state of the art to provide highly integrated solutions for the wireless industry. It's a great time to be interested in radio and homebrewing!

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Windsor, CA 95492
shageman@sonic.net

How the Source Works

The CW signal generator, as I have designed it, (see Fig 1) is one of a class of synthesizers employing "coherent direct synthesis."[1] *Coherent* means that all the frequencies are derived from one master oscillator—the 40-MHz clock on the Numerically Controlled Oscillator (NCO) board in this design. The synthesizer is direct because it uses frequencies generated by PLL circuits in a single mixer to generate the output.

Coherent synthesis is advantageous because it tends to eliminate frequency-drift problems. In fact, if the reference oscillator in this synthesizer drifts, the output frequency drifts in exactly the same direction and at the same rate. When multiple frequency references are used, this may not be the case.[2]

The synthesized source described here uses many excellent ideas presented previously.[3, 4, 5] These are combined to make a 2-250 MHz, programmable VHF source.

In Rhode's 1994 article (see Reference 2) he described using an NCO as a "micro-adjustable" frequency reference to a PLL circuit. This topology has the advantage of allowing the PLL to operate at a high reference frequency, thus allowing a high loop bandwidth (BW), while having sub-hertz channel spacings.

[1]Notes appear on page 14.

While Reference 1 describes many other methods of achieving small channel spacing, I have adopted this approach as the simplest way of achieving the disparate design goals of a high reference frequency and small step size.

Looking at the block diagram shown in Fig 1, the NCO board (also Fig 2) contains the 40-MHz master oscillator, the NCO circuit and a divide-by-four circuit that generates a separate 10-MHz reference. The 32-bit NCO, a Harris HSP45102 device, yields a step size of:

$$df = \frac{40 \bullet 10^6}{2^{32}} = 0.009 \text{ Hz} \qquad \text{(Eq 1)}$$

The output of the NCO is set for 10.7 MHz. As per Rhode's assessment of the design problem, using a common IF allows the use of off-the-shelf, low-cost filters. I chose a ceramic filter because it provides more than adequate sideband suppression and its impedance is such that I could match it with simple resistors while keeping plenty of signal level to drive the PLL's reference input.

I also considered 455 kHz. This would have had the advantage of lower distortion from the NCO, but the available ceramic filters need to be matched in the 1 to 3-kΩ range. This would have complicated the matching circuitry to keep the losses at an acceptable level.

The filtered NCO output drives the PLL 0 board (Fig 3). This PLL design uses the readily available, highly integrated Motorola MC145191. The MC145191 is a low-power, 1-GHz PLL that contains an internal reference divider, N + A feedback divider and charge-pump phase detector. The reference, or "R" divider is used to set the PLL's channel spacing. Here, the 10.7 MHz of the NCO is divided by 107 to get a reference frequency of 100 kHz. When the value of the N + A counter is changed by one, the output of the VCO steps by 100 kHz. So the exact NCO frequency is used to interpolate between the PLL's basic 100-kHz channel spacing.

Mathematically, this works out to the total "frequency gain" of the circuit (750 MHz / 10.7 MHz = 70). For a step size of 1 Hz, the NCO must have a frequency resolution of no greater than 1 Hz / 70 = 0.014 Hz. As shown above, this requirement is easily met by the NCO circuit used.

The output of PLL 0 is an adjustable 500 to 748-MHz signal that is used by the mixer board to synthesize the source's 2 to 250-MHz output. PLL 1 is similar to PLL 0. In fact, it uses the same PC board and parts. PLL 1 produces a fixed frequency of 750 MHz, however. The input to PLL 1 is fixed at 10 MHz (divided from the 40-MHz crystal oscillator) and the R divider is set to 100 to yield a 100-kHz reference frequency to the phase detector (same as PLL 0).

The PLL outputs are used as inputs to the Mixer board. The Mixer board looks much like a standard down-converting receiver because that's essentially what it is.

The 750-MHz PLL 1 signal is used as fixed-frequency local oscillator (LO) drive to a mixer, while the 500 to 748-MHz PLL 0 signal is used as the RF input. Both PLLs are low-pass filtered to remove any harmonics. As we will see later, this greatly improves the spurious performance of the synthesizer overall.

The filtered PLL signals are routed to a modular doubly balanced mixer (Fig 4). The desired output from the mixer is the difference frequency of 2 to 250 MHz. To ensure that the mixer operates with low intermodulation distortion, its IF output is broadband-terminated at 50 Ω with a Hewlett-Packard MSA-1105, 1.3-GHz MMIC amplifier. The MSA-1105 provides a reasonable input match up to 3 GHz and provides about 12 dB of gain. The mixer's sum-frequency output is filtered by a 270-MHz low-pass filter at the output of U4 as shown in Fig 4.

After several more stages of amplification, the RF signal is fed through a PIN-diode attenuator, final amplifier and directional coupler. The directional coupler is used to separate the forward output power from any that is reflected, so that the true forward power is leveled correctly. This standard scheme has been used for decades in leveling the power of RF sources (see Reference 4).

The forward-coupled path is detected in the leveling loop by an Analog Devices AD8307 broadband log amplifier to form a dc voltage that is proportional to the output power in dBm. This dc voltage is compared to the reference voltage produced by a digital-to-analog converter (DAC) on the CPU board and the PIN-diode attenuator is adjusted by the leveling loop until the equilibrium is achieved. By changing the leveling-DAC output voltage, the output level of the synthesizer may be changed over the range of –20 to +15 dBm. A directional coupler is used because it separates the true output power of the synthesizer from any reflections, thus increasing the accuracy of the output level in the face of possible wildly varying termination mismatches (see Reference 4).

The CPU board (Fig 5) contains a Microchip Technology PIC16C63, single-chip microprocessor. The PIC is used as a smart UART; that is, it receives command packets from a PC over a 19.2-kbps RS-232 port and executes instructions to set the hardware or shift serial data out to the PLLs, NCO and leveling DAC. The PIC operates as a "smart-logic" device that can be quickly reprogrammed during development. It essentially replaces dozens of discrete logic chips with just one package. The CPU also monitors the lock status of the PLLs and function of the leveling loop, then sends this information back to the PC.

As shown in Fig 5, the RS-232 interface includes a "link port." This connection allows up to four sources to share one RS-232 port. The address-select switch must be set to the desired address for each source on the link port. When a command is sent to a source, only the source with the proper address acts on that command.

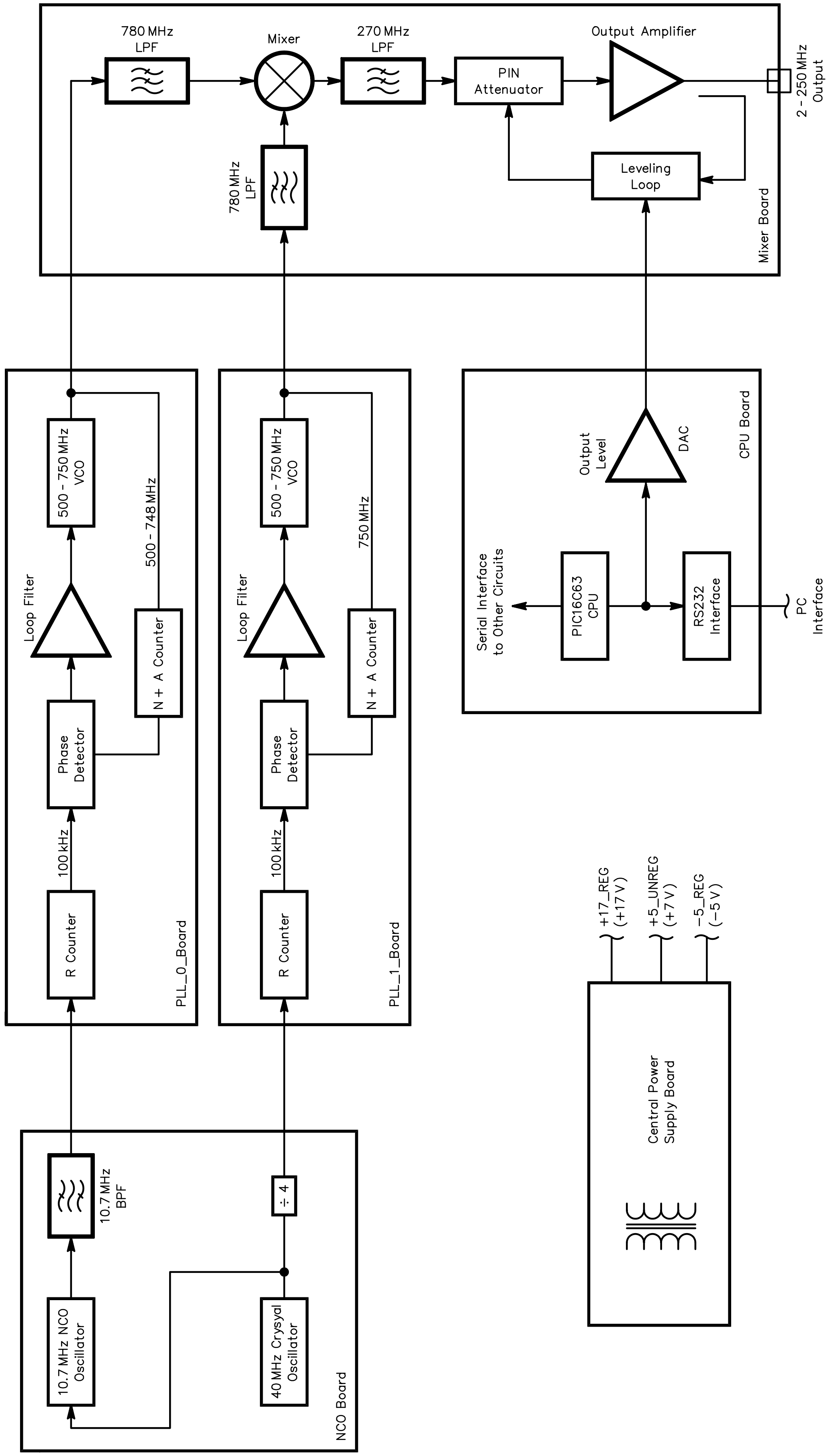
The PIC does no significant calculations; it merely acts as go-between for the PC and the actual hardware. As described later, the PC program does all the calculations and sends binary information to the PIC via the RS-232 serial port. The PIC converts this binary information to a bit stream that sets the PLLs and other components to desired states.

Here's the bottom line of this whole scheme: By using a fixed frequency of 750 MHz for the LO and mixing this with a variable frequency of 500 to 748 MHz, we can produce a 2 to 250 MHz output. Conventional VCOs cannot be made to tune much more than an octave. By using a heterodyne technique, we can make a synthesizer that tunes over seven octaves in a single band.

Some Design Specifics

One thing I've found while working with PLLs is that it is very easy to get the frequency you want. The tough part is not getting the frequencies you don't want! In fact, until one learns to

Fig 1 (see right)—As this block diagram of the synthesized source shows, the design is partitioned into five functional blocks (the PLLs share the same circuit design). Each of the RF blocks is housed in a die-cast metal enclosure to suppress cross-talk between modules and contain any RF radiation.


780 MHz
LPF
Mixer
270 MHz
LPF
PIN
Attenuator
Output Amplifier
2 - 250 MHz
Output
780 MHz
LPF
Leveling
Loop
Mixer Board
500 - 750 MHz
VCO
500 - 748 MHz
Loop Filter
N + A Counter
Phase
Detector
100 kHz
R Counter
PLL_0_Board
500 - 750 MHz
VCO
750 MHz
Loop Filter
N + A Counter
Phase
Detector
100 kHz
R Counter
PLL_1_Board
Output
Level
DAC
CPU Board
Serial Interface
to Other Circuits
PIC16C63
CPU
RS232
Interface
PC
Interface
+17_REG
(+17 V)
+5_UNREG
(+7 V)
−5_REG
(−5 V)
Central Power
Supply Board
10.7 MHz
BPF
÷ 4
10.7 MHz NCO
Oscillator
40 MHz Crysyal
Oscillator
NCO Board

measure phase noise and discrete spurious frequencies, life goes on quite merrily.

The frequencies you don't want are discrete spurs—such as power-line-related, PLL-reference feed-through—and mixer-intermodulation spurs, as well as the PM noise commonly referred to as phase noise.

Reference feed-through is addressed by using a low PLL bandwidth such that the reference frequencies are easily filtered out due to their much higher frequencies. Power-line-related spurs fall into two main categories:

1. *120-Hz ripple:* Since most linear supplies are full-wave rectified, 120 Hz is the main interfering frequency output of the power supply. This ripple, if inadequately filtered, will show up as modulation on the output of the VCO. A 60-Hz subharmonic may also appear because the rectifier diodes do not conduct the same on each half cycle, although when this does happen, the amplitude is usually 10-40 dB below the 120-Hz component. 120-Hz contamination is predominately a problem of conduction through the power and ground wires.

2. *Magnetic field coupling:* The frequencies most commonly seen are 60 Hz and 180 Hz. 60 Hz is the fundamental frequency of the magnetic field in the transformer. When a magnetic-core transformer is driven relatively

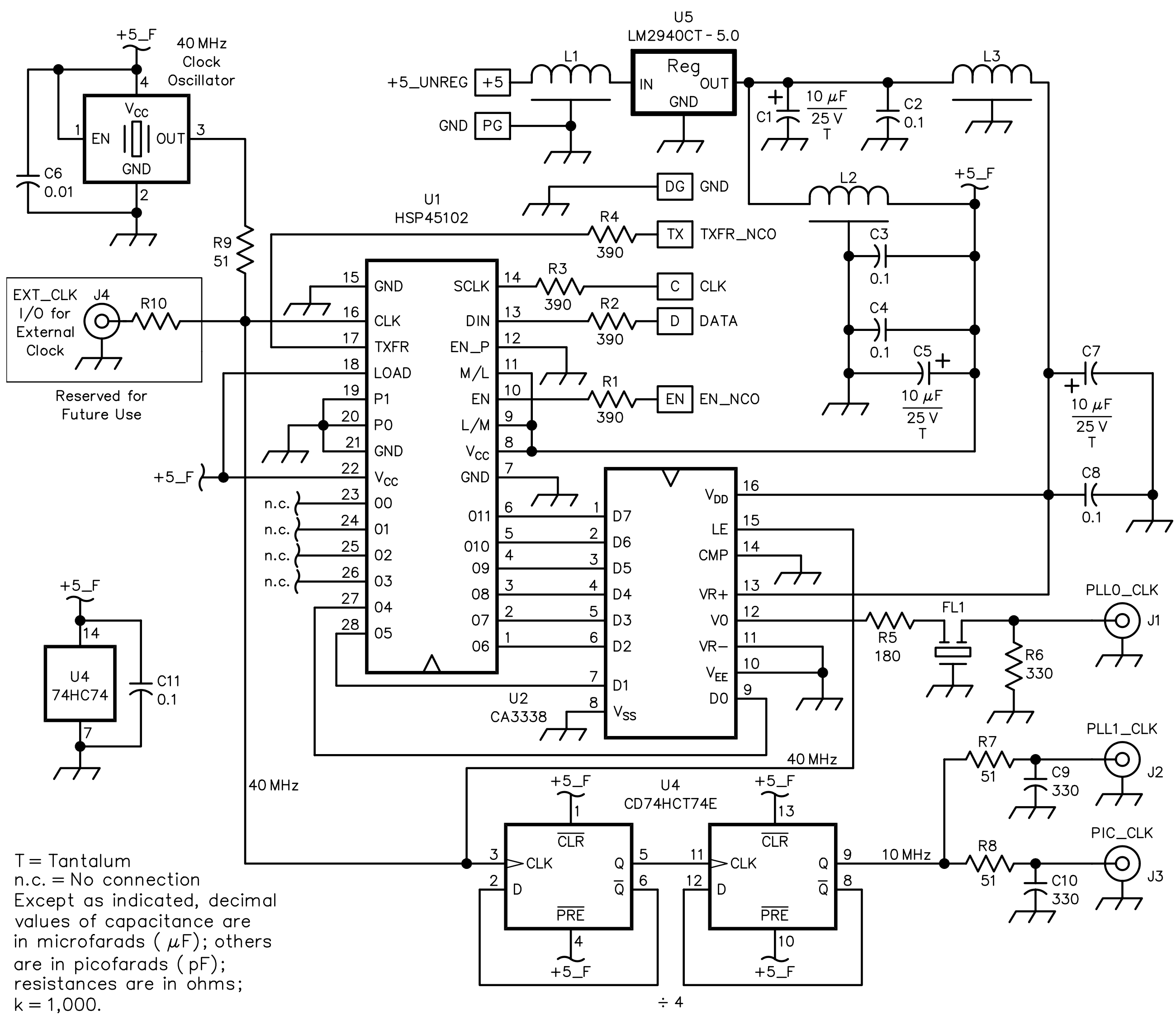


Fig 2—The NCO circuit is used to derive the master-clock references for the rest of the source. The NCO-driven, 10.7-MHz output is used as a super-fine adjustable reference for the PLL 0 board. PLL 1 and the PIC microprocessor are driven by fixed 10-MHz clocks. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. All capacitors are 50-V X7R ceramics unless otherwise noted.

J1-J3—SMB connector, PC board mount (Digi-Key #J648-ND)
L1-L3—Panasonic EMI filters (Digi-Key #P9807CT-ND)
U1—Harris HSP45102 NCO (Allied #HSP45102PC-40)
U2—Harris CA3338AE DAC (Allied #CA3338AE)
U3—40 MHz clock oscillator (Digi-Key #XC279-ND)
U4—CD74HCT74E (Digi-Key #CD74HCT74E-ND)
U5—LM2940CT 5 V regulator (Digi-Key #LM2940CT-5.0-ND)

hard, the magnetic field waveform begins to look like a square wave near saturation. This leads to the appearance of odd harmonics. So 60 Hz, 180 Hz (third harmonic) and 300 Hz (fifth harmonic) tend to show up on VCOs that are placed too close to power-line transformers.

The other common discrete spurs that show up in synthesizer circuits are unwanted mixing products and feed-through of various clocks. Unwanted mixing spurs are best remedied by removing spurious frequencies before mixing. I have done this here by low-pass filtering the PLL signals, by careful frequency planning and by using the proper drive levels to the mixer to limit spurious responses caused by overdrive, especially on the mixer's RF input.

Phase noise is a different sort of problem. Instead of discrete-frequency leakage, it is best thought of as "frequency smearing." On a spectrum analyzer, phase noise looks like an increase in the noise floor around a signal rather than a discrete spur. Phase-noise energy is spread across many frequencies and decreases in energy with increasing offset from the carrier.

Every component in this synthesizer potentially adds phase noise to the desired signal. In oscillator circuits, two basic tenets help improve the phase-noise performance:

- Use a high-Q resonator.
- Use lots of power in the resonator circuit (see Reference 5).

The master 40-MHz reference clock is a crystal-oscillator design and, as one would expect because of its high Q, it has very little phase noise: better than –100 dBc / Hz at 1 kHz offset from the carrier. Given the block diagram and the reference's phase noise, a calculation of the absolute-best phase noise at the output of the signal generator is possible. Since the reference oscillator is at 40 MHz and the output frequency of the PLL is 750 MHz, the frequency multiplication is 750 / 40 = 18.75. The phase noise increases by 20 dB for every 10 times the frequency is multiplied.

For a multiplication factor of 18.75, the phase noise will increase by at least:

$$20\log(18.75) = 25 \text{ dBc/Hz increase} \qquad \text{(Eq 2)}$$

So at the 1-kHz offset, we can expect a best-case phase noise of –100 + 25 = –75 dBc / Hz.

A frequency divider reduces the phase noise of a signal by a similar amount; ie, $-20\log N$ dB, where N is the counter's division ratio. The action of the NCO is that of a divider also, reducing the phase noise by its division ratio. Since the output frequency of the NCO is 10.7 MHz and the input clock is 40 MHz, the phase noise reduction is:

$$20\log\left(\frac{40}{10.7}\right) \approx 12 \text{ dBc/Hz} \qquad \text{(Eq 3)}$$

It's interesting that although we continue to divide the reference frequency in the PLLs to 100 kHz, the phase noise of the synthesizer overall will not decrease because of this division. This is because we need a frequency higher than the reference in the first place, so we eventually need to multiply the reference up, and the PLLs used here do so. They multiply the 100-kHz reference up to the order of 750 MHz, or 7500 times.

Division has limits, however. The CMOS dividers used in the MC145191 have a reported phase-noise floor of approximately –160 dBc / Hz at large offsets; below 10 Hz, the noise floor of the dividers also tends to increase (see Reference 1). So in reality, infinitely dividing a noisy reference can produce no better than the divider's phase-noise floor, approximately –160 dBc/Hz. This is low, but when multiplied up again by 7500 (a 77 dBc / Hz increase), the phase noise increases again.

This concept is more intuitive if we consider an analog example: If a signal is divided in amplitude by 1000, then amplified by 1000, it is pretty obvious that the output signal-to-noise ratio (S/N) will degrade from that of the original signal. The power divider has a noise figure equal to its attenuation and the amplifier has its own noise figure; hence, we add noise figure to the signal at every stage, decreasing its S/N. In the PLL-synthesizer's case, this increase in noise figure at every stage acts to increase the phase noise of the synthesizer.

Addition or subtraction of frequency —as in the mixer portion of this circuit—does not decrease phase noise; rather, the phase noise increases by the square root of the sum of the squares of the original signals being mixed. Therefore, if one source having lots of noise is mixed with a quiet one, then the total noise is essentially just the noisy source's contribution. Of course, a doubly balanced mixer such as the one used here also has loss and adds approximately 7 dB to the composite noise figure.

Now you should see what I mean by the "ignorance is bliss" statement above. Every signal path has the opportunity to worsen the phase noise of the original signal.

In my circuit, as in most PLL-based synthesizers, the greatest phase-noise contribution is that of the VCO. I used a commercial VCO for several reasons. First, if I used world-class VCO circuits like those Rhode described in Reference 2, the component count would have skyrocketed beyond what was practical for this project. Second, we can get a very good performance-to-price ratio and save building time by using a modular VCO. We must pay attention to PLL loop design and take care that the rest of the signal paths to not add unduly to the overall phase noise.

A major part of the PLL board's circuitry (Fig 3) is contained in the Motorola MC145191 IC. This device contributes very little phase noise, and its use of a current-output charge pump helps to limit noise. This is so because the output essentially goes to a high-impedance state when locked and only small correction pulses are added from time to time to maintain stability.

The charge pump and error amplifier can have significant impact on phase noise for several reasons, including noise in bias currents, input-referred noise and loop-filter design. This is because any unwanted signal on the VCO-tune line will show up as modulation on the VCO output. The VCO used has a tuning sensitivity of about 25 MHz / V. So if there's a microvolt of extra noise on the tuning line, output modulation of 25 Hz will be seen. If this extra microvolt is random in nature and has a moderately wide bandwidth, the effective phase-noise floor of the entire circuit will suffer.

I made one concession to greatly improve performance in the PLL circuit; I use an industry-workhorse, low-noise operational amplifier: the OP-27.[6] The OP-27 has noise at least 20 times lower than most commonly available, single-supply op amps on the market. This required the addition of a negative-bias supply (thus complicating the power-supply circuitry), but I felt the increased performance was worth the cost.

You hear of PLL circuits having a bandwidth. This bandwidth is set by the RC frequency-shaping networks between the phase-detector output and the VCO tuning-voltage input; it is required to be within certain ranges by the laws of circuit design and practically. Bandwidths less than 300 Hz, or so, tend to suffer from microphonics. On the other hand, the maximum bandwidth is on the order of 5 to 10 times less than the PLL reference frequency. Since the phase detector employs a

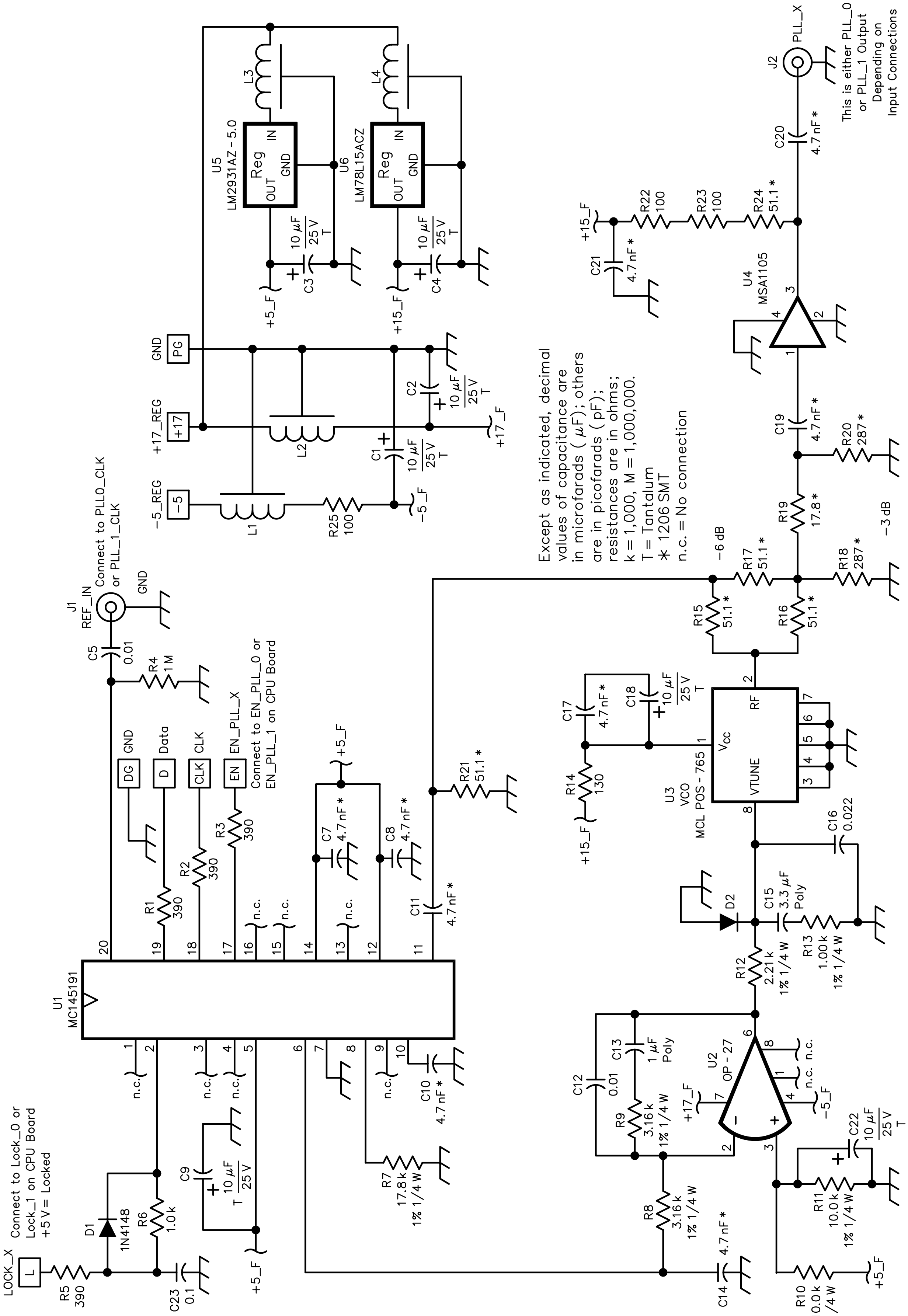

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000, M = 1,000,000.
T = Tantalum
∗ 1206 SMT
n.c. = No connection
Connect to Lock_0 or Lock_1 on CPU Board +5 V = Locked
Connect to PLL0_CLK or PLL_1_CLK
Connect to EN_PLL_0 or EN_PLL_1 on CPU Board
This is either PLL_0 or PLL_1 Output Depending on Input Connections
U1 MC145191
U2 OP-27
U3 VCO MCL POS-765
U4 MSA1105
U5 LM2931AZ-5.0
U6 LM78L15ACZ

Fig 3

sampling process, the Nyquist bandwidth limits also apply. Between these two extremes are the design tradeoffs that can consume many days of experimentation and analysis (see References 1, 2 and 5). Large bandwidth is desirable for fast switching, but less bandwidth is needed to limit reference frequency feed-through.

Below the crossover frequency of a carefully designed PLL, the loop can be made to reduce phase noise by an amount equal to the loop gain of circuit. This can be used to advantage because it will help to clean up a noisy VCO. The limit on how clean the VCO can be made is set by the characteristics of the reference frequency. Above the loop bandwidth, the output noise is dominated by the VCO noise alone because the loop is not able to correct for perturbations.

Usually, phase noise is traded off against settling-time and reference-spur requirements. I chose a loop bandwidth of approximately 1 kHz, since this met several design goals simultaneously. The multiplied reference-frequency phase noise crosses the VCO phase noise at approximately this frequency. A 1-kHz crossover frequency allows for a PLL settling time of approximately three milliseconds (to 0.1%). The PLL susceptibility to microphonics with a 1-kHz bandwidth is reduced and the 100-kHz reference is pretty easy to filter with passive parts without seriously effect-ing the loop's gain and phase-margin performance.

Power-Supply Design

At offsets below 500 Hz, the power-supply circuitry is critical. I used some off-the-shelf toroidal transformers (see Fig 6) in the power supply because these units significantly reduce any 60, 180 and 300-Hz magnetic fields without adding any special µ-metal shielding.

Within a toroidal-coil structure, the magnetic field is theoretically contained completely inside the core. This is not the case with transformers using E-I laminations, where the field "blows" out at the corners. In addition, the use of a distributed-power architecture—where each circuit has its own local regulators—serves to limit conducted noise in the sensitive PLL and output circuits. Remember: A microvolt at the VCO input pin produces 25-Hz FM on the output, so conducted and radiated noise must be contained before it gets to any of those circuits. Putting each RF circuit block in its own die-cast, metal box serves to shield any spurious signals entering or exiting to any of the other circuit blocks, too.

Fig 3 (see left)—The same circuitry is used for PLL 0 and PLL 1 boards. The heart of each PLL board is the Motorola MC145191. A Mini-Circuits VCO is used because of its high performance and because use of a commercial VCO minimizes assembly time. Unless otherwise specified, use $^{1}/_{4}$ W, 5%-tolerance carbon composition or film resistors. All capacitors are 50-V X7R ceramics unless otherwise noted. "Poly" indicates polyester capacitors.
J1, J2—SMB connector, PC board mount (Digi-Key J648-ND)
L1-L4—Panasonic EMI filters (Digi-Key P9807CT-ND)
U1—MC145191 PLL (Newark #MC145191F)
U2—OP-27 op amp (Newark #OP-27GP)
U3—465-765 MHz VCO module (Mini-Circuits Labs #POS-765)
U4—HP MSA-1105 MMIC (Newark #MSA-1105)
U5—LM2931AZ 5.0 V regulator (Digi-Key #LM2931AZ-5.0-ND)
U6—LM78L15ACZ 15 V regulator (Digi-Key #LM78L15ACZ-ND)

Performance

The source's phase-noise performance is shown in Fig 7. Note that respectable phase noise of better than –91 dBc / Hz is achieved at 10 kHz offset from the carrier. This should be more than acceptable for most homebrewing test needs. Fig 8 shows the close-in spurious performance. The reference feed-through is better than –79 dBc.

Harmonic performance is one area in which a substantial design tradeoff was made in the name of simplicity. In a commercial synthesizer, sub-octave filters might have been used to reduce the harmonics generated by the out-put amplifier chain. This would have required much more complexity than was deemed practical for this project. This source has one band-limiting filter in the forward signal path. The 270-MHz low-pass filter I used strips off the sum frequencies of the mixer and helps to reduce the RF and LO feed-through of the mixer, but does little to improve harmonics below 150 MHz. Even so, at an output power of 0 dBm, harmonics are typically below –45 dBc from 20 MHz to 250 MHz. Below 20 MHz, the PIN-diode attenuator used on the mixer board adds most of the distortion.[7] At an output frequency of 2 MHz, the harmonics rise to –30 dBc. If improved harmonic performance is needed for a special situation, Mini Circuits BNC filters[8] can be added to the output as required.

Non-harmonic spurious responses also result when signals are mixed. The frequency plan of this design was constrained by the availability of suitable VCOs and the desire to make it reasonably repeatable by a homebrewer.[9] Even with these restrictions, the frequency plan produces very few "crossing spurs." Crossing spurs occur where $(nF_1 + mF_2)$ coincides with the desired output.

The worst crossing spur happens at 250 MHz, where a rather large third-order and two smaller seventh- and eighth-order (LO to RF) spurs cross the desired output. The next-largest spur crossing happens at 187.5 MHz, where fifth- and ninth-order spurs cross the desired signal. A much less problematic crossing occurs at 150 MHz, where a seventh-order spur crosses. The final spur crossing, at 125 MHz, is a seventh-order spur that (at least with my spectrum analyzer) is in the noise level (ie –80 dBc).

A spur-avoidance technique is applied in the PC control program (see below) that basically shifts the LO frequency down to 740 MHz within 1 MHz either side of the spur crossings at 250, 187.5 and 150 MHz. The RF is similarly shifted to keep the spurs away from the carrier by at least 4 MHz. This is perhaps reasonable for most general uses, but if the source were used as a LO in a receiver, we would also have to avoid the image frequencies of the receiver's mixer. Since I can't envision every possible use of the source, I can't take care of these crossings for every possible case without significantly increasing the control program's complexity. The control program has the capability of disabling its spur-avoidance behavior if needed, however. Refer to Table 1 for a summary of typical performance.

Software for the Source

Software support is provided in two forms. For stand-alone applications, a 32-bit, *Windows-95* program[10] is provided to drive the source. With this program, the frequency and output amplitude may be set quickly to any desired value by means of the Windows graphical user interface (GUI).

For computerized test applications, I have also provided what is known as an "Active-X" control. This 32-bit control is much like a dynamic linked library (DLL). Many *Windows-95*-compliant programs may use Active-X controls, including Word and Excel. Programming languages such as *Visual Basic* and the specific-test generation program HP *VEE*[11] can also use Active-X controls. Basically, the Active-X control is a standard library format that encapsulates the functionality of the source into a single well defined component that many Windows programs can plug into and use.

All of the software described including the *Visual Basic* 5 source code and the firmware for the PIC may be downloaded from the ARRL's Web site.[12]

Source Construction

If you have never worked with surface mount technology (SMT) but wish to learn with a relatively simple project, then look no further. The source's circuitry has been designed and optimized to use the absolute-minimum number and variety of SMT parts. In fact, only eight different SMT parts are required for construction! Three of these are ICs; the rest are resistors and ceramic capacitors. The shipping charges for the passive components is greater than the part costs!

Recent editions of *The ARRL Handbook* and recent *QST* articles[13] provide describe SMT-assembly techniques and—with a small soldering iron tip—it's not difficult at all. In fact, now that I'm used to the technology, I find prototype building with SMT parts faster and easier than using leaded components. SMT really is something to be learned and used—not avoided!

The synthesizer circuitry is partitioned exactly as is the block diagram. All SMT components mount on the trace side of the circuit boards[14] as shown in Figs 9 and 10. This allows easy construction of the SMT parts first, followed by the addition of the through-hole parts.

Each circuit board is sized to fit in a die-cast aluminum box, obtained from Jameco. The RF connectors (marked "Jx" on the schematic) are all on one side of each PC board, so that the box may be drilled and the PC board slipped inside as shown in Fig 11. The use of die-cast boxes greatly reduces radiation coupling between circuit sections.

The dc and programming connec-tions may be made by passing insu-lated wires through holes in the box. As long as holes are kept smaller than 1/4 inch, feed-through capacitors are not needed. The only precaution I took with the PLLs was to pass the power leads through a Radio Shack snap-on common-mode filter (#273-105). It is a reasonable precaution to put one of these filters on the RS-232 connection and the power cord also, just to be sure that no "nasties" get into or out of the source.

The NCO and Mixer PC boards have large, TO-220-packaged voltage regulators on one side of the board. The tabs should be attached to the die-cast box for extra heat sinking. The TO-220 tabs need no insulation as they are at ground potential.

The RF connections between modules use SMB, snap-on RF connectors. The mating sockets (Digi-Key #ARF1235-ND) are very easy to assemble using RG-174 subminiature coax. I find these SMB cables are easier to put together than BNCs, which never seem to come out right for me.

The Mini-Circuits modules are supplied in small, steel-pin-mount packages. After the basic functionality of the circuit is tested, these packages should be tack-soldered to the circuit board's topside ground plane on both long edges. This helps to reduce the inductance between the package and circuit ground and improves shielding.

I also took the precaution of mounting the PLL modules using double-sided foam tape and rubber grommets. This helps damp any vibration from other pieces of test equipment.

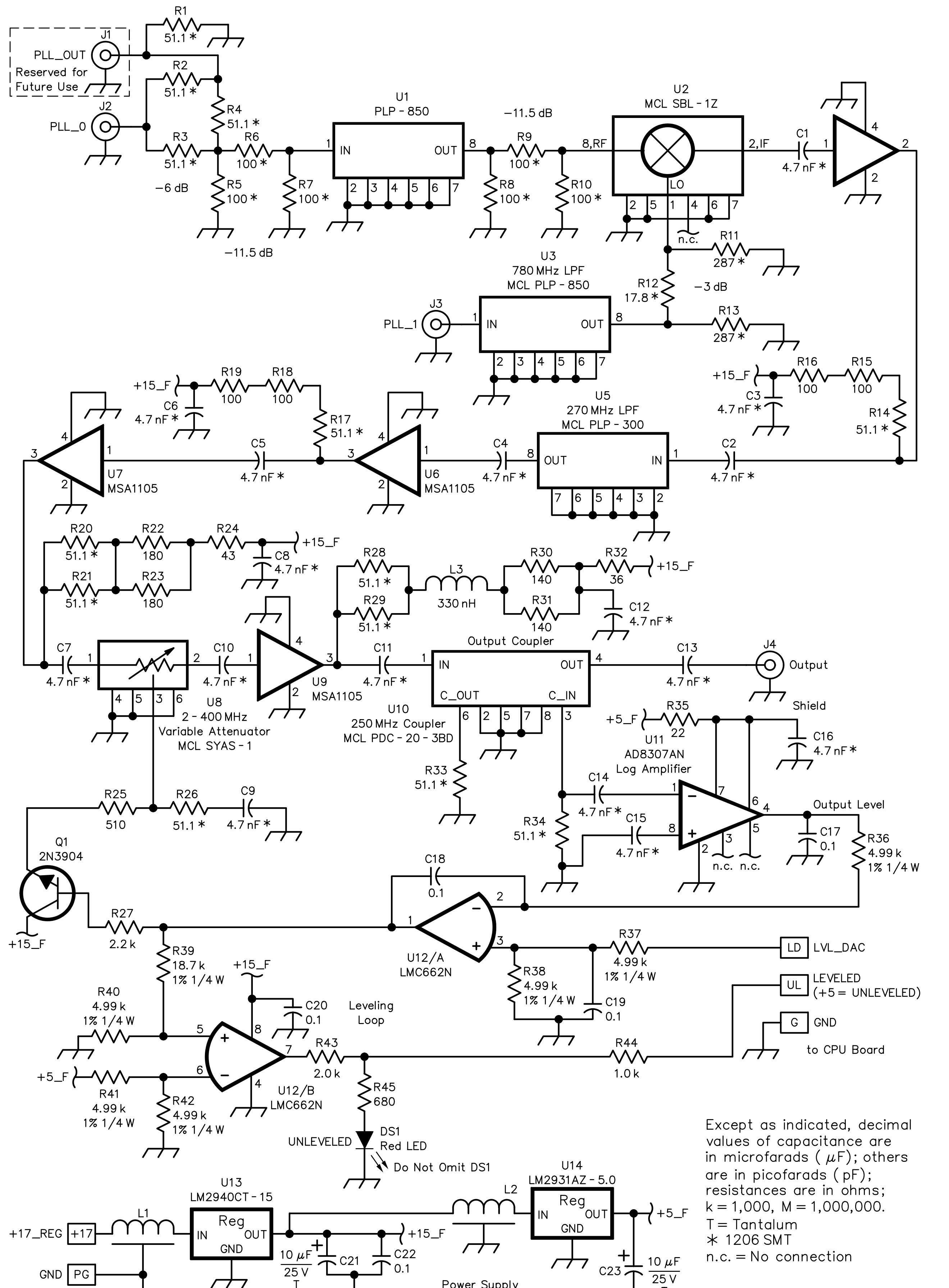
If you are going to use the source as an RF input to a receiver, you should also take some special precautions in

Table 1—Typical Performance Summary

Frequency Range	2-250 MHz
Frequency Resolution	1 Hz
Amplitude Range (calibrated)	–15 to +15 dBm
Amplitude Resolution	0.1 dB
Maximum output	+17 dBm
Amplitude Accuracy	±0.5 dB over full frequency and power range
Frequency Accuracy Error	Correctable to 0 Hz frequency drift per hour (after warmup): 0.0002 %
Frequency Stability	0.001 % over range of 15-35°C
Phase Noise	–64 dBc/Hz @ 1 kHz
	–90 dBc/Hz @ 10 kHz
	–110 dBc/Hz @ 100 kHz
Harmonics:	2 MHz to 20 MHz < –30 dBc
	20 MHz to 250 MHz < –45 dBc
Spurious:	2 to 220 Mhz < –65 dBc
	220 to 250 MHz < –35 dBc
LO and RF feed-through	< –55 dBm
Clock feed-through (fundamental and harmonics)	< –85 dBm
Output Match (SWR)	2 MHz to 7 MHz < 2:1
	7 MHz to 250 MHz < 1.3:1

Note: Output at 100 MHz, 0 dBm output power and temperature of 25°C, unless otherwise specified.

Fig 4 (see right)—The Mixer board is where the source comes together. The PLL-module outputs are difference-mixed to provide an output of 2 to 250 MHz. A leveling circuit keeps the output level fixed regardless of load mismatch. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. All capacitors are 50-V X7R ceramics unless otherwise noted. "Poly" indicates polyester capacitors.
J1, J4—SMB connector, PC board mount (Digi-Key J648-ND)
L1-L2—Panasonic EMI filters (Digi-Key P9807CT-ND)
L3—330 nH 100 mA RFC (Digi-Key M9R33-ND)
U1, U3—780 MHz low-pass filter (Mini-Circuits Labs #PLP-850)
U2—10-1000 MHz mixer (Mini-Circuits Labs #SBL-1Z)
U4, U6, U7, U9—HP MSA-1105 MMIC (Newark #MSA-1105)
U5—270 MHz low-pass filter (Mini-Circuits Labs #PLP-300)
U8—2-400 MHz variable attenuator (Mini-Circuits Labs #SYAS-1)
U10—250 MHz coupler module (Mini-Circuits Labs #PDC-20-3BD)
U11—AD8307AN logarithmic amplifier (Newark # AD8307AN)
U12—LM662N op amp (Digi-Key #LMC662N-ND)
U13—LM2940CT 15 V regulator (Digi-Key #LM2940CT-15.0-ND)
U14—LM2931AZ 5.0 V regulator (Digi-Key #LM2931AZ-5.0-ND)


J1
PLL_OUT
Reserved for
Future Use
J2
PLL_0
R1
51.1 *
R2
51.1 *
R3
51.1 *
R4
51.1 *
R5
100 *
R6
100 *
R7
100 *
−6 dB
−11.5 dB
U1
PLP - 850
IN
OUT
−11.5 dB
R9
100 *
R8
100 *
R10
100 *
U2
MCL SBL - 1Z
8,RF
2,IF
LO
n.c.
C1
4.7 nF *
R11
287 *
R12
17.8 *
−3 dB
R13
287 *
U3
780 MHz LPF
MCL PLP - 850
J3
PLL_1
+15_F
C6
4.7 nF *
R19
100
R18
100
R17
51.1 *
U7
MSA1105
C5
4.7 nF *
U6
MSA1105
C4
4.7 nF *
U5
270 MHz LPF
MCL PLP - 300
C2
4.7 nF *
+15_F
C3
4.7 nF *
R16
100
R15
100
R14
51.1 *
R20
51.1 *
R21
51.1 *
R22
180
R23
180
R24
43
C8
4.7 nF *
+15_F
R28
51.1 *
R29
51.1 *
L3
330 nH
R30
140
R31
140
R32
36
+15_F
C12
4.7 nF *
C7
4.7 nF *
C10
4.7 nF *
U8
2 - 400 MHz
Variable Attenuator
MCL SYAS - 1
U9
MSA1105
C11
4.7 nF *
Output Coupler
IN
OUT
C_OUT
C_IN
U10
250 MHz Coupler
MCL PDC - 20 - 3BD
C13
4.7 nF *
J4
Output
Shield
+5_F
R35
22
C16
4.7 nF *
U11
AD8307AN
Log Amplifier
R33
51.1 *
R25
510
R26
51.1 *
C9
4.7 nF *
C14
4.7 nF *
C15
4.7 nF *
R34
51.1 *
Output Level
C17
0.1
R36
4.99 k
1% 1/4 W
n.c. n.c.
Q1
2N3904
+15_F
C18
0.1
R27
2.2 k
U12/A
LMC662N
R37
4.99 k
1% 1/4 W
LD LVL_DAC
R38
4.99 k
1% 1/4 W
C19
0.1
R39
18.7 k
1% 1/4 W
+15_F
C20
0.1
Leveling
Loop
UL LEVELED
(+5 = UNLEVELED)
G GND
to CPU Board
R40
4.99 k
1% 1/4 W
+5_F
R41
4.99 k
1% 1/4 W
R42
4.99 k
1% 1/4 W
U12/B
LMC662N
R43
2.0 k
R44
1.0 k
R45
680
DS1
Red LED
UNLEVELED
Do Not Omit DS1
Except as indicated, decimal
values of capacitance are
in microfarads (μF); others
are in picofarads (pF);
resistances are in ohms;
k = 1,000, M = 1,000,000.
T = Tantalum
* 1206 SMT
n.c. = No connection
U13
LM2940CT - 15
Reg
IN
OUT
GND
+17_REG +17
L1
GND PG
10 μF
25 V
T
C21
C22
0.1
+15_F
L2
Power Supply
U14
LM2931AZ - 5.0
Reg
IN
OUT
GND
+5_F
C23
10 μF
25 V
T

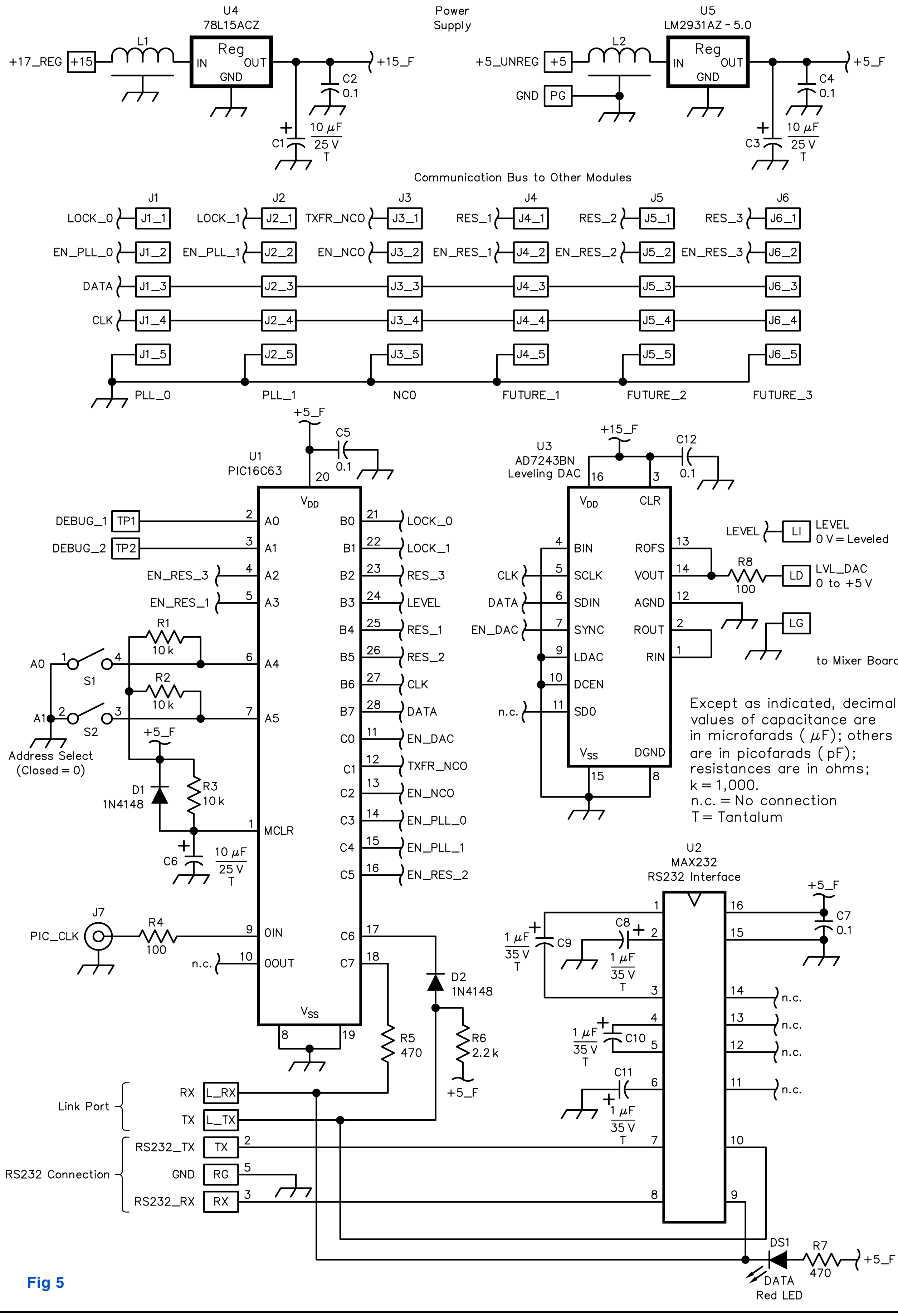

Power Supply
U4
78L15ACZ
Reg
IN
OUT
GND
+17_REG
+15
L1
+15_F
C2
0.1
C1
10 μF
25 V
T
U5
LM2931AZ - 5.0
+5_UNREG
+5
L2
GND
PG
C4
C3
+5_F
Communication Bus to Other Modules
J1
J2
J3
J4
J5
J6
LOCK_0
LOCK_1
TXFR_NCO
RES_1
RES_2
RES_3
EN_PLL_0
EN_PLL_1
EN_NCO
EN_RES_1
EN_RES_2
EN_RES_3
DATA
CLK
PLL_0
PLL_1
NCO
FUTURE_1
FUTURE_2
FUTURE_3
U1
PIC16C63
C5
0.1
VDD
VSS
DEBUG_1
TP1
DEBUG_2
TP2
A0
A1
A2
A3
A4
A5
B0
B1
B2
B3
B4
B5
B6
B7
C0
C1
C2
C3
C4
C5
C6
C7
MCLR
0IN
0OUT
LEVEL
EN_DAC
EN_PLL_0
EN_PLL_1
R1
10 k
R2
10 k
S1
S2
Address Select
(Closed = 0)
D1
1N4148
R3
10 k
C6
10 μF
25 V
T
J7
PIC_CLK
R4
100
n.c.
D2
1N4148
R5
470
R6
2.2 k
U3
AD7243BN
Leveling DAC
+15_F
C12
0.1
CLR
BIN
SCLK
SDIN
SYNC
LDAC
DCEN
SDO
ROFS
VOUT
AGND
ROUT
RIN
DGND
LEVEL
LI
LEVEL
0 V = Leveled
R8
100
LD
LVL_DAC
0 to +5 V
LG
to Mixer Board
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
n.c. = No connection
T = Tantalum
U2
MAX232
RS232 Interface
C7
0.1
C8
C9
C10
C11
1 μF
35 V
T
Link Port
RX
L_RX
TX
L_TX
RS232 Connection
RS232_TX
TX
GND
RG
RS232_RX
RX
DS1
R7
470
DATA
Red LED

Fig 5

the housing of the completed unit to achieve minimum RF leakage. As Fig 11 shows, the chassis I used for the construction of the source is painted with a nice, scratch-resistant finish. While pretty, it does not allow the chassis to be sealed, in an RF sense. For minimum leakage, all the joints of the chassis must be sanded to bare metal, so that each piece makes good contact with the others. Another option is to build a sub-chassis from scrap copper-clad PC board material and place all the RF modules in this box, then place this sub-box inside the painted chassis. The power supply, as shown in Fig 12, is constructed bread-board-style on a piece of Vector "perf" board because of its simplicity.

Calibration

Notice that there is not a single adjustment in this entire design—well, okay, except for the power supply. There are only two error sources of concern in the design. The first is the accuracy of the frequency reference. A 0.005% (50 ppm) error in the 40-MHz reference translates directly to a 0.005% error in the source's output frequency. This is very easy to correct in software. If an accurate measurement can be made of source frequency using, say, a frequency counter, then the error is corrected with the PC control program.

The other important error source is the output-leveling circuit. Since the coupler and log amplifier are subject to gain and offset errors, some means

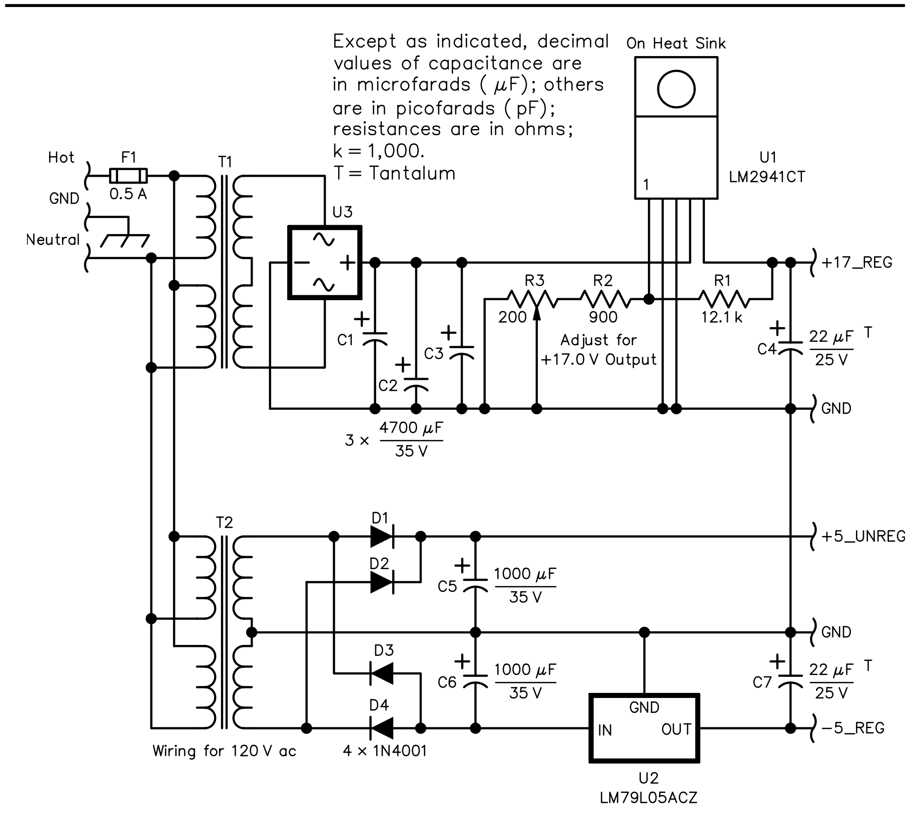
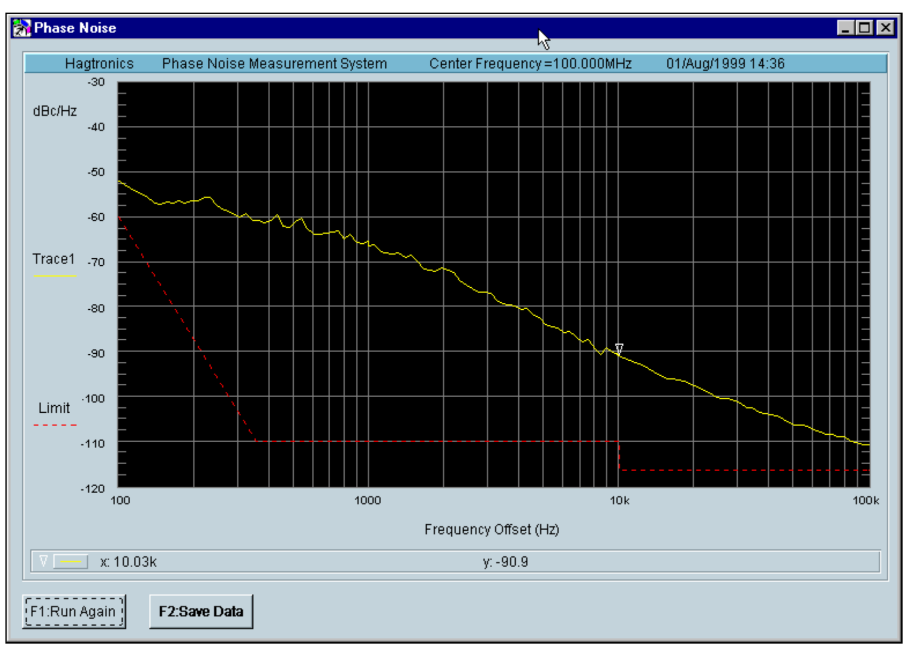


Fig 6—The power supply uses toroidal transformers to prevent radiated magnetic fields from modulating the source output. As a further noise-reduction technique, all of the critical RF modules contain their own local linear voltage regulators to prevent cross-talk. Unless otherwise specified, use 1/4 W, 1%-tolerance carbon composition or film resistors.

R3—200 Ω single-turn trimpot, 5%
T1—120/240-V primary, 14-V center-tap secondary, 1.07 A (manufactured by Talema Inc, Digi-Key #TE70050-ND)
T2—120/240-V primary, 14-V center-tap secondary, 0.5 A (manufactured by Talema Inc, Digi-Key #TE70030-ND)
U1—LM2941CT 17-V regulator (Digi-Key #LM2941CT-ND)
U2—LM79L05ACZ 5 V regulator (Digi-Key #NJM79L05A-ND)
U3—4-A, 100 PIV bridge rectifier (RadioShack #276-1171)

Fig 5 (see left)—The PIC16C63 microprocessor acts as a "smart UART," receiving commands from the PC via an RS-232 port and setting the source hardware appropriately. The RS-232 connection operates at a speedy 19.2 kbaud. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors.
L1-L2—Panasonic EMI filters (Digi-Key P9807CT-ND)
S1, S2—Two-position DIP switch (Digi-Key #GH1102-ND)
U1—PIC16C63 programmed, see text (Digi-Key #PIC16C63-20ND)
U2—MAX232CPE RS-232 ASIC (Digi-Key #MAX232CPE-ND)
U3—AD7243BN (Newark # AD7243BN)
U4—LM78L15ACZ 15 V regulator (Digi-Key #LM78L15ACZ-ND)
U5—LM2931AZ 5.0 V regulator (Digi-Key #LM2931AZ-5.0-ND)

Fig 7(see right)—The phase noise of the completed source is a respectable –91 dBc/Hz at a 10 kHz offset from the carrier.



of correction is needed. This is done using a DAC on the CPU board. The DAC output range and resolution exceed those needed in the circuit. The procedure is to set a level with the PC control program and measure the output with a power meter, spectrum analyzer or scope (in order of decreasing accuracy). The frequency used for leveling-circuit power correction is 10 MHz, making a low-frequency scope measurement practical.

A linear-regression, straight-line curve is then fit to the data and input to the PC control program. This linear equation is then used to determine the proper DAC output corresponding to power levels from –20 to +15 dBm.

Other leveling-circuit errors are caused by the coupler and log-amplifier amplitude-versus-frequency errors. These errors are accounted for in a manner similar to the above. At 10-MHz frequency intervals, the power is measured with the output set to 0 dBm. A simple correction curve is fit to the data.

These two leveling corrections provide better than 0.5-dB output accuracy over the entire frequency and output-power ranges of the source. If you don't have the equipment available to calibrate the source, I will supply default data that should typically provide better than 1.5 dB accuracy in the leveling circuits. The *Windows-95* control program also allows a pure power offset to be added to the calculated powers. This is useful to correct for the gain or loss of external circuits connected to the source.

The Most Popular Question

If this project is like every other I have done, the most popular question will be: "Can I extend the frequency to 500 MHz?" Well, as in all projects, design goals must be traded against complexity, performance and practicality. 250 MHz was chosen because it is a reasonable upper limit of what can be assembled breadboard-style by a reasonably skilled RF builder. Getting back to the question however, the answer is: "Yes, the frequency can be extended, but not in the sense of getting higher-frequency VCOs or mixers." An external frequency doubler (Mini Circuits Model FD-2) can be used on the output of the generator to get to 500 MHz, with certain distinct drawbacks. Since the doubler conversion loss is 13 dB, a maximum output of only +2 dBm at 500 MHz is possible. The fourth harmonic from this doubler is only 18 dB below the second, however, so a 1000-MHz signal can be generated at about –16 dBm. The phase noise degrades by 6 dB per doubling, but the resulting performance is still very respectable.[15]

Notes

[1]V. Manassewitsch, *Frequency Synthesizers* (New York: John Wiley and Sons, 1987).

[2]U. Rohde, "Key Components of Modern Receiver Design," Parts 1 and 2, *QST*, May pp 29-32, and June pp 27-31, 1994.

[3]Qualcomm is assigned a patent on this approach (US Patent Number 4,965,533). Many books and articles reference the technique but not the patent (Rhode mentions both and a prior use in a Rhode & Schwartz synthesized source). The Qualcomm Web site has information on this technique and many other interesting products at **www.qualcomm.com**. [My research reveals use of the DDS-driven PLL as early as 1984. It's likely QEX will soon publish an article written by one of the first investigators of this

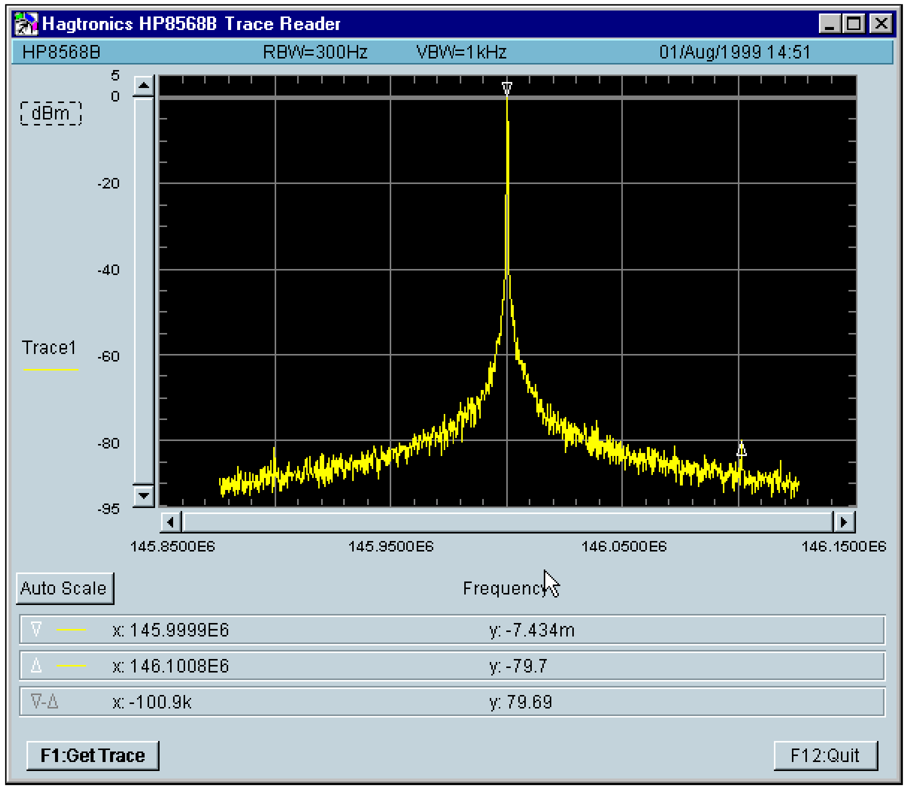


Fig 8—The close-in spurious performance of the source shows a reference rejection of –80 dBc. The span for this display is ±125 kHz from the carrier.

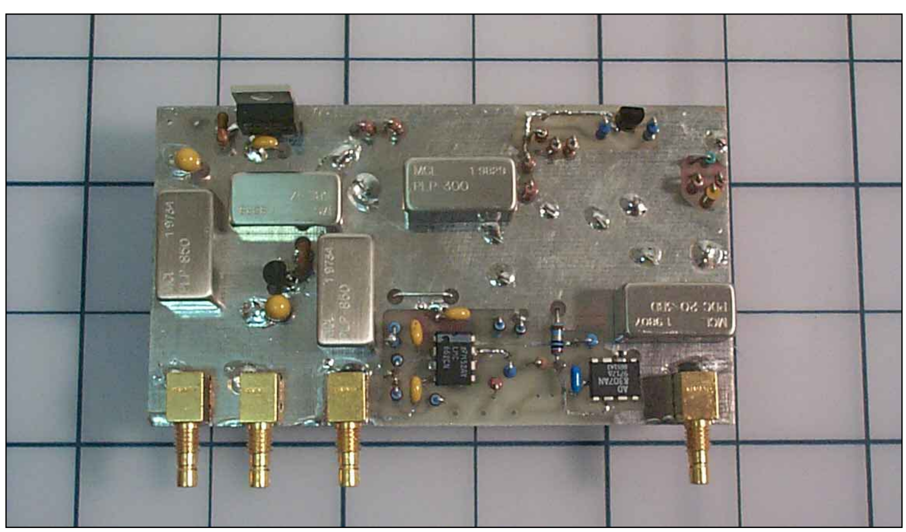


Fig 9—The PC boards are constructed with the through-hole components mounted to the top of the board. This is a view of the Mixer board component side.

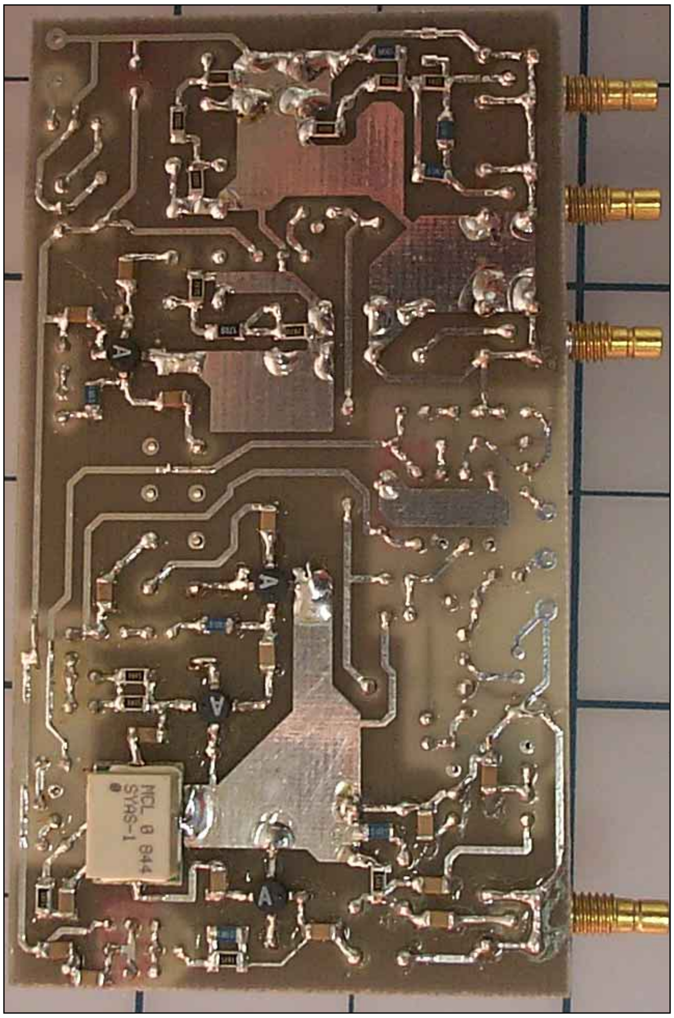


Fig 10—The SMT components are attached to the solder side of the PC boards. This is a view of the Mixer board solder side showing SMT construction.

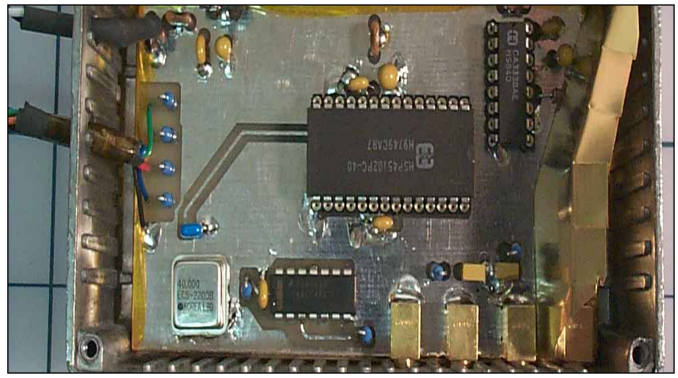


Fig 11—All major circuit blocks in the source are designed to be mounted in low-cost, die-cast enclosures from Jameco. Here is a view of a completed NCO board with some added shielding (on the right side) around the 10.7-MHz filter to reduce any clock feed-through to the 10.7 MHz PLL reference signal.

Fig 12—What all your hard work can produce—your own VHF synthesized source. I used a new Bud chassis to house my source, but a surplus enclosure can work equally well.

technology. I think it will shed some light on things.—*Ed.*]

[4]S. Adam, *Microwave Theory And Applications* (New Jersey: Prentice-Hall, 1969).

[5]R. Rhea, *Oscillator Design and Computer Simulation*, Second Ed. (New York: McGraw-Hill, 1995).

[6]The OP-27 data sheet is available from Analog Devices at **www.analog.com**. It is interesting to note that Analog Devices has recently introduced some low-noise, single-supply op amps, but their availability to homebrewers is limited at this time.

[7]The PIN-diode attenuator and the low-frequency performance of the mixer/amplifier chain are what set the low-frequency limit of this design. The basic topology (with more complex circuitry) can theoretically produce an output to dc, limited only by the residual FM of the PLL.

[8]Mini Circuits Labs, PO Box 350166, Brooklyn, NY 11235-0003; tel 800-654-7949, 718-934-4500, fax 718-332-4661; **http://www.minicircuits.com/**.

[9]This design can be built by an experienced RF builder with little difficulty. In fact, that is how this project started. To see a photo of my original breadboard, visit my Web site at **www.sonic.net/~shageman**.

[10]The control program requires *Windows* 95, 98 or NT with one 16650-controlled serial port available for communication and 3 MB of available hard-disk space.

[11]HP *VEE* is a graphical test-generation program that can use Active-X components. Visit **www.tmo.hp.com** and search for "VEE."

[12]You can download this package from the ARRL Web site at **www.arrl.org/files/qex/**. It includes complete parts lists. Look for 0100HAGE.ZIP. A programmed PIC is available from the author for $30.00 US, postpaid in the US and Canada. Add $5 to cover postage and handling to Europe, $15 to South America. Send payment as a check or money order, no credit cards.

[13]S. Ulbing, "Surface Mount Technology—You Can Work with It!" Parts 1, 2 and 3, *QST*, April-June, 1999.

[14]FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; tel 847-836-9148 (Voice mail), fax: 847-836-9148 (same as voice mail); **farcir@ais.net**; **http://www.cl.ais.net/farcir/**.

[15]I maintain a FAQ page at **www.sonic.net/~shageman**. Be sure to check there for updated information on this project and others.

Effects of Boom and Element Diameters on Yagi Element Lengths at 144, 432 and 1296 MHz

Want to build some VHF/UHF/Microwave Yagis? Some up-front measurements can cut the time required for tuning. Use this method to determine boom-correction factors for Yagi elements mounted through the middle of —and in good electrical contact with—a metal boom.

By Guy Fletcher, VK2KU

Editor's note: Since this article originally appeared in Amateur Radio *magazine of the WIA in March 1999, corrections and clarifications have been provided by the author.*

My experiments were performed at frequencies of 144.2, 432.2 and 1296.2 MHz and provide data for boom diameters of up to 0.08 λ. They also explore the effect of element diameter. The results show clearly that the correction depends not only on boom diameter but also on element diameter and element length.

The effect of the boom on each element may be represented by a *negative* reactance at the center of the element. A simple empirical formula for this reactance agrees well with all the experimental data and allows correction for any combination of boom diameter and element diameter. These results are given in the form of a universal graph.

The observed dependence on element length is intrinsic to the model of boom reactance and leads to a correction that tapers as the element length decreases. This may be adequately represented in practice by a simple power-law modification of the value for a standard element length of 0.42 λ taken from the graph.

The use of tapered corrections for the different element lengths (rather than a single fixed correction) has been applied to examples of practical Yagis. The difference is negligible at 144 MHz and small at 432 MHz. At 1296 MHz, however, where boom diameters may be relatively large (in terms of wavelength—*Ed.*), the use of a fixed correction appears to change the performance parameters of the antenna quite significantly.

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Background

Boom correction factors are discussed by Günter Hoch, DL6WU, in the *VHF / UHF DX Book* (edited by Ian White, G3SEK) and other similar references.[1] Günter's corrections may

[1] I. White, G3SEK, Ed., *VHF/UHF DX Book.* This book is available from your local ARRL dealer or directly from the ARRL as #5668. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line at **http://www.arrl.org/catalog**.

be embodied in a formula developed by Ian:

$$\frac{C}{B} = 25.195\left(\frac{B}{\lambda}\right) - 229\left(\frac{B}{\lambda}\right)^2 \qquad \text{(Eq 1)}$$

where C is the correction and B is the boom diameter, both in millimeters. This formula is not valid for boom diameters greater than 0.055 λ, although diameters of up to 16 mm (0.07 λ) are common at 1296 MHz. Ian's formula is plotted in Fig 1. It includes no dependence on element diameter or length. Also, the curve is assumed to pass through the origin, although there is no real reason to expect this. C is obviously zero when B is zero, but the ratio C/B need not be zero to make C zero.

There seem to be no data available for larger boom diameters, which is perhaps why some amateurs have remarked on the difficulty of matching antennas correctly at 1296 MHz. The experiments to measure boom corrections are in fact quite straightforward, so I decided to make some simple measurements. The scope of the project expanded rapidly as the unexpected nature of the results appeared.

Theory and Model

The complex voltage reflection coefficient (ρ) represents the magnitude and relative phase of the ratio of the reflected voltage wave to the forward voltage wave at a load. In these experiments, the reflected power (P_R) and the forward power (P_F) were measured rather than the voltages:

$$|\rho| = \left(\frac{P_R}{P_F}\right)^{\frac{1}{2}} \qquad \text{(Eq 2)}$$

The voltage standing wave ratio (SWR), σ, is:

$$\sigma = \frac{(1+|\rho|)}{(1-|\rho|)} \qquad \text{(Eq 3)}$$

Any element of a Yagi antenna has energy stored in the fields surrounding it. Near the element center, the current is large and the voltage small; near the ends, the current is small and the voltage large. If the element passes through a larger conductive boom at its center, the skin effect forces the current to flow around the outside of the boom instead of directly along the element's surface. This reduces the volume of the magnetic field around the element and therefore reduces the energy stored there. Since the stored energy is directly proportional to the self-inductance (L) of the element, the effect of the boom is to contribute a negative reactance to the element's impedance, Z. This negative reactance contribution increases in magnitude as the boom diameter increases.

For thicker elements, the volume of the magnetic field is reduced anyway, because the field is limited to the region outside the element. Thus, there is less field volume for the boom to remove, so the effect of thicker elements will be to reduce the magnitude of the boom's effect, hence also reducing the correction required. [See the sidebar "Plumber's Delight Meets EM Theory" for another way of looking at this.—*Ed.*]

The element-plus-boom can be restored (approximately) to its original electrical state by lengthening the element so as to contribute a positive reactance to offset the boom effect. This is the boom correction. Brian Beezley, K6STI, writes in the handbook to his Yagi design and analysis program *YO6* that elements of different diameter are electrically equivalent when the *phase angles* of the complex self-impedances are the same. This differs from simply equating the imaginary components (the reactive parts) of Z.

The Experiments

Thirteen experimental measurements were made with the boom (B) and element (d) diameters shown in Table 1, as limited by available materials. The signal source was a Yaesu

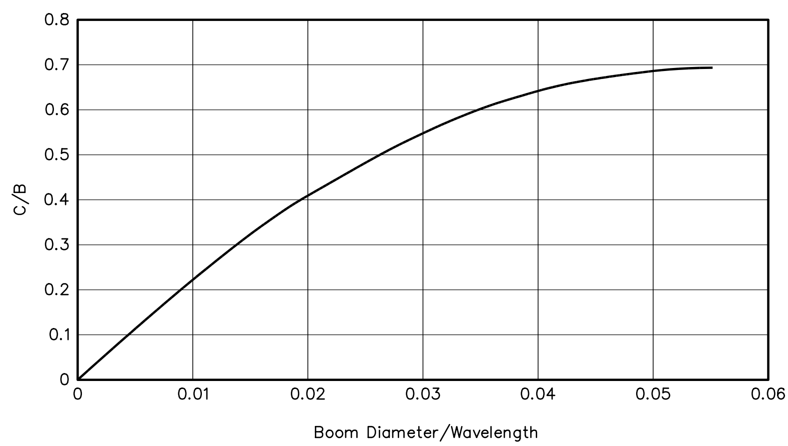


Fig 1—Plot of G3SEK's formula for boom correction.

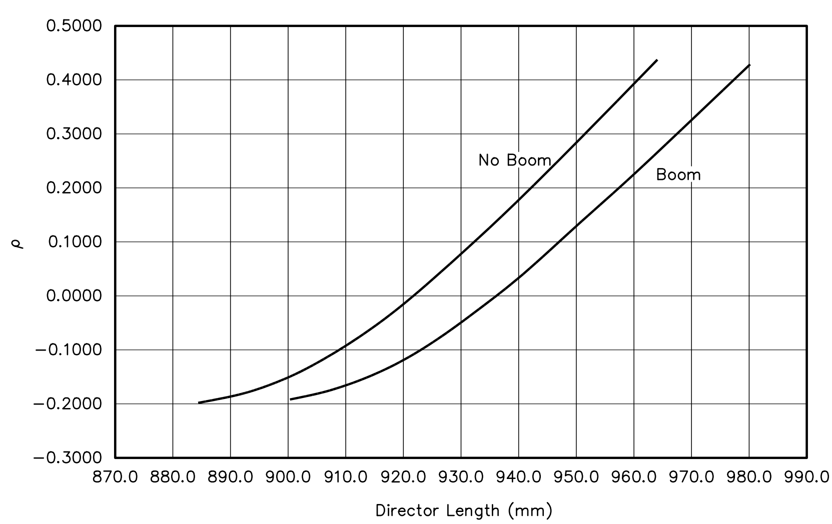


Fig 2—Data for frequency 144.2 MHz, *B* = 32.0 mm, *d* = 6.35 mm.

FT-736R transceiver, delivering 25 W on 144 MHz and 432 MHz and 10 W on 1296 MHz.

Forward and reflected powers were measured with a Bird 43 wattmeter using different plug-in elements for forward and reflected power. The relative precision for the reflected power was about 0.02 W on 144 MHz, 0.04 W on 432 MHz and 0.01 W on 1296 MHz. The absolute measurement accuracy was not nearly as good as this, but the experiments consisted essentially of comparing different antennas to obtain the same reflected power, so calibration errors are not as important as reading precision.

For each frequency and boom diameter, a simple three-element Yagi was designed using *YO6* and constructed on a dry wooden boom (usually rectangular). The feed impedance was around 25 Ω and T-matching was used with a conventional 4:1 balun. In each case, the elements were cut to the expected length; then the T-bars and the length of the driven element (DE) were adjusted for zero reflected power with no metal boom sleeve in place. The metal boom for the director (D1) was an exact sliding fit over the wooden boom and extended about half of the distance back toward the DE and a similar distance forward. Elements were pinned in place with self-tapping screws—which made no observable difference to any reading—to ensure good electrical contact between elements and boom. This arrangement guaranteed that the director could be repeatedly removed and replaced in exactly the same position.

To avoid ground effects, each antenna was mounted to radiate vertically upward. With the boom sleeve in place and the director cut deliberately long, the forward and reflected powers were recorded for each director length (*L1*), as the length was systematically reduced by small amounts until the reflected power was near zero. (Sometimes the measurements were continued well beyond this point.) The boom sleeve was then removed and the process repeated over a similar range of reflected powers. The reflection coefficient (ρ, equal to the square root of the power reflection coefficient) was plotted against *L1*. The expectation was that two parallel curves would result, their separation being the desired boom correction. In fact, the curves were not quite parallel!

Element lengths were measured with a steel ruler on 144 and 432 MHz to a precision of about 0.2 mm and with dial calipers on 1296 MHz to a precision of 0.01 mm. These two methods are not equivalent, in that the ruler measures a length averaged by eye over the end faces, whereas the calipers measure between the high points on each end face. However since all measurements in any one experiment were made consistently, the accuracy of the experimental boom correction factors found from a length difference should approach twice the appropriate precision above. The smoothness of the raw data curves supports this belief.

The Results of the Experiments

Figs 2, 3 and 4 are typical of the 13 graphs obtained for reflection coefficient ρ as a function of director length (*L1*) with and without a metal boom

Table 1—Thirteen experimental situations

144.2 MHz:	B = 32.0 mm, d = 4.76 mm, 6.35 mm
432.2 MHz:	B = 16.2 mm, d = 2.40 mm, 3.18 mm, 4.76 mm
	B = 20.2 mm, d = 2.40 mm, 3.18 mm, 4.76 mm, 6.35 mm
1296.2 MHz:	B = 16.2 mm, d = 1.60 mm, 2.40 mm, 3.18 mm, 4.76 mm.

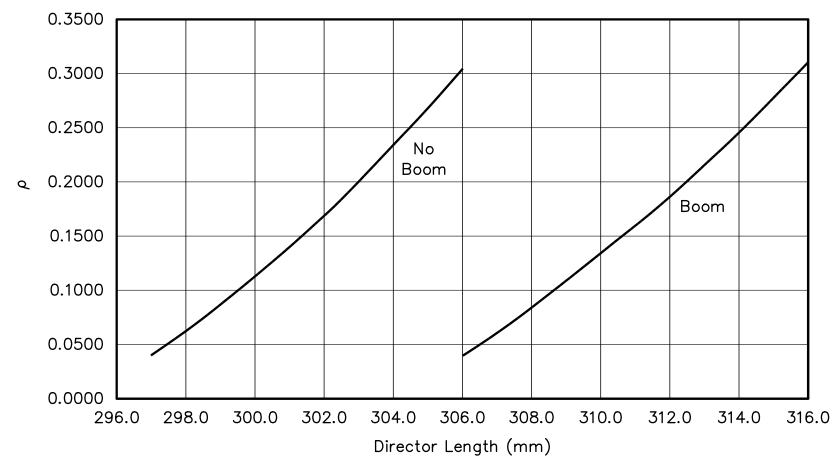


Fig 3—Data for frequency 432.2 MHz, *B* = 20.2 mm, *d* = 4.76 mm.

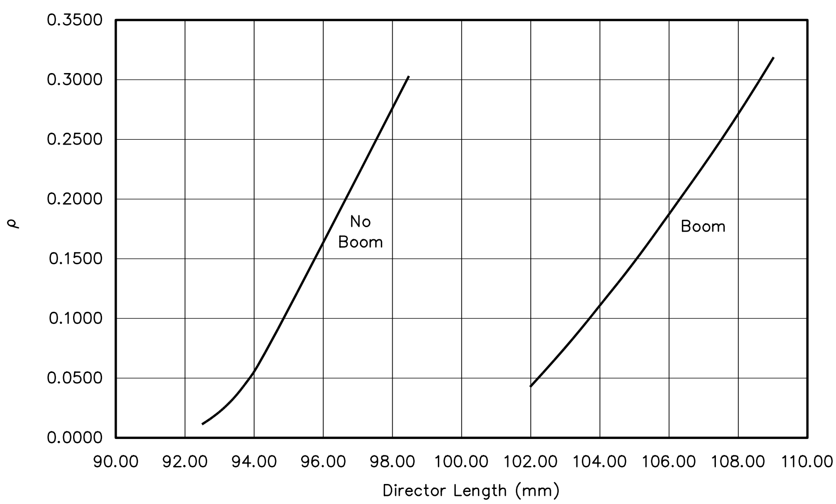


Fig 4—Data for frequency 1296.2 MHz, *B* = 16.2 mm, *d* = 4.76 mm.

sleeve. Careful study of these and other graphs shows that the boom correction, measured by the separation of the two curves, decreases slightly as the director length is reduced. This finding is not very surprising, but is significant because such a dependence has not previously been suggested.

From each graph, the director lengths with and without the boom sleeve were tabulated at several values of reflection coefficient ρ (eg, 0.1, 0.15, 0.2, 0.25 and 0.3) and a set of boom corrections was found. For each of the more than 50 pairs of director lengths, *YO6* was used to find the complex element impedance, Z. The program actually requires a reflector to be present, so this was placed 100 meters behind the driven element, where it would have no discernible effect. (The use of a particular program such as *YO6* to find element impedance is open to some criticism. This important point will be discussed below.)

The impedances from each pair of director lengths were used to find the negative reactance (X) contributed by the boom. This is best illustrated by an example. The results from Fig 3 for ρ = 0.1 are reproduced in Table 2.

The boom correction is here 9.15 mm. The X in the column for impedance with the boom sleeve present represents the unknown contribution of the boom to Z. The value of X was found by equating the phase angle (ϕ) of Z with and without the boom sleeve, so that the two situations are electrically equivalent. This gives $X = -j18.50\ \Omega$. Originally, the comparison was made by simply equating the imaginary components of Z, but this procedure led to model curves that did not converge in the way actually observed, so the phase-angle method was adopted. The values of X found in this way were reasonably consistent over the whole range of ρ and were averaged.

Finally, the value of X was used to predict boom corrections over a wider range of element lengths typical of a long Yagi by reversing the procedure. For the example in Table 2, the calculated boom corrections range from 7 mm (for the shortest director) to 12 mm for the reflector. This shows clearly the variation of boom corrections to be expected over the length of such a Yagi and the errors introduced by using a fixed boom correction for all elements. A table of such calculated boom corrections (which include the experimental values as a subset) was generated for all 13 experiments.

Each of these 13 tables of calculated corrections was plotted against director length $L1$, and they were all found to fit closely to a simple power-law relationship. The optimum value of the power varied slightly across the experiments, but a satisfactory fit for all the data was given by:

$$C = k(L1)^{1.8} \qquad \text{(Eq 4)}$$

or

$$\left(\frac{C}{C0}\right) = \left(\frac{L1}{L0}\right)^{1.8} \qquad \text{(Eq 5)}$$

where $C0$ and $L0$ correspond to some standard director length. For various reasons, this standard element length was chosen to be 0.42 λ, and the final graph presented in Fig 7 corresponds to this standard length.

Table 2 —Results taken from Fig 3 for ρ = 0.1

	No Boom Sleeve	*With Boom Sleeve*
Director Length (mm)	299.45	308.60
Impedance Z (Ω)	51.5 – j35.7	55.5 – j20.0+X

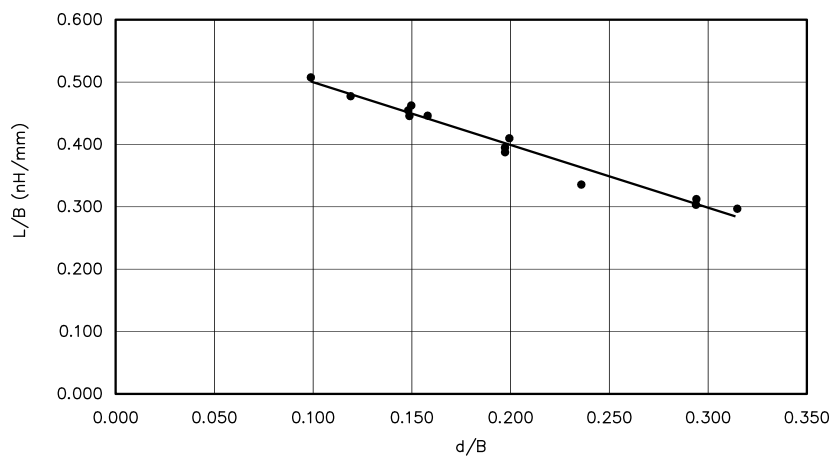


Fig 5—Dependence of inductance *L* on *B* and *d*.

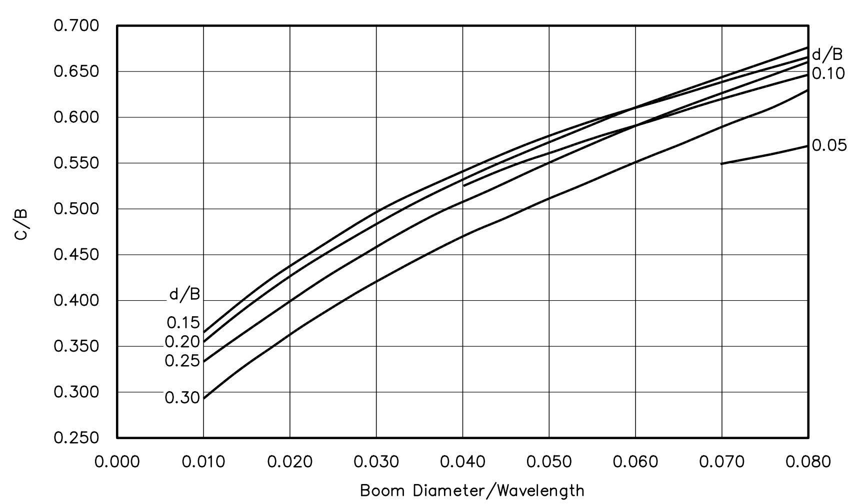


Fig 6—Boom corrections *C*/*B* for elements of length 0.42 λ and for various *d*/*B* values.

Effect of Boom and Element Diameters

In spite of trying many different plots, it has not proved possible to represent the dependence of the length correction (C) on element diameter (d) in any simple way. This is not entirely surprising because of the complexity of the effect on element impedance of varying the tip length to compensate for the effect of the boom. It has proved very helpful to break the problem into two separate parts:

1. *The effect of the boom on the element impedance:* As explained above, this can be represented as a pure negative reactance, the value of which depends also to a lesser extent on the element diameter.

2. *The increase in length required to compensate for this reactance:* so as to restore the original phase to the element impedance.

The boom reactances from the 13 different experiments have been converted to inductance (L) and plotted in Fig 5 as L/B versus d/B. The inductance is plotted as a positive quantity for convenience, but remember that it contributes negatively to Z.

The graph of L/B versus d/B shows a remarkable linear relationship:

$$\frac{L}{B} = 0.5994 - 0.999\left(\frac{d}{B}\right) \quad \text{(Eq 6)}$$

Only one point (at 432.2 MHz) departs appreciably from the line of best fit. Having due regard to the accuracy of the data, this relationship is most simply expressed as:

$$L = 0.6B - 1.0d \quad \text{(Eq 7)}$$

In this simple and elegant expression (Guy's Rule!), L is the value of the negative reactance contributed by the boom/element combination to the element impedance Z. For the values of the constants as presented, L is in nanohenries, while B and d are in millimeters. With this rule, the reactance of *any boom and element combination* can be predicted with reasonable confidence.

Calculation of Boom Corrections

It is straightforward—but not particularly convenient—to use the inductance value given by my rule to calculate a value for the boom

Seven Simple Steps

1. Calculate the wavelength (in millimeters) from:

$$\lambda = \frac{299792.5}{f} \quad \text{(Eq 8)}$$

where f is in megahertz: Eg, f = 1296.2 MHz, λ = 231.3 mm.

2. Choose a boom diameter B and element diameter d, both in millimeters. Eg, B = 16.2 mm, d = 3.18 mm.
3. Calculate the ratios B/λ and d/B. Eg, B/λ = 0.070, d/B = 0.196.
4. Refer to Fig 7. Draw a vertical line corresponding to the value of d/B and read off the value of C/B from the appropriate curve. Interpolate between the curves as necessary. Eg, C/B = 0.645.
5. Calculate C (in mm) from C/B by multiplying by B. This is the boom correction for an element of length equal to the standard length, $L0$. Eg, $C0$ = 10.4 mm.
6. Calculate the standard length $L0$ from $L0$ = 0.42 λ. Eg, $L0$ = 97.1 mm.
7. Calculate the correction C for any element of length L from Eq 5. Eg, for L = 90.0 mm, C = 9.1 mm.

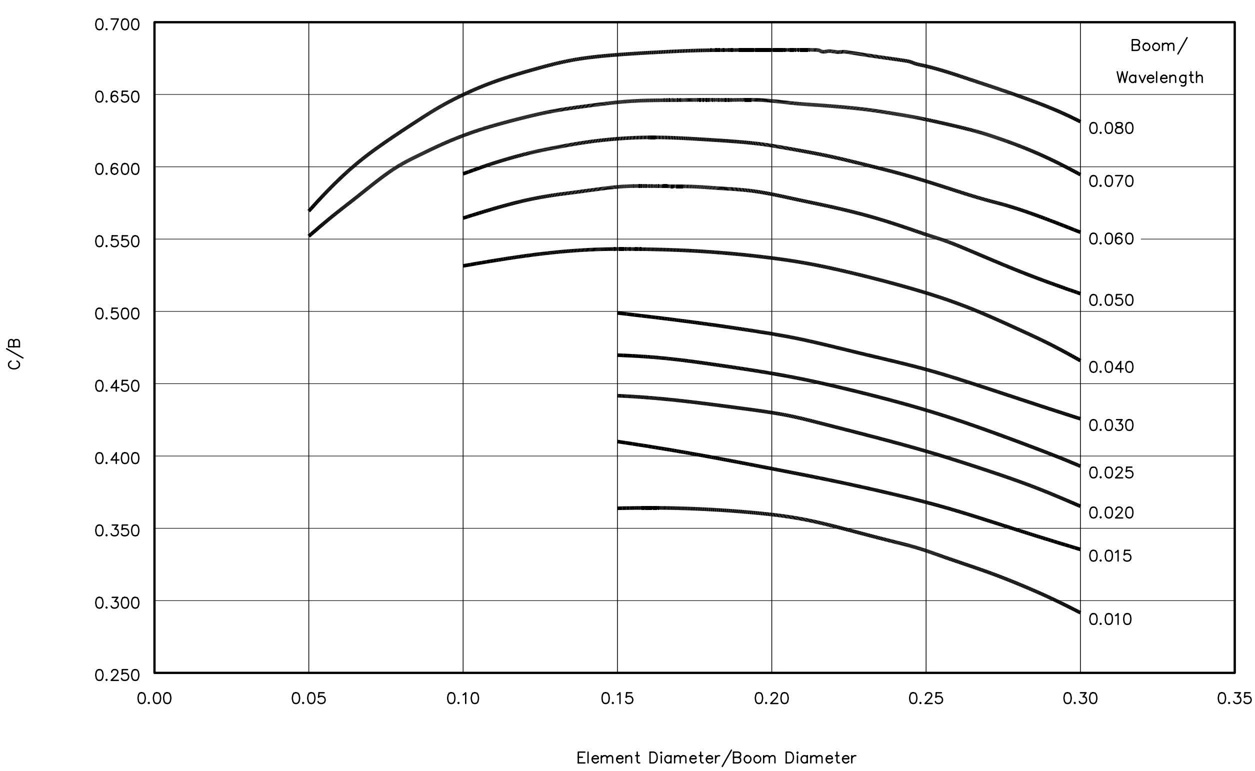


Fig 7—Boom corrections *C/B* for elements of length 0.42 λ. *B/λ* marked for each curve.

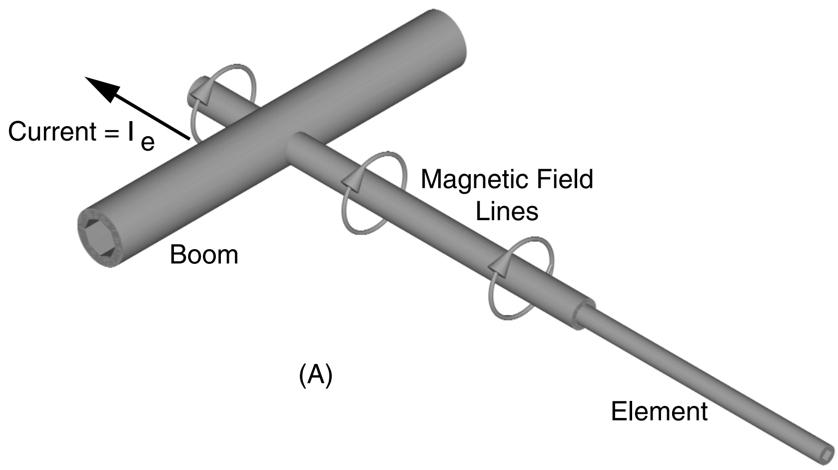
Plumber's Delight Meets EM Theory

Guy's theory about why Yagi elements appear electrically shorter when attached to a conductive boom proved difficult for me to understand at first. The following represents my perception of the effect as clarified by him. I hope it helps you, too.

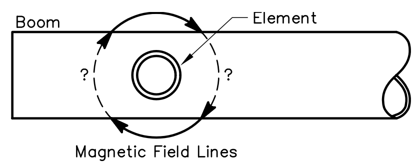
Current flowing in a conductor such as a Yagi element produces a magnetic field aligned everywhere at right angles to the direction of current flow. The shape of magnetic field lines satisfying this condition is a circle concentric to the element, as shown in Fig A.

The presence of a conductive boom interferes with the development of the magnetic field near the boom, thereby affecting the current flow there. See Fig B. Experiments verify the electrical shortening of the element. Using knowledge of how EM fields interact with matter, it seems to me this theory could be developed to predict the exact magnitude of the effect. What properties of the boom material are included in the solution? What geometries minimize the effect?

I do not find much mention or analysis of this idea in the literature. It appears to merit further consideration in modeling of antennas, especially at UHF and microwave frequencies. What do *you* think?—*Doug Smith, KF6DX.*



(A)



(B) Side View

correction C for any combination of boom diameter, element diameter and element length. This involves using *YO6*, first to find the complex impedance Z of the uncorrected element length and hence its phase ϕ. Then by a process of trial and error to find a new element length which, when combined with the negative reactance contributed by the boom, has the same phase.

Instead, for the standard element length of 0.42 λ, boom corrections have been calculated over a wide range of boom diameters (B) and element diameters (d) covering all the sizes likely to be met in practice. The results may be plotted in the form C/B versus B/λ, as in Fig 6, with separate curves for different element diameters. Alternatively, C/B may be plotted versus d/B, as in Fig 7, with separate curves for different boom diameters. Other possibilities include using d/λ in place of d/B.

Fig 6 may be compared directly with Fig 1, based on the G3SEK formula. The curve shapes in Fig 6 are generally similar to that in Fig 1, but it is apparent that the intercepts on the vertical axis of Fig 6 are well above zero for all values of d/B. The curves in Fig 6 also intersect, making it hard to use in practice. The reason for these intersecting curves is clearer in Fig 7. In general, as the element diameter increases, the boom correction factor C/B decreases as expected from the discussion earlier. In the case of thick booms however, the boom correction factor also falls for very thin elements, for which the reactive component is large. This is, in fact, a consequence of the use of a standard element of fixed length rather than fixed phase. The use of a standard length is much easier to use in practice, but leads to curves that intersect when boom diameter is used as the horizontal coordinate.

For the practical prediction of boom corrections, Fig 7 is significantly easier to use than Fig 6 because the various curves are well separated and generally less inclined.

Practical Significance of Length-Dependent Boom Corrections

The detailed results described in this article are novel in that they lead to boom correction factors that depend not only on the boom diameter, but also on element diameter and length. It is reasonable to wonder whether this has any real practical significance when compared with the simpler system of a fixed correction factor presently in widespread use. If the corrections should indeed taper from larger values for longer directors and the reflector to smaller values for the shorter directors, then the effect of using a single, fixed correction is to apply a correction that is too small for the longer elements and too big for the shorter ones.

This can be easily simulated in an antenna analysis program such as *YO6* by adding the fixed correction to every element and then subtracting the tapered corrections. Such simulations lead to the conclusion that—at 144 MHz—the difference between the two approaches is negligible. This is not at all surprising since the corrections are a small fraction of the element lengths. At 432 MHz, small differences are apparent, but do not appear very significant. At 1296 MHz, however, the fixed and tapered corrections differ by considerably more than acceptable construction tolerances. The predicted antenna properties also differ significantly, with some loss of gain when a fixed correction is used and major differences in the feed impedance. This is consistent with the matching difficulties previously experienced at 1296 MHz by some amateurs.

Several local amateurs have now constructed long Yagis for 1296 MHz using the VK2KU tapered corrections, and in each case, they have reported that matching the Yagi proved quite straightforward.

Conclusions

The raw data graphs—such as Figs 2, 3, and 4—appear to show unequivocally that boom corrections depend not only on the boom diameter (as a fraction of wavelength), but also on element diameter and element length. The dependence on element diameter may not be very startling, but the dependence on element length appears to be a novel idea that was initially unexpected.

The formula for calculating the negative reactance contributed by the boom/element combination is also new, but it fits the experimental data very well. This rule is the key to calculating boom corrections for any combination of boom and element diameter. It may well be that the use of a different computer program for finding element impedance would lead to somewhat different values for the negative reactance, and so to slightly different constants in the formula. When the procedure is reversed, however, and the same program is used to find the corrections in other situations, such differences between programs should largely be eliminated. In effect, the computer modeling is used to interpolate between boom correction factors that were found directly by experiment. Thus, I believe that the graphical results as presented in Fig 7 are substantially independent of the computer modeling and represent a close approximation to the truth.

The length dependence of the corrections appears to be best described by a power law of order 1.8, though this value does not seem to be very critical. The fixed boom corrections commonly used elsewhere extend up to a boom diameter of 0.055 λ. The experiments described in this article extend this range up to 0.070 λ and calculations have been carried out up to 0.080 λ, thus covering the important range of booms thicker than 12 mm at 1296 MHz.

Acknowledgements

I particularly thank Gordon McDonald, VK2ZAB, and Ian White, G3SEK, for their encouragement and advice in a project that threatened to get out of hand, growing rapidly from a planned single measurement at 1296 MHz into a comprehensive survey over three frequency bands, three boom diameters and five element diameters.

Guy Fletcher was a senior lecturer (associate professor) in Physics (electromagnetism and optics) at Macquarie University in Sydney until his retirement in 1997. He holds a masters degree from Cambridge, UK, and a doctorate from Macquarie.

Guy's previous call signs are G3LNX (UK 1957-1967) and VK2BBF (1967-1998). His amateur interests include VHF/UHF DX on 144, 432 and 1296 MHz, mainly SSB, some CW. He also studies propagation modes on those frequencies, including tropo', sporadic E, ducting and aircraft scatter. He is also active in antenna design and construction.

A Scanner Controller for Varactor-Tuned Receivers

Forget about optical encoders—here's a controller board that provides an easy, push-button frequency control and scanning interface for varactor-tuned receivers.

By Thomas K. Duncan, KG4CUY

Microcontrollers appear in many construction projects in current Amateur Radio literature, but their usefulness is not limited to newly constructed equipment. In this project, which served as my introduction to embedded microcontrollers in radios, I grafted a scanner controller into a single-conversion, varactor-tuned receiver.

With increasing activity on 10 meters, I wanted a radio to periodically monitor certain haunts for upper-sideband voice activity. My store-bought scanner that covers these frequencies handles only AM and FM. Besides, I was looking for an excuse to get my feet wet with simple microcontrollers. A 10-meter receiver built in the early months of increasing solar activity kept whispering "Use me!" every time it was switched on. I wanted to maintain that radio in working order in case this project was a complete failure. The original design —in particular, the board layout—was therefore modified slightly to add the necessary hooks for the controller. These modifications are quite simple; it should be possible to add such a controller to any voltage-tuned receiver.

The Microchip PIC product line was chosen for ease of programming: A programmer could be bought or built at reasonable price. Development software was widely available at little or no cost, and certain chips are FLASH programmable, so no EPROM eraser was needed. A 16-character LCD with an on-board controller and a set of five momentary-contact push-buttons (**FREQUENCY UP**, **FREQUENCY DOWN**, **SCAN**, **SELECT CHANNEL** and **PROGRAM CHANNEL**) comprise the new tuning mechanism. Recent *QST* articles[1, 2, 3] aimed me toward these design decisions.

Description

As the push-button functions imply, the unit operates as a minimal-function scanner:

- Manual tuning takes place via the frequency up/down buttons. Tuning speed is increased by a factor of eight each time the **SELECT CHANNEL** button is pressed while holding **FREQUENCY UP** or **FREQUENCY DOWN**. The display tracks frequency while tuning.
- The current frequency may be stored into the current-channel EEPROM space by pressing the **PROGRAM CHANNEL** button.
- The current channel is incremented by pressing **SELECT CHANNEL**, thereby tuning the receiver to the value currently stored for that channel. There are 10 channels.
- To scan continuously through all channels, press **SCAN**. Scanning pauses when the received-signal line goes high and stops if it remains high for five seconds.

In addition, a calibration mode may be entered by pressing **FREQUENCY UP** and **FREQUENCY DOWN** simultaneously. In this mode, the frequency counter may be adjusted to accommodate the exact frequency of the ceramic resonator or crystal used in the clock oscillator. Similarly, the IF-offset mode allows the IF to be set so that the display shows the received frequency rather than the LO frequency, which the counter actually measures.

Some microcontrollers have built-in D/A converters that might generate tuning voltages. They are generally limited to 8-10 bits of resolution. It was my desire to cover 28.0 to 29.2 MHz at 100 Hz or better resolution, thus

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[1]Notes appear on page 30.

requiring at least (1,200,000/100) = 12,000 steps, or 14 bits. Because I wanted to use FLASH-programmable devices, on which I/O pins were scarce until recently,[4] a serial interface to any external D/A converter was desirable. Serial D/As with the required resolution were rare in DIP packages, so I chose an Analog Devices AD8402 dual RDAC, or "digital potentiometer," to feed an op amp that develops the tuning voltage. Each RDAC section has 8-bits of resolution, multiplied in the op amp to provide the desired tuning granularity. A CMOS rail-to-rail op amp, the National LMC662CN, is used to provide adequate tuning-voltage swing.

With outputs driving the LCD and RDAC, and inputs from push-buttons, a frequency counter and received-signal line, the 13 I/O pins of one 16F84 would not suffice. Two 16F84s are therefore used, communicating with each other via two lines: One line is a gate pulse for the frequency counter, and the other carries unidirectional serial I/O. One PIC services the display and provides the gate pulse. The other measures frequency, detects the received-signal input and handles the push-button inputs.

A block diagram of the controller appears in Fig 1. While most of the important work takes place in PIC1, the interface to the Optrex LCD module is not entirely trivial. It takes much code to convert the LO frequency from the rather obscure units sent over by PIC1 into displayable digits at the input frequency. This division of labor also allows PIC2 to serve as an additional timer used to gate PIC1's frequency-counter software during the 16-ms counting periods (each 16F84 has one timer).

Modifications to the Original Receiver

While the original radio already had a LO buffer amplifier to drive another PIC design (one of Almost All Digital Electronics' DFD-series displays),[5] another single-stage FET buffer was added

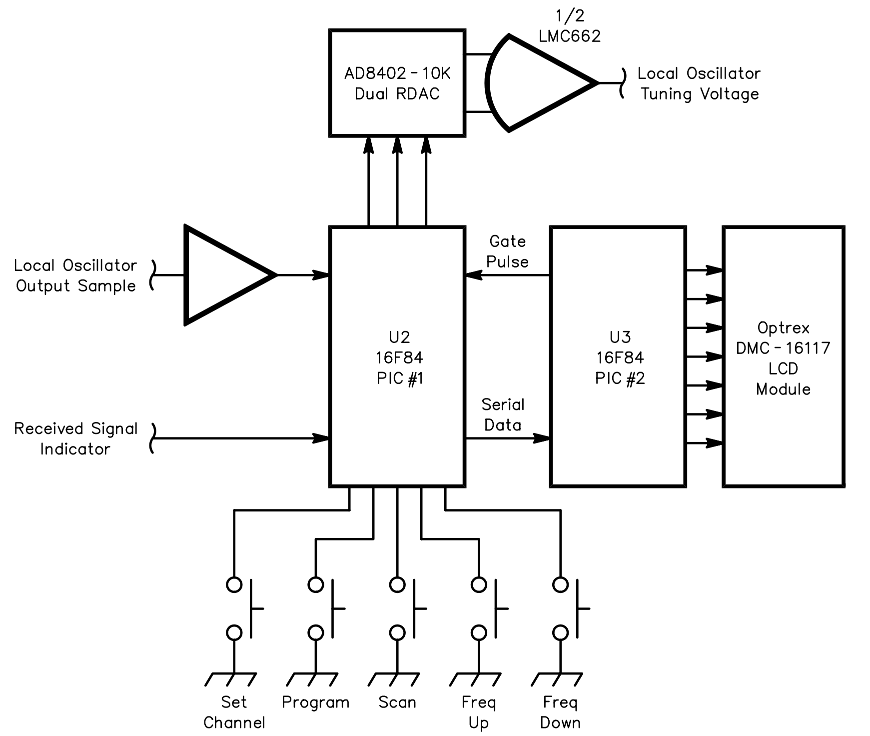


Fig 1—Controller block diagram.

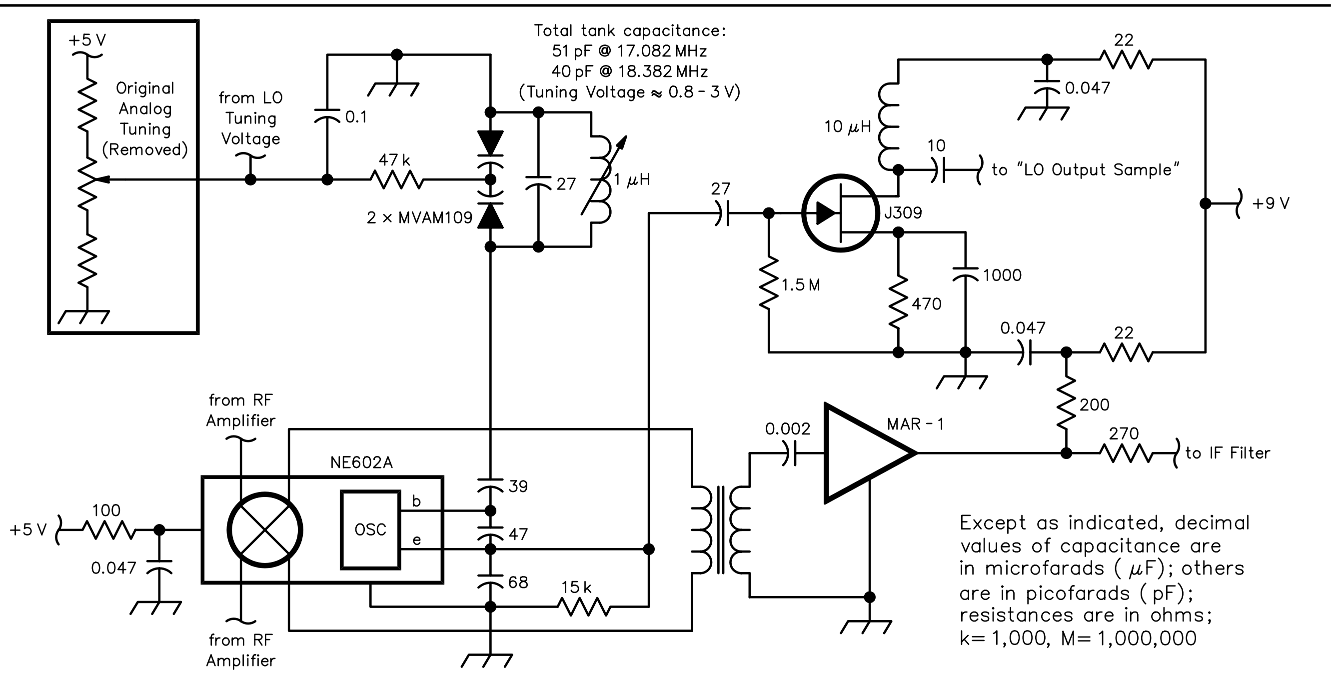


Fig 2—Receiver modifications.

at the controller end since the radio and controller would be on two separate boards, connected by a length of coax. The schematic in Fig 2 shows the tuning voltage and LO-sampling connections.

The received-signal input required new circuitry. The original radio didn't have AGC, so the spare section of an audio preamplifier, LM1458, and an FET were used to generate about +4 V in the presence of audio and less than +1.5 V with no audio. In one of its early versions, the controller supported only manual recall of stored frequencies, so the received-signal input wasn't required. The control signal is only needed to stop scanning when a received signal is found. The schematic of Fig 3 shows the received-signal circuitry added to the original radio.

Controller Hardware Details

Fig 4 is the complete controller schematic. The PIC has internal pull-up resistors on the five push-button ports, so the push buttons need only short these pins to ground. An 8-MHz ceramic resonator is used as the clock for each PIC. Inexpensive 8-MHz crystals could be used as well, but they require capacitors between each end of the crystal and ground. The 39-kΩ/0.47-µF network on each MCLR input guards the PICs against false starts when power is applied. In retrospect, these pins could probably have been connected directly to +5 V without violating the power rise-time rules. These and all other 16F84 details are available in documentation available on Microchip's Web site.

The 32 Ω trimpot R2 in the tuning-voltage circuit is used to adjust for monotonically decreasing tuning voltage with increasing RDAC setting. Tuning voltage is adjusted in two parts: One section of the RDAC drives the non-inverting op-amp input, and the other part drives the inverting input. The pot adjusts the range of the least-significant RDAC driving the inverting input so that the tuning direction does not reverse at either the low or high ends. This adjustment is most easily made by connecting a frequency counter with at least 10-Hz resolution to the controller's LO input and pressing the **FREQUENCY UP** or **FREQUENCY DOWN** button momentarily several hundred times. If when tuning the counter momentarily reverses direction (while pressing **FREQUENCY UP** the counter goes down), adjust the trimpot and try the offending range of frequencies again. If the button can be pressed at least 256 times (ensuring the complete 8-bit range of the least significant RDAC has been traversed) and no frequency reversal occurs, the adjustment is correct.

Controller Software

Microchip provides PIC software-development tools at no charge on its Web site, **www.microchip.com**. The MPLAB package is an integrated development environment that runs under *Windows*: I used MPLAB 3.99 under *Windows NT* 4.0. While it is possible to program PICs in *C* and *BASIC*, I chose to stay very close to the hardware by using assembly language. The instruction set is simple and straightforward, and Microchip supplies many support routines written in the assembly languages of the various PIC products.

The two PICs comprising the controller perform the major tasks shown in Table 1. Some of the tasks deserve special explanation.

Measuring Frequency

The LO frequency applied to PIC 1 feeds a prescaler that divides the input frequency by 16. This, in turn, drives an eight-bit counter which interrupts PIC 1 when it rolls over from a full count of $FF to $00 [a preceding "$" indicates a hexadecimal number—*Ed.*]. When PIC 1 needs to measure frequency, it asks PIC 2 for a gate pulse, which is 16 ms long. When the gate pulse rises from its quiescent low value to high, PIC 1 enables the timer interrupt, clears the timer, and waits for the gate pulse to go back to low. During this wait loop, timer interrupts occur. On every interrupt, PIC 1 increases the 16-bit frequency count by 256 by incrementing the high-order byte by one. When the gate pulse falls to low, PIC 1 disables interrupts and moves the current counter

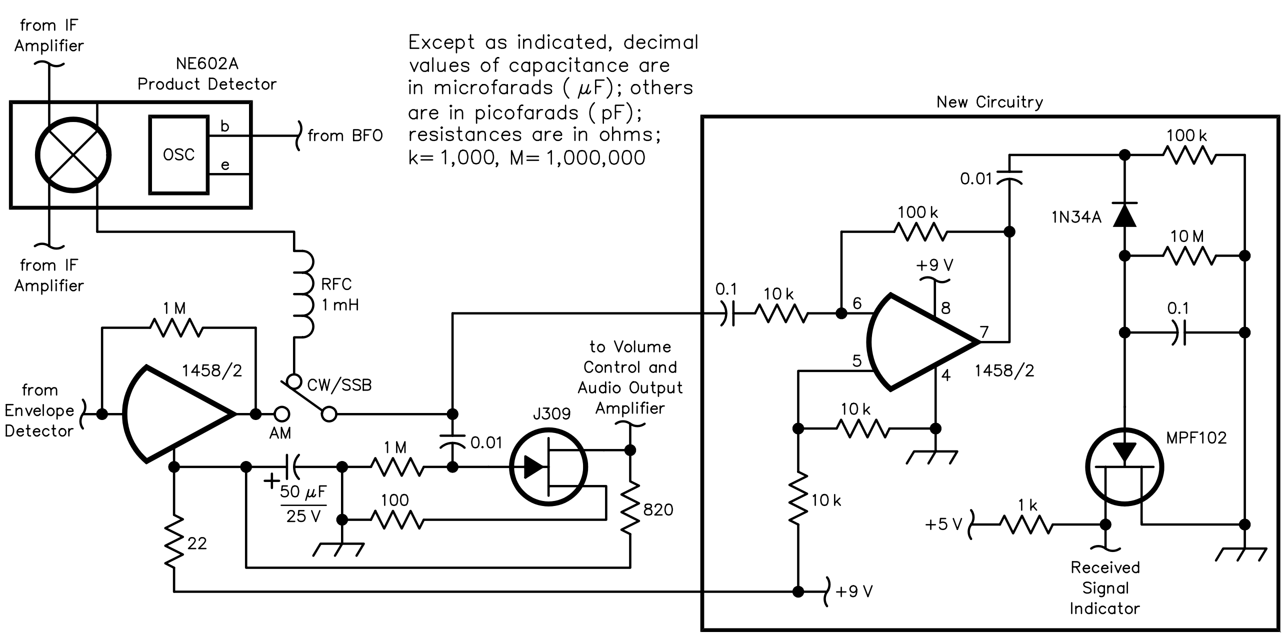


Fig 3—Received-signal indicator circuit.

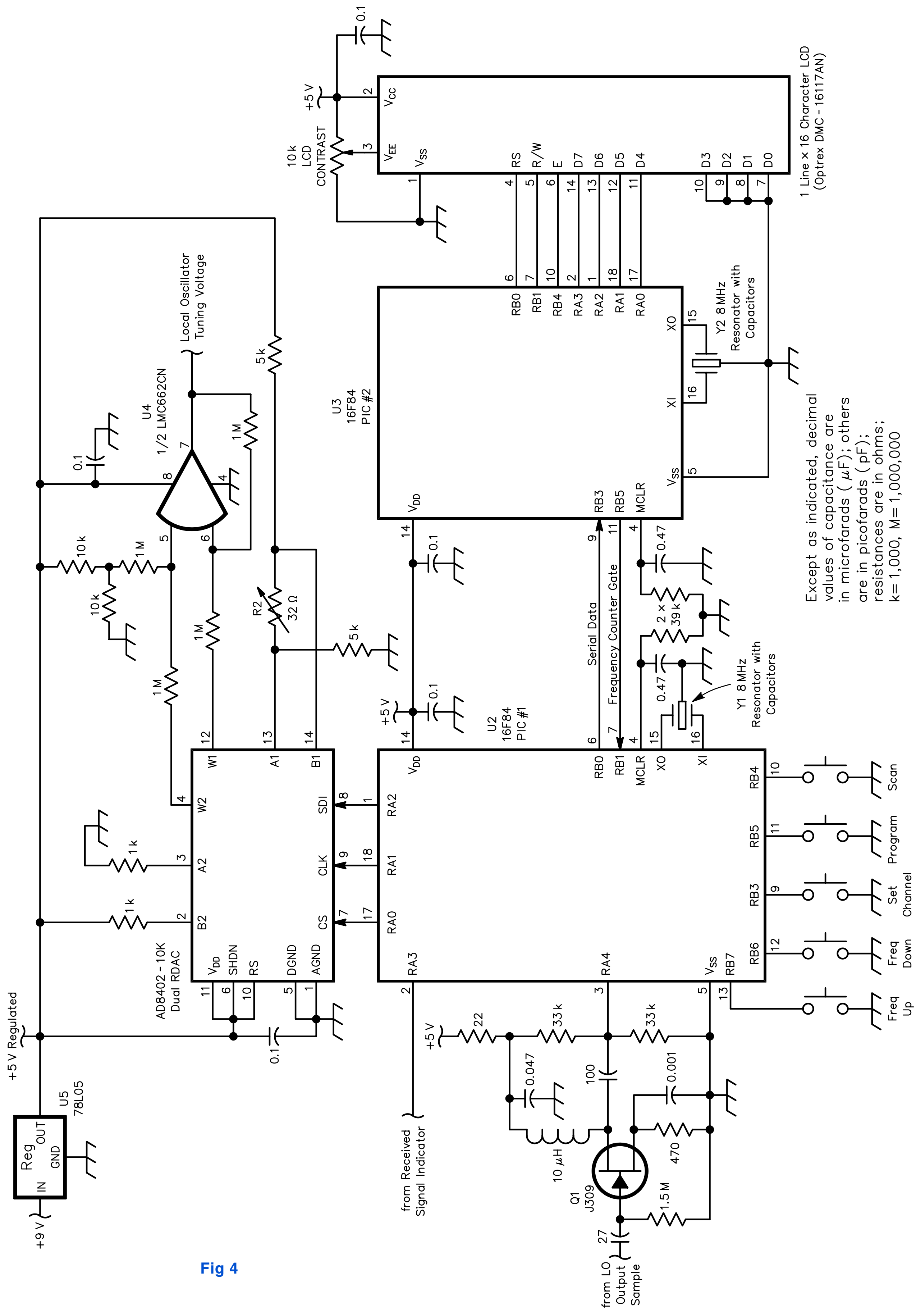

1 Line × 16 Character LCD
(Optrex DMC-16117AN)
10 k
LCD
CONTRAST
+5 V
U3
16F84
PIC #2
U2
16F84
PIC #1
Y2 8 MHz
Resonator with
Capacitors
Y1 8 MHz
Resonator with
Capacitors
Serial Data
Frequency Counter Gate
U4
1/2 LMC662CN
Local Oscillator
Tuning Voltage
AD8402-10K
Dual RDAC
R2
32 Ω
+5 V Regulated
U5
78L05
Reg
IN
GND
OUT
+9 V
from Received
Signal Indicator
from LO
Output
Sample
Q1
J309
10 µH
Scan
Program
Set
Channel
Freq
Down
Freq
Up
Except as indicated, decimal
values of capacitance are
in microfarads (µF); others
are in picofarads (pF);
resistances are in ohms;
k= 1,000, M= 1,000,000

Fig 4

value—equal to the number of prescaler counts since the last interrupt—into the low-order byte of the 16-bit frequency count.

The 16-bit frequency count, C_{LO}, therefore represents the input frequency divided by the prescaler setting (16) and by the quantity (1 s) / (16 ms) = 62.5, for a net division by 1000, which represents the LO frequency in kilohertz. This number is sent to PIC 2.

Converting LO Frequency to Tuned Frequency

Depending on the number of CPU cycles in the gate-pulse timing routine and PIC 2's clock frequency, the gate pulse may not be exactly 16 ms long, so an adjustment is provided. PIC 2 calculates the tuned frequency as:

$$F_{tuned} = F_{IF} + \frac{(C_{LO})(GPM)}{10^4} \qquad \text{(Eq 1)}$$

Where GPM is the gate pulse multiplier, around 10,000. Its value is stored in the PIC 2 EEPROM, and it may be adjusted by commands from PIC 1 to PIC 2. Similarly, the additive term F_{IF} accounts for the intermediate frequency. It is stored in the PIC 2 EEPROM and is adjustable.

There are no multiply and divide instructions in the PIC instruction set, but Microchip provides routines on their Web side for both fixed- and floating-point operations. Fixed-point routines[6] are used in the frequency calculation, with the multiplication done before division to provide the greatest precision.

Setting Frequency

PIC 1 stores a channel's C_{LO} in EEPROM when the **PROGRAM CHANNEL** button is pressed. The RDAC setting corresponding to this frequency is not stored. Further, the relationship between RDAC setting and frequency depends on oscillator stability. In order to set the LO to a particular frequency, PIC 1 does a binary search across the LO tuning range. Beginning with the extreme RDAC settings ($0000 to $FFFF), the frequency at each extreme is measured. The RDAC setting at the midpoint is calculated, and the frequency measured at that point. If the desired frequency lies at the midpoint, tuning is complete. If the desired frequency lies above the midpoint, the low extreme is set to the midpoint; otherwise, the high

Fig 4 (See left)—Complete controller schematic diagram. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. Parenthetical references show (part manufacturer, source and source part number). Equivalent parts may be substituted for those shown.
DS1—16×1 character LCD module (Optrex, Digikey #DMC-16117AN)
Q1—J309 FET
U1—AD8402-10 dual 10-kΩ RDAC (Analog Devices, Allied #AD8402AN10)
U2, U3—PIC 16F84, 10 MHz (Microchip, Digikey #PIC16F84-10/P-ND). PICs must be programmed to function.
U4—LMC662 dual CMOS rail-to-rail op amp (National, Allied #LMC662CN)
U5—78L05 +5-V regulator
X1, X2—8 MHz ceramic resonator with capacitors (Murata, Allied #CST8.00MTW

Table 1—PIC Tasks

PIC 1 Tasks

1. Measure LO frequency by asking PIC 2 for a gate pulse and counting-timer overflows driven by the LO signal while the gate pulse is high
2. When the **SELECT CHANNEL** button is pressed, increment the channel, retrieve its frequency from EEPROM, set the LO to that frequency and send the channel number and frequency to PIC 2 for display
3. Respond to the **PROGRAM CHANNEL** button by storing the current frequency in the currently selected channel's EEPROM slot
4. Respond to the **FREQUENCY UP** and **FREQUENCY DOWN** buttons by adjusting RDAC settings accordingly
5. When the **SCAN** button is pressed, continuously step through channels 0 through 9, pausing five seconds on each channel until the received-signal line is active more than momentarily or any button is pressed, then stop scanning
6. Send gate-pulse-multiplier calibration messages (increment multiplier, decrement multiplier and store multiplier) to PIC 2 when in the gate-pulse-calibration mode
7. Send IF-offset-calibration messages (increment IF, decrement IF and store IF) to PIC 2 when in the IF-offset-calibration mode.

PIC 2 Tasks

1. Generate frequency-counter gate pulses;
2. Convert PIC 1 LO counts to tuned frequency and display it on the LCD;
3. Display the current channel number;
4. Adjust and store the gate-pulse multiplier in response to commands from PIC 1;
5. Adjust and store the IF offset in response to commands from PIC 1.

Table 2—Serial Command Structure

Command	*Command Byte*	*# Argument Bytes*	*Example*
Display raw decimal number	N	2	Nxx, where xx is a 16-bit binary number.
Convert LO counts to tuned frequency and display	F	2	Fxx, where xx is a 16-bit binary number.
Convert LO counts to LO frequency and display	L	2	Lxx, where xx is a 16-bit binary number.
Display channel number	C	1	Cx, where x is an 8-bit binary number between 0 and 9.
Send gate pulse	P	0	P
Blank display	B	0	B
Display gate pulse multiplier value	S	0	S
Increment gate pulse multiplier value by 1	G	0	G
Decrement gate pulse multiplier value by 1	g	0	g
Rewrite gate pulse multiplier to EEPROM	W	0	W
Display text string	T	Null-terminated	eg, THELLO<0> displays HELLO

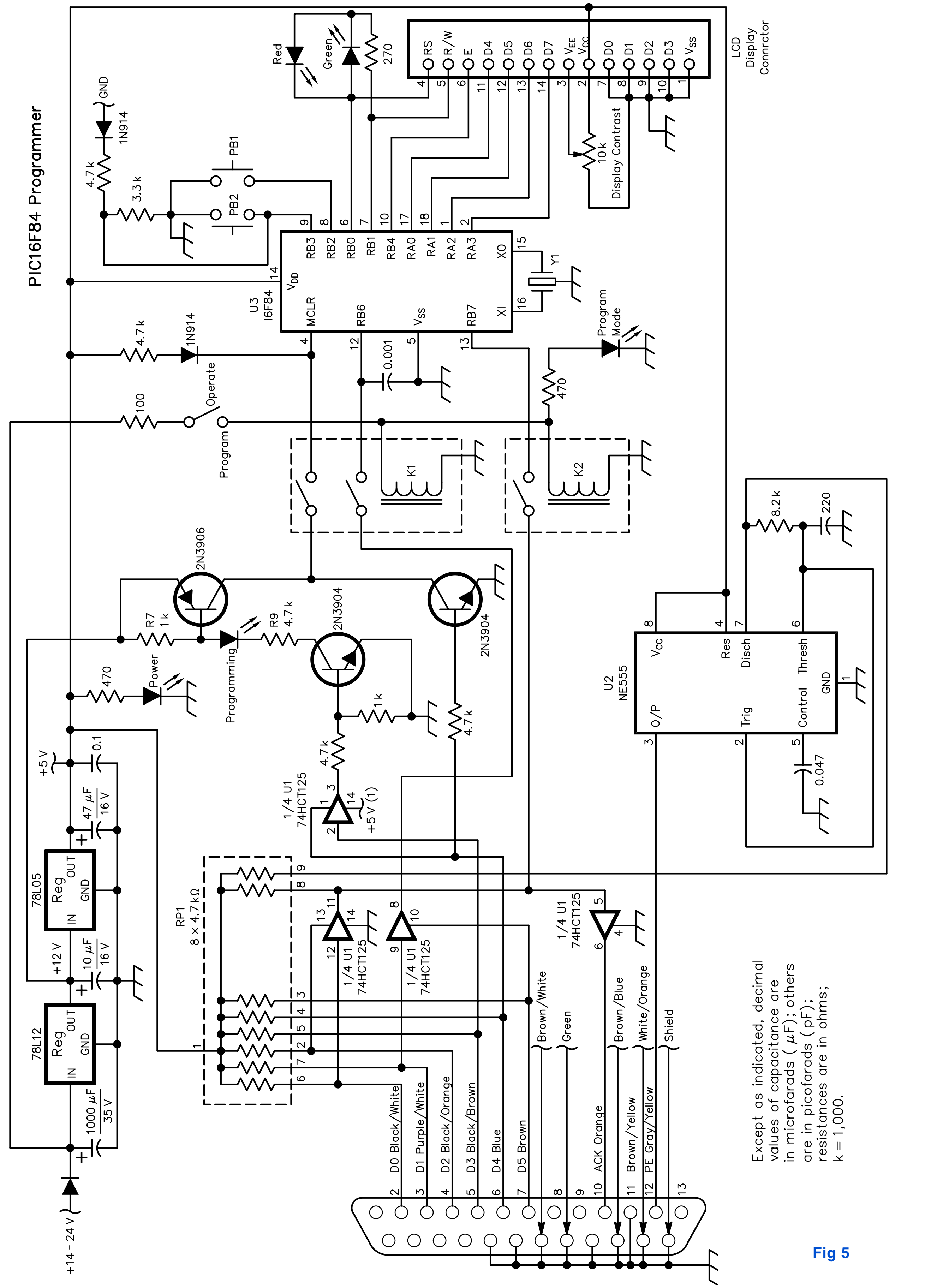

PIC16F84 Programmer
U3
16F84
U2
NE555
LCD
Display
Connrctor
Display Contrast
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.

Fig 5

extreme is set to the midpoint. This operation converges on the desired frequency provided tuning is monotonic with respect to RDAC setting as detailed above.

Serial Protocol

Serial I/O routines are adapted from

Fig 5 (See left)—PIC programmer schematic diagram. Unless otherwise specified, use ¼ W, 5%-tolerance carbon composition or film resistors. Equivalent parts may be substituted for those shown.
K1, K2—5 V DPDT relay, 5-V coil
Y1—4-MHz ceramic resonator with capacitors

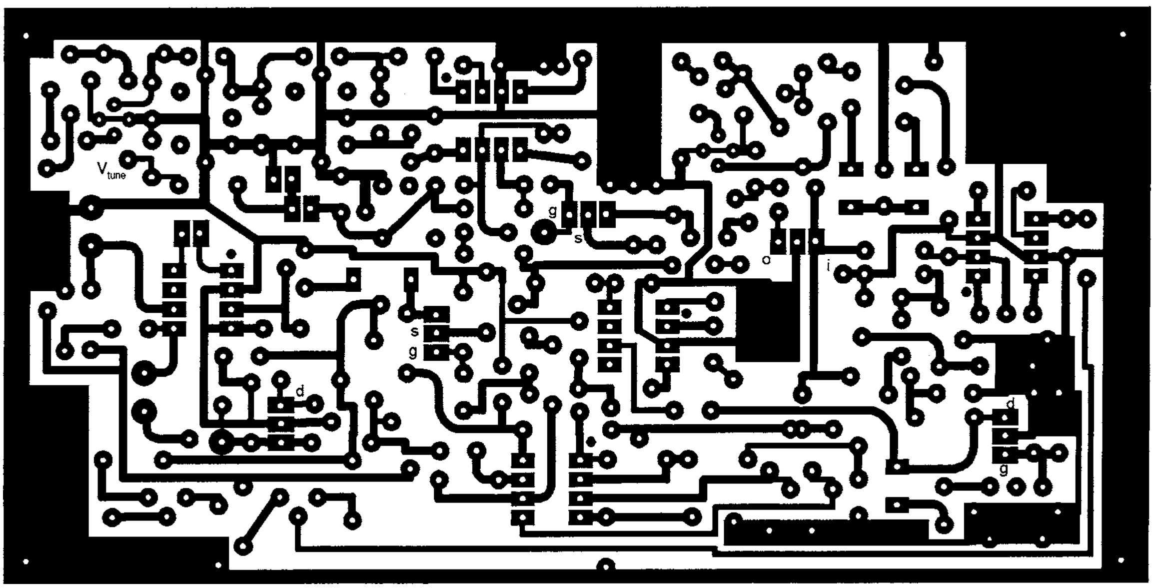
a Microchip application note.[7] Communication is unidirectional, with PIC 1 issuing commands. Except for the send-gate-pulse command, PIC 2 does not indicate completion of a command, so PIC 1 must allow sufficient time for the command to complete. Table 2 describes the various serial commands.

PIC Programming

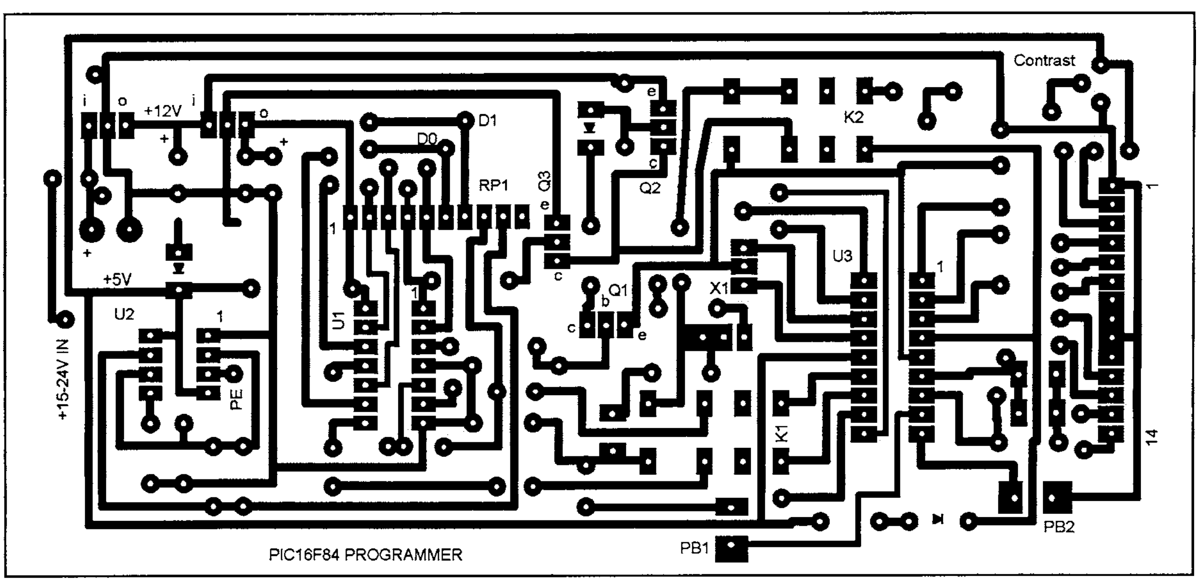
MPLAB produces the Intel-format object code files necessary to program the PICs, but a programmer is required. Programmers are available at reasonable cost from various companies, but they are also easily built. I chose to modify a programmer circuit given on the University of Nottingham's Microchip Web site maintained by Steve Marchant.[8] This site covers the programmer hardware (PROG84) and several software packages to drive the programmer. I chose Massimo Veneziano's *PPF84*, which runs under *DOS* and *Windows 95*. The schematic and PC-board layout for my variant of Steve's programmer are shown in Figs 5 and 6. Images of the etching patterns are available on the *QEX* Web site.[9]

Operation

The controller enters its normal



(A)



(B)

Fig 6—(A) PIC controller etching pattern, full size 3.87×6.07 inches. (B) PIC programmer etching pattern, full size 2.92×6.21 inches.

operating mode on power-up. Tuning proceeds very slowly, one RDAC count at a time, when either **FREQUENCY UP** or **FREQUENCY DOWN** is pressed. To increase the tuning speed (ie, RDAC increment and decrement), momentarily press **SCAN** while holding **FREQUENCY UP** or **FREQUENCY DOWN** in place. Tuning speed is indicated by a single digit on the LCD; it appears only while tuning. Table 3 shows the operating modes of the controller, the information displayed and actions taken for commands given in each mode.

Table 3—Operating Displays and Actions

Operating Mode (Display Tuned Frequency and Channel Number)

Buttons Pressed	*Action*
FREQUENCY UP	Increment frequency
FREQUENCY DOWN	Decrement frequency
SELECT CHANNEL	Go to the next channel and tune to its stored frequency.
PROGRAM CHANNEL	Store the tuned frequency for the current channel in EEPROM
SCAN	Scan through all channels
FREQUENCY UP + FREQUENCY DOWN	Enter gate pulse calibration mode
FREQUENCY UP + SCAN	Increase frequency increment
FREQUENCY DOWN + SCAN	Increase frequency decrement

Gate-Pulse-Calibration Mode Functions (Display Gate-Pulse Multiplier and LO Frequency)

Buttons Pressed	*Action*
FREQUENCY UP	Increment GPM
FREQUENCY DOWN	Decrement GPM
SELECT CHANNEL	Enter IF Offset Calibration Mode
PROGRAM CHANNEL	Store GPM in EEPROM
SCAN	Return to operating mode

IF Offset Calibration Mode Functions (Display IF and LO Frequencies)

Buttons Pressed	*Action*
FREQUENCY UP	Increment IF Offset
FREQUENCY DOWN	Decrement IF Offset
PROGRAM CHANNEL	Store IF Offset in EEPROM
SCAN	Return to operating mode

Using the Controller with Other Receivers

The controller should work with most HF voltage-tuned receivers. Operation at higher frequencies is limited by the PIC prescaler; operation at much lower frequencies probably requires a higher-valued drain choke in the LO buffer. With a 16-bit tuning resolution and limited tuning voltage swing, the controller is probably useful only with single-band receivers.

There are certain assumptions in the software that may need to be modified for your particular radio:

1. The LO is assumed to tune on the low side such that IF = RF – LO. For high-side tuning, the PIC 2 code must change to subtract the IF offset, rather than add it. For direct-conversion receivers, the IF offset is set to zero.
2. The displayed frequency resolution is 1 kHz, which is adequate for my purposes in the 10-meter band. Resolution may be increased by increasing the width of the gate pulse generated by PIC 2, decreasing PIC 1's prescaler ratio, or both, and suitably adjusting the formula that calculates tuned frequency.
3. It may be convenient to generate a received-signal indicator that goes low when a signal is received. If so, the received-signal-detection logic needs to be reversed.
4. If your application differs from mine, the default values for the gate-pulse multiplier (10,075) and IF (10,916 kHz) should be changed in the PIC 2 code to make initial adjustments easier.

Construction Tips

With two PIC clocks generating 8-MHz square waves and other digital activity depending on which buttons are being pushed, the controller board generates quite a bit of noise. It is impossible to shield the board and interconnecting wiring too thoroughly. If you are adding the controller to an existing receiver, I recommend housing the two in separate shielded enclosures. My receiver and controller are mounted in the same enclosure, and I am still working on shielding.

If you will be making any changes to the software, make sure to plug the PICs into 18-pin DIP sockets that move with the PIC. There will be one socket on the controller board, and one attached to the PIC until software work is complete. Tearing off socket pins during repeated insertions and extractions is much less expensive than tearing off PIC pins. The DIP sockets fit just as well as IC pins into the zero-insertion-force socket on the programmer.

Notes

[1]J. Hansen, W2FS, "Using PIC Microcontrollers in Amateur Radio Projects," *QST*, Oct 1998, pp 34-40.

[2]S. Hageman, "A Synthesized 2-Meter FM Receiver with PC Control," *QST*, Feb 1999, pp 35-40.

[3]S. Hageman, "PIC Development on a Shoestring," *QST*, Mar 1999, pp 49-51.

[4]The newer PIC 16F87X series of FLASH-programmable microcontrollers alleviates this problem.

[5]N. Heckt, "A PIC-Based Digital Frequency Display," *QST*, May 1997, pp 36-38.

[6]F. Testa, AN617: "Fixed-Point Routines," Microchip Technology, Inc, 1996.

[7]S. D'Souza, AN593: "Serial Port Routines Without Using Timer0," Microchip Technology, Inc, 1997.

[8]**www.ccc.nottingham.ac.uk/~cczsteve/pic84.html**

[9]You can download this package from the ARRL Web **http://www.arrl.org/files/qex/**. Look for 0001DUNC.ZIP.

Thomas has been a radio amateur for only a few months, having after about 30 years of procrastinating finally developed his code skill enough to get a Tech Plus license. His credentials include BSEE and MSEE degrees from Georgia Tech and a First Class Radiotelephone license issued in 1969, despite which he has never been employed as an electrical engineer or technician. For the past 23 years, he has been a computer programmer for scientific and engineering systems.

ARRL Technical Awards

Nominations are now open for an exciting slate of national ARRL awards:
- ARRL Technical Service Award
- ARRL Technical Innovation Award
- ARRL Microwave Development Award

The ARRL Board of Directors has created a program of technical awards to recognize service, innovation and microwave development. In taking this action, the Board answered international, federal and organization motivations: International regulations recognize the "technical investigations" conducted by the Amateur Service. The FCC recognizes the amateur's "proven ability to contribute to the advancement of the radio art," the need to promote technical skills and expand the "existing reservoir . . . of trained operators, technicians and electronics experts." The League's purposes include promoting experimentation and education in the field of electronic communication, research, development and dissemination of technical, educational and scientific information.

All of these purposes are fundamental to the amateur service and they deserve recognition. Hence, the Board's initiative to encourage and reward achievement in technical areas. The ARRL technical awards encourage and tangibly reward amateurs who have given outstanding technical service. The awards provide an opportunity to publicly recognize amateur's service, Amateur Radio and its many benefits for society.

ARRL Technical Service Award

The annual Technical Service Award is given to a licensed radio amateur whose service to the amateur community and/or society is exemplary in the area of Amateur Radio technical activities. These include, but are not limited to:

- Leadership or participation in technically-oriented organizational affairs at the local or national level.
- Service as an official ARRL technical volunteer: Technical Advisor, Technical Coordinator, Technical Specialist.
- Communication of technical education and achievements with others through articles in Amateur Radio literature or presentations at club meetings, hamfests and conventions—this includes encouraging others to do the same.
- Promotion of technical advances and experimentation at VHF/UHF, with specialized modes and work with enthusiasts in these fields.
- Service as a technical advisor to clubs sponsoring classes to obtain or upgrade amateur licenses.
- Service as a technical advisor to Amateur Radio service providers, government and relief agencies establishing emergency-communication networks.
- Aid to amateurs needing specialized technical advice by referral to appropriate sources.
- Aid local clubs to develop RFI/TVI committees and render technical assistance to them as needed.
- Aid to local technical-program committees to arrange suitable programs for ARRL hamfests and conventions.

Each Technical Service Award winner receives an engraved plaque and travel expenses to attend an ARRL convention for the formal award presentation.

ARRL Technical Innovation Award

The amateur community has witnessed great technological changes over the past 75 years. Amateurs have been at the heart of many advances in the radio art. Enduring ARRL policy encourages amateurs to remain at the forefront of technological advancement. The ARRL's annual Technical Innovation Award is granted to a licensed radio amateur whose accomplishments and contributions are most exemplary in the areas of technical research, development and application of new ideas and future systems of Amateur Radio. These include, but are not limited to:

- Promotion and development of higher-speed modems and improved packet radio protocols.
- Promotion of personal computers in Amateur Radio applications.
- Activities to increase efficient use of the amateur spectrum.
- Work to alleviate long-standing technical problems, such as antenna restriction, competition for spectrum and EMC.
- Digital voice experimentation.
- Amateur satellite development work to improve portable and mobile communication.
- Improvements in solar, natural and alternative power sources for field applications.
- Practical compressed video transmission systems.
- Spread-spectrum technology and applications.
- Technology development to assist disabled amateurs.

Each Technical Innovation Award winner receives a cash award of $500, an engraved plaque and travel expenses to attend an ARRL convention for the formal presentation.

ARRL Microwave Development Award

The microwave and millimeter bands are great frontiers for amateurs. They present amateurs with a vast test area for development of both new and traditional modes. The annual ARRL Microwave Development Award is given to an amateur individual or group whose accomplishments and contributions are exemplary in the area of microwave development. That is, research and application of new or refined activity in the amateur microwave bands at 1 GHz and above. This includes adaptation of new modes in both terrestrial and satellite techniques.

Each Microwave Development Award winner receives an engraved plaque and travel expenses to attend an ARRL convention for the formal presentation.

Nominate Now!

Count yourself among those who know that technical advancement is not a lost ideal in the amateur community. Now is the time to nominate your colleagues for one or all of the awards described above!

Formal nominations may be made by any ARRL member. Submit support information along with the nomination document, including endorsements of ARRL affiliated clubs and League officials. Nominations should thoroughly document the nominee's record of technical service during the previous calendar year. Information concerning the character of the nominee should be as complete as possible. The ARRL's Amateur Radio Technology Working Group will serve as the award panel, review the nominations and select the winners.

Send nominations to: ARRL Technical Awards, 225 Main St, Newington, CT 06111. Nominations must be received at Headquarters by March 31, 2000. Send any questions to Headquarters or e-mail **jwolfgang@arrl.org**.

Automotive RFI Elimination

Modern vehicles are RF-noisy environments. Come learn how to identify and silence your mobile noise sources.

By Stuart G. Downs, P.E., WA6PDP
ARRL Technical Specialist

[Editor's note: Readers should check with their automobile manufacturers' EMC departments before performing any modifications that may void vehicle warranties.]

Automotive RFI can drive hams crazy and I was no exception. Even with a good noise blanker, RFI makes it very difficult—if not impossible—to hear weak stations. It is also difficult to identify noise sources using a mobile HF radio as a noise-proximity tester, that is, to obtain noise signatures by picking up other noisy vehicles, power lines, traffic lights, etc. It is theoretically possible to eliminate all automobile RFI noise.

Automotive Background

Automobile manufacturers have minimized conducted and radiated ignition, electric-motor and generator noise problems by adding inductive/resistive spark-plug wires, resistive spark plugs and judiciously placed capacitors. This was done primarily to reduce noise in the AM and FM broad-cast bands. The theory behind this is quite simple. By placing resistive and inductive distributed elements in a series circuit, voltage rise times are increased and current is decreased. In addition, a capacitor placed between power and ground provides a low-impedance return path for low-frequency noise. The problem is the capacitor looks inductive (lead reactance) as the frequency gets higher. At some frequency, the component self-resonates. Rise time is related to the spectral content by the following approximation:

$$f_{\text{(occupied BW)}} \approx \frac{0.35}{t_{\text{(rise time)}}} \qquad \text{(Eq 1)}$$

If the rise time increases, the spectral energy content decreases. With charge flow lessened by the increased inductance and resistance, the magnitude of the radiated electromagnetic field is also reduced. These noise-reduction techniques were not intended for HF ham radios, but they do help a little. The problem is that it's not enough. Broadband noise is picked up by our HF rigs! Essentially, we drive around with spark-gap transmitters under our hoods, connected to ignition-wire antennas!

To help understand how things radiate, let's look at things in mathematical terms. The time rate of change of voltage (dv/dt) and current (di/dt) both are reduced using the previously mentioned methods. This is easily derived from the first-order differential equations for resistive or inductive spark-plug wire systems. Keep in mind that accelerating charge radiates electromagnetic energy. To understand this, consider the following:

$$I = \frac{dq}{dt} \qquad \text{(Eq 2)}$$

Current flow (in amperes) is the rate of change of charge (in coulombs) per unit time. A constant dc current moves charge (electrons) through a wire and, for all practical purposes creates a static (unchanging) magnetic field. When the current is time varying, say:

$$I(t) = K \sin(\omega t) \qquad \text{(Eq 3)}$$

where K is the peak magnitude and sin (wt) describes the function of time ($\omega = 2\pi f$), things start to happen: A travelling electromagnetic wave is produced. Taking the time derivative of Eq 3's current yields:

$$\frac{d^2 q}{dt^2} = K\omega \cos(\omega t) \qquad \text{(Eq 4)}$$

This equation represents accelerating charge that radiates electromagnetic energy. With dc current, there is no electromagnetic radiation; with a time-varying current, you get electromagnetic radiation. See the Appendix for proof.

Solutions to ignition impulse noise have already been found and successfully demonstrated in aircraft with internal-combustion engines. The VHF, AM aircraft band is especially susceptible to ignition noise. Aircraft manufactures fixed these problems so aviators could use their AM radios. Next time you have a chance to look at an aircraft engine, observe the shield-

10983 Shy Bird Ln
San Diego, CA 92128-4017
stu.downs@trw.com

ing on the ignition wires.

My Mobile Rig

After getting on HF again following a long absence, I decided to install my newly purchased Kenwood TS-570D transceiver into my 1995 Toyota 4WD 4-Runner. After routing power and ground wires directly from the battery terminals and using a network analyzer to tune my ham-stick antennas (courtesy of IDAFAB's Glenn Borland, KF6VZK), I thought I was ready for my first QSO driving while hamming. I couldn't have been more disappointed. The RFI level was between S-7 and S-9. Murphy had stricken again. Oh boy—what to do?

Not long after this, back at home I heard a mobile-7 station on Interstate 10 east of Tucson working the Philippines. I just about fell out of my chair. Weeks later, I was on a one-day picnic trip to Anza-Borrego Desert State Park. With the engine turned off, I heard a station in Uganda on 10 meters and worked him! I never would have been able to hear him with the engine on: There would have been too much RFI. I realized right then that I had to eliminate the noise. Elimination of RFI noise is not black magic, but it follows theory just like everything else. It usually involves a multi-step process.

Conducted and Radiated Noise

Two different types of noise need to be considered in automobile RFI control: conducted and radiated. Conducted noise is called that because it comes through wires and other conductors directly from the noise source. Radiated noise enters the receiver at the antenna via electromagnetic fields from the source. To effectively get rid of any noise source, one must first understand if it is conducted or radiated and then determine where it's coming from. If one has no theoretical understanding, it's all the more difficult and one is left with the process of "Easter-egging." That is, you might get lucky and find the cause by adding all sorts of time-tried solutions.

The noise sources in my automobile were determined to be in the ignition system, which includes the spark plugs, wires, distributor and coil, dc motors and a noisy oil-pressure sender.

Conducted-Noise Tests

The radio amateur needs to first determine if the conducted noise on their vehicle's 12-V power system is producing noise in the their rig. To determine this, do the following:

1. Start the engine and switch the rig on. This must be done without an antenna because you don't want to confuse conducted noise with radiated noise that is picked up through the antenna. Make sure the air conditioning, all fans, blowers, blinkers and windshield wipers are turned off.
2. Turn your transceiver's mode switch to AM and listen on HF, then increase the engine speed a little.
3. Do you hear any popping, whining, crackling or any other noise that increases with engine speed?
4. Turn the engine off. Now turn on the fans, blowers and air conditioning one at a time.
5. Do you hear any whining?

Did you notice any noise in steps 3 or 5? If you did, the noise you heard or saw on the S meter is conducted through your dc bus. If no noise was heard, your rig is not susceptible to the conducted 12-V power noise. If you heard noise in steps 3 or 5 you will need to install a low-pass filter that is capable of operating with your trans-mitter's load current, typically about 20 A for a 100-W HF rig. With derating, the filter should be able to pass around 25 A or so with a low voltage drop. Obviously, the filter's dc resistance needs to be quite low or you'll get a significant voltage drop between the battery and your rig.

The ideal filter for this application would be of the L-C variety, containing a series inductor and shunt capacitor. This filter's forward voltage transfer function decreases noise on the output as a function of frequency. Dc output

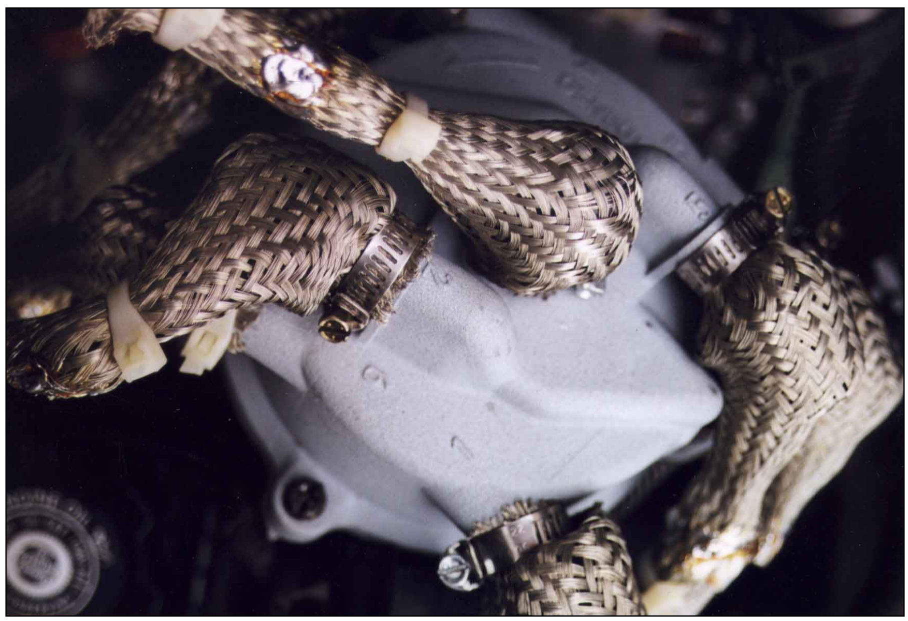


Fig 1—Braid installed over spark-plug wires at distributor.

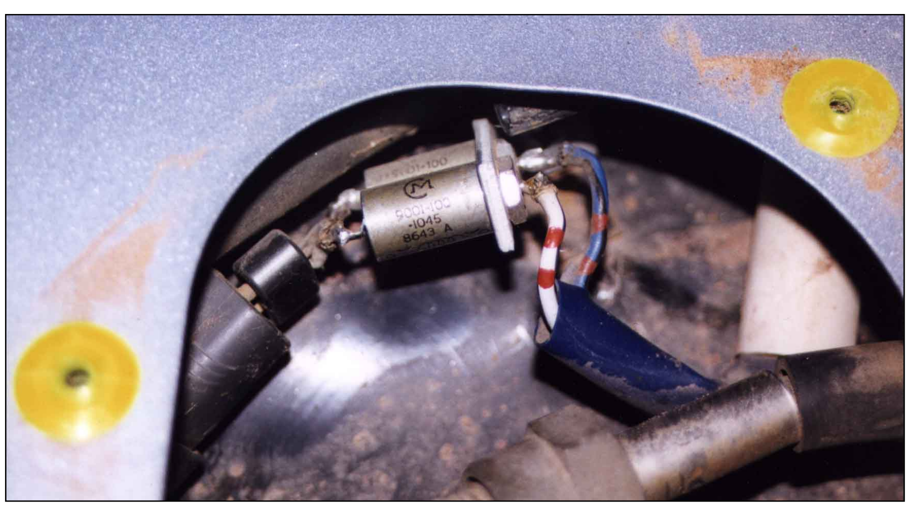


Fig 2—Filter cons installed on bracket.

current and voltage will not be affected. Do not use a pi filter here. Such a filter may couple noise from the dc bus down through the first capacitor to ground, then up through the second capacitor into your rig! The inductor side should be attached to the battery and the inductor-capacitor node to the radio side. The best type of power filter for this application is called a "filter con."

Filters

To reduce RFI, one must understand filter basics. Filters can reduce or eliminate both conducted and radiated noise. Choosing the correct filter for a particular application can be a significant task in itself. Say, for instance, you want to get rid of noise on a power supply's output. You think that adding a capacitor will solve the problem; but after placing the capacitor, you find it did no good! Among other things, filter placement depends on source impedance. Noise from a low-impedance source cannot be reduced by a capacitor alone. It is simply a matter of voltage division:

$$V_{out} = V_{in}\left(\frac{Z_{load}}{Z_{source} + Z_{load}}\right) \qquad \text{(Eq 5)}$$

If Z_{source} is small, you can take it to zero. Then you're left with:

$$V_{out} = V_{in} \qquad \text{(Eq 6)}$$

The point is to think about what is happening in the circuit. The source impedance may determine what filter to add. To get good attenuation from a low-impedance source, one needs to add an L-C filter. Batteries are typically low-impedance sources!

To understand this better, think of a voltage source having zero output impedance in series with an impedance *Z*. *Z* can be resistive, inductive, or some combination of both; it represents the opposition to current flow. In the case of the lone capacitor, it had no impedance to work with. Therefore, the noise currents charged up the cap quickly because its impedance was larger than that of the source.

Power Filter Cons

Power filter cons are nice solutions to ham radio automotive noise problems for the following reasons:

1. L-C (series-L, shunt C), C-L-C (shunt-C, series-L, shunt-C, a pi network) and L-C-L (series-L, shunt-C, series-L, a T network) varieties are available at surplus and from a variety of manufacturers.
2. The filter components are hermetically sealed in a metal tube; so they can't radiate and are easily mounted with a bracket, washer and nut.
3. The parasitic reactances have already been minimized.
4. The self-resonant frequencies of the filters are high.
5. They come in various current ratings.

Second-order filter cons typically attenuate noise at –40 dB/decade of frequency above the 3-dB point.

$$f_{-3dB} \approx \frac{1}{2\pi\sqrt{LC}} \qquad \text{(Eq 7)}$$

Filter cons are good up to a couple hundred megahertz. Typical series-inductance values for 25-A dc power-line filters would be in the range of 100 to 200 µH, with a shunt capacitor of about 1 to 2 µF. The break-point frequencies of these filters are about 15 kHz, or so. Essentially, this filter forms an ac voltage divider with the noise source's impedance, thus reducing ac noise at its output—more about this later. Smaller filter cons rated at 5 A or so have less inductance.

This type of filter can be large depending upon the voltage and current ratings: The inductor must not saturate. It must support the full load current, which can be up to 25 A for a 100-W transmitter—and don't forget derating. Inductor size is directly proportional to the amount current flowing through it. If the inductor saturates, you have only the wire resistance and little additional inductance. In this case, you would just have a first-order filter composed of nothing more than an R and C—not very good! Take the time to select the right filter up front.

Make sure that the ground return on your rig goes right to a solder lug on the bracket as close to the filter con as possible. After the filter is installed, repeat the above tests to see how effective it is.

Most transceivers have filters on the dc power line coming into the transceiver. In my case, the Kenwood TS-570D did not have a conducted noise problem. I did not see any indication on the S meter, nor did I hear any conducted noise. I also looked on an oscilloscope connected across my battery and noticed low-frequency ripple similar to what one would see out of a full-wave bridge rectifier: about 100 mV (P-P) punctuated with high-frequency, damped sinusoids. It was obvious that the filter in the rig was adequate.

Radiated-Noise Tests

1. Switch on your rig with an antenna connected and notice the ambient noise level with the engine off. It is better to do radiated tests on days when the ambient noise is not readable on the S meter.
2. Next, turn your mode switch to AM—with the noise blanker off—and start your engine. You'll probably hear a popping noise. Note the S-meter reading.
3. Increase the engine speed. Do you notice that the popping noise increases with engine speed? The popping noise is impulse noise from the ignition system; it is caused by the current flowing in various parts of the engine. In another words, your ignition system is radiating, and your antenna is picking it up. This is the noise problem that you should probably solve first.

If you already have resistive spark plugs and wires, the voltage and current rates of change d*v*/d*t* and d*i*/d*t* should not be altered. Something else must be done. Grounding the hood will not eliminate this noise. We must attenuate the electromagnetic radiation without affecting d*v*/d*t* or d*i*/d*t*, both of which are necessary for proper engine performance.

Electromagnetic radiation of this sort *can* be greatly reduced. Maxwell's equations—the four that describe all electromagnetic waves—can yield great insight to the behavior of fields at boundary conditions. One boundary condition for E fields occurs as follows: The tangential E-field component of a propagating electromagnetic wave decreases to zero at the surface of a perfect conductor (metal) and is continuous across the boundary. That is to say: The E field shrinks to a small value across the skin depth of a metal shield, since there is no perfect conductor. The normal component of the H field passes through the metal boundary. The E-field attenuation typically will be on the order of 60 dB or so. That's a 1000-to-1 change! To greatly reduce or eliminate ignition noise, therefore, we must shield the ignition system—it's the only way. Don't worry; it's a job, but not as hard as you might think.

In shielding my ignition system, I wanted to ensure that I had good RF grounds and low-impedance current-return paths. That is, I wanted to be certain that—while each spark-plug fires—the return current travels via a shield over the spark-plug wire. The current path from the coil through the spark plug wires to the spark plug and back to the coil should be as direct and non-inductive as possible. Any other ground paths on an engine might look inductive and present a virtually open circuit at RF! The plan, therefore, was to use flexible braid as a low-impedance ground path.

Shielding My Toyota's Ignition Spark-Gap Transmitter

My ignition system was shielded in the following manner. First, I had copper-tube spark plug covers made for all six spark plugs. They snap into place over the base of the plugs (ground potential) with a slight interference fit. I then soldered flexible braid to the outside of each spark plug's copper cover. In this way, I was able to snap the cover over the plug at its base to provide a good ground, at the same time plugging the electrical-connection boot onto the spark plug in one motion. The copper spark-plug covers were made for me by Tooling Associates.[1]

There are two common spark-plug-base sizes. Toyota spark-plugs in a 4-Runner are of the smaller-diameter. To shield them, I used 1/2-inch flexible braid (Belden #8669)[2] over each spark-plug wire all the way from the spark plug to the distributor. To increase dielectric strength, I wrapped the spark-plug wires with insulated flexible tubing before installing the braid. With ground now located so close to the wires, I sure didn't want any high-voltage breakdown. I used 3/4-inch braid (Belden #8662) to cover the larger-diameter boots at the distributor.

Belden braid comes in 50-foot lengths, so maybe a club or group purchase would be best. You can find the braid in the Newark catalogue. So far, the silicone spark plug wires have exhibited no high-voltage breakdown.

[1]Notes appear on page 36.

Appendix: Strength of EM Fields Radiated by Accelerating Charge

The following proof quantifies the field strengths of an electromagnetic wave as created by accelerating charge in a wire [or antenna element—*Ed.*]. This derivation begins with one of the four basic laws of electromagnetism known as Maxwell's equations. These laws were formulated as a set by Maxwell, but certain parts of them were obtained directly from the work of others such as Gauss, Ampère and Thomson.

The first law may be stated: "The work required to carry a unit magnetic pole around a closed path is equal to the total current passing through any surface that has the path for its boundary." Vector calculus says the same thing more succinctly, here in integral form:

$$\oint H \bullet dl = I_{conduction} + \frac{\partial \phi_D}{\partial t} \qquad \text{(Eq 8)}$$

where the left-hand side represents a line integral around a closed path, dl is a vector element of length along that path, H is the magnetic-field vector and ϕ_D is the electric flux linking the path. See Fig A. The time rate of change of ϕ_D is written as a partial derivative to show that the path of integration is not moving in space.

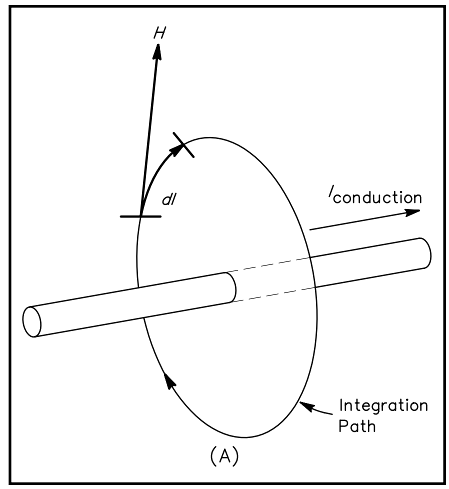


Fig A—The situation of Eqs 8 through 12.

In a wire, there is no E field, so we are left with:

$$\int_0^{2\pi r} H \bullet dl = I_{conduction} \qquad \text{(Eq 9)}$$

$$= 2\pi rH$$

where r is the distance from the wire. But, current is the time rate of change of charge, q:

$$I_{conduction} = \frac{\partial q}{\partial t} \qquad \text{(Eq 10)}$$

and so:

$$\frac{\partial q}{\partial t} = 2\pi rH \qquad \text{(Eq 11)}$$

Now give the charge some acceleration (changing velocity). Taking the time derivative of both sides:

$$\frac{\partial^2 q}{\partial t^2} = 2\pi r\frac{\partial H}{\partial t} \qquad \text{(Eq 12)}$$

This shows that an accelerating charge generates a magnetic field that changes with time. A changing magnetic field is related to the electric field by Maxwell's second law, which may be stated: "The work required to carry a unit electric charge around a closed path is equal to minus the time rate of change of the magnetic flux through that path." Again using integral form:

$$\oint E \bullet dl = -\frac{\partial \phi_B}{\partial t} \qquad \text{(Eq 13)}$$

where E is the electric-field vector and ϕ_B is the magnetic flux linking the path of integration. See Fig B. We have an electromagnetic wave!

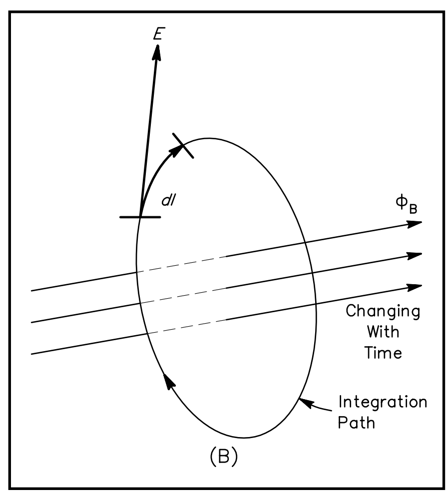


Fig B—The situation of Eq 13.

Maxwell's equations may also be written in differential form. An excellent book that explains this vector calculus stuff in understandable ways is H. M. Schey, *Div, Grad, Curl and All That*, 3rd edition, ISBN 0-393-96997-5.

Second, the distributor was zinc "flame-sprayed" to create a volumetric metal enclosure (shield) only a few mils thick, but still thick enough that the E field exponentially dies off and meets the E-field boundary criteria. This could also have been done with the coil. Flame Spray, Inc of San Diego[3] graciously sandblasted and flame-sprayed the distributor. The distributor had to be masked so that the metal did not cover areas where its was not needed. Plastic parts are routinely sprayed for EMI/RFI protection in military and commercial applications, and the flame-spray people are in this business.

Taping of the distributor can be a task; otherwise high-voltage arcing can occur. I taped the coil myself with 1-inch-wide, 1-mil-thick copper tape, which was supplied by Chomerics of Woburn, Massachusetts.[4]

Third, I used narrow hose clamps to connect the ends of the braid to the wires to the distributor and coil below the insulating boots. It worked quite well. See Fig 1.

After completely shielding the ignition wires, coil and distributor, I studied the ignition circuitry for *sneaky* RF paths. Next, I put two filter cons between the +12-V power line, the igniter and the coil. I mounted them on a bracket with a good RF ground next to the coil and igniter. The reason for this was simple: I did not want broadband RF energy in the wire from the distributor to the coil to get into my 12-V power line and radiate.

All coils and transformers are self-resonant and start to look capacitive—rather than inductive—at some frequency. The Toyota's high-voltage ignition coil is no exception. Remember that all current—even RF—wants to return to its source. Why not give it a low-impedance path by making the path short? Minimizing the current-loop size minimizes the radiated energy, as well.

Ah, but what kind of filter con should I select? In this case, an L-C filter con was the optimum choice because the output capacitor shunts the dc power at the coil and igniter right at its source, creating a very short return path for RF. The series inductance presents a large inductive reactance that keeps RF off the +12-V power line.

Judging from the wire gauge from the battery to the igniter and coil, the dc current must be on the order of a couple of amperes. Therefore, I used a 5-A filter con that's about the size of a fat pencil and about 1-inch long. Its dc working voltage is 50. See Fig 2.

Fuel-Pump Noise

When I finished the job, I started the motor, switched the radio to AM and noticed that the ignition noise was gone! Now I heard another noise that had been masked by the first noise source: a whine that did not change with engine speed. What was it? — Some kind of motor that was radiating! Since all other motors and fans were turned off, it could only be the fuel-pump motor inside the gas tank. With the engine on, I pulled the fuse to the fuel system. The noise stopped—I had found it!

Again, I thought about EM theory. I knew it was motor noise and that every motor has a rotor and a stator. I thought there must be some sort of brushes making sparks. Since sparks make broadband noise and rotors and stators can self-resonate, I theorized that the RF energy must be coupling capacitively through the rotor or stator and looping back through my power system. After some ground testing, I found two ground returns for the fuel pump system: one through the vehicle ground and one via a power-return wire! I wanted to put the filter cons right at the source, but that was going to be tough unless I removed the gas tank—which I did not want to do. As it turned out, there was an access panel on top of the gas tank, accessible through the back seat. Again, I wanted to short out any RF return currents and short out the sneak path.

The place to do this was at the top of the gas tank. I mounted two 5-A filter cons on a bracket, in turn connected to a tab through a self-tapping screw made for holding the gas-tank wires to the top of the tank. Fuel-pump power and return passed through both of the filter cons. By now, you're probably wondering what kind I used. I used a low-pass pi network. It worked! When I turned the engine back on, the noise from the fuel pump was gone!

Oil-Pressure-Sender Rheostat Noise

Well, by now I had spent considerable effort eliminating noise problems, but I wasn't going to quit now! With the engine running for a while, I noticed that I had no noise for a couple of seconds—then all heck broke loose. The next noise did not whine but crackled. When I increased the engine speed, it stayed nearly constant. When I decreased engine speed, it went away for a couple of seconds, then came back. I wondered what would cause this.

I took out my scope and looked on the +12-V bus. I saw some noise but thought it was too low in frequency to cause what I was hearing. Obviously, the real source was something that switched on shortly after engine start and changed after decreasing engine speed. I thought it might be the sender in the fuel pump and therefore pulled the gauge fuse, opening all sender circuits. The noise stopped! I then knew it was some sort of sender, but which one?

Then it hit me: Oil pressure rises when the engine is first started, until it reaches its proper value and it changes again with engine speed—some sort of rheostat. I pulled the connector from the oil-pressure sender and the noise stopped! I looked at my S meter. There was no noise indication, and I could not hear any vehicle noise! I removed the sender unit from the engine. With help from Glenn Borland, KF6VZK, I cleaned the sensor body of "chem" film and soldered a filter con onto it. We used a high-power soldering iron for this because we wanted the job done quickly to avoid damaging the sender. This requires a large, hot thermal mass. There was no good place to put a bracket. Best was to mount the filter con directly onto the sensor and that's what we did.

Shake-Down Test

Since I had put many hours into solving my RFI problems, Janice and I decided to go to Anza-Borrego Desert State Park for a picnic and a hike. The HF rig went along too, although it was in the back seat. (Never put the rig on the right-front seat with the girlfriend or wife along—it might find its way onto the highway!)

With the engine on, there was no noise indication on the S meter in the AM mode. I quickly turned to SSB. I could hear all kinds of stations—even weak ones; I worked several stations and thought how nice it was to be able to finally use my rig to its full potential without limiting RFI. We hiked up Palm Canyon and were rewarded a sighting of a bighorn sheep—a borrego. It was a great day!

Notes

[1]Contact Pete Bird, 15570 Corte Montanoso, San Diego, CA 92127; tel 619-672-1949.

[2]Belden Wire & Cable Co, PO Box 1980, Richmond, IN 47374; tel 765-983-5200, fax 765-983-5257; http://www.belden.com/.

[3]Flame-Spray Inc, 4674 Alvarado Canyon Rd, San Diego, CA 92120-4304; tel 619-283-2007; contact Gary Logan.

[4]Chomerics Division, Parker Hannifin Corp, 77 Dragon Ct, Woburn, MA 01888-4014; tel 781-935-4850, fax 781-933-4318; mailbox @chomerics.com; http://www.chomerics .com/. Contact Dennis Hennigan, WA1HOG, extension 4140.

A Homebrew Logic Analyzer

Digital designers, experimenters and technicians need to look at signal sequences. Their tool of choice is a logic analyzer. This homebrew adapter and software turn a PC into an eight-channel analyzer suitable for serious work and digital exploration.

By Larry Cicchinelli, K3PTO

With the increasing importance of digital communication in Amateur Radio, this project should be of interest to many hams. The logic analyzer (LA) described in this article will meet the needs of most home experimenters. The hardware design is such that individual builders may make changes and still use the software I have developed. The software is available as shareware. A free version is available; it uses no hardware other than your printer port (EPP mode).[1]

[1]You can download this package from the ARRL Web **http://www.arrl.org/files/qex/**. Look for LA0999.ZIP. It includes a complete set of documentation.

15116 Bellmont Ave
Allen Park, MI 48101
k3pto@arrl.net

This LA is very similar to earlier commercial instruments. It takes a sample every clock cycle and stores the data for display. The memory depth defines the number of samples, as well as the total sample interval. In this case, the memory size is 131 kB. Therefore, with a sample period of 200 ns, the sample interval is approximately 26 ms.

Newer LAs have much faster sample clocks—usually with periods of picoseconds. Instead of storing data at each sample period, they store data whenever a state change is detected. The time of the change is stored along with the data. This allows much greater flexibility in the data analysis, but also requires more memory, because the time and value must be stored with high precision for each sample.

Some of the general specifications are:

- 8 channels
- 5-V logic only
- Time resolution from 200 ns to 1 ms per sample
- *Windows* 3.1 or *Windows* 95 compatibility
- Enhanced parallel port required
- User configuration via .INI file
- Pre-Trigger capability of 4096 (2^{12}) samples
- Memory depth of 131 kB (2^{17} bytes)
- Circuits built with HCMOS parts—usually good to at least 25 MHz.

The operation of each circuit is described so that the reader might gain some insight as to how they work, as well as ideas for modification. A few possible suggestions for modifications are:

- Increase clock speed to 20 MHz to allow 50-ns sampling
- Change the Pre-Trigger memory size

• Allow more sampling-rate choices, including an external input

Here are a few notes on the circuit descriptions. The suffix characters of the signal names have the following meanings:

PL = Pulse Low, the active level is a pulse with a steady-state value of 5 V, but pulses low to 0 V for a short period of time

PH = Pulse High, similar to above, but pulses high

L = Active level of 0 V

H = Active level of 5 V

timing values are usually indicated with Hex notation ($FFF, for example)

I use this system because I do a lot of programming in *C*; only those signals that go between the circuits have been named.

The strip menu has a **DEBUG** entry, which allows the builder to test the system. I did most of my hardware debugging using this facility. Several of the circuits can be tested as they are built. Not all of the ICs need be installed before testing begins. A suggested construction order is: Control, Time Base, Address Generator, Input, Trigger, RAM and Pre-Trigger. The Time Base and Address Generator can be built and tested independently of the Control circuit, if desired.

The **HELP** menu is not available at the time of this writing. I hope to have it ready by the time this article hits the streets. It will contain much of the operational information in this article, as well as details on the **DEBUG** options.

I am very interested in any feedback that anyone may have about the operation of this system. I will try to incorporate any changes that enhance its operation.

Theory of Circuit Operation

Fig 1 shows a block diagram of the LA. Refer to it and the individual circuit schematics as you read.

Address Generator

This circuit (Fig 2, built on a daughter board) is constructed using synchronous binary counters. As such, they generate a binary sequence creating the address information for both the RAM IC and the Pre-Trigger address. The advantage of a synchronous counter is that all outputs change state on the same clock edge. This insures that there is no delay accumulation from the least to the most-significant stages.

The operation of the circuit is as follows: Until a trigger signal occurs, U404 and U405 are disabled, thus allowing U401 through U403 to form the address counter for the Pre-Trigger memory. U401 through U403 start counting immediately following the AddrReset-PL signal. They will continue to count until they reach $FFF ($4095_{10}$), at which point the count rolls over, and begins again at $000. This continues until a trigger signal occurs. At that time, both the Triggered-PL and the Triggered-H signals will occur. The Triggered-PL will preset the address counter to a value of $21000. Please note two significant facts about this value: (1) the $20000 bit, bit 17, is *not* used as an address bit, and (2) the address is one count beyond the Pre-Trigger range.

The Triggered-H (via HighAddrEna-H) signal enables the last two ICs in the circuit. Eventually the count will reach $3FFFF and roll over to $40000. This inhibits any further counting, because the AddrDone-L signal, derived from bit 17, is fed back to one of the enable inputs of U401, the first IC in the counter chain. The AddrDone-H signal is used to tell the PC that all the samples have been stored.

When the system is sampling the input data, the AddrClk-PH signal is derived from the Time Base circuit. When the data needs to be dumped to the PC, the signal is derived from the PC via one of the control-port bits.

Note that the same signal is used to advance the counter as well as to store the data value into RAM. This is valid because the RAM chip does not require any data hold time to store the data. Therefore, by the time the address lines have changed (specified as 30 ns, typical), the RAM is done with the data for the address just ended.

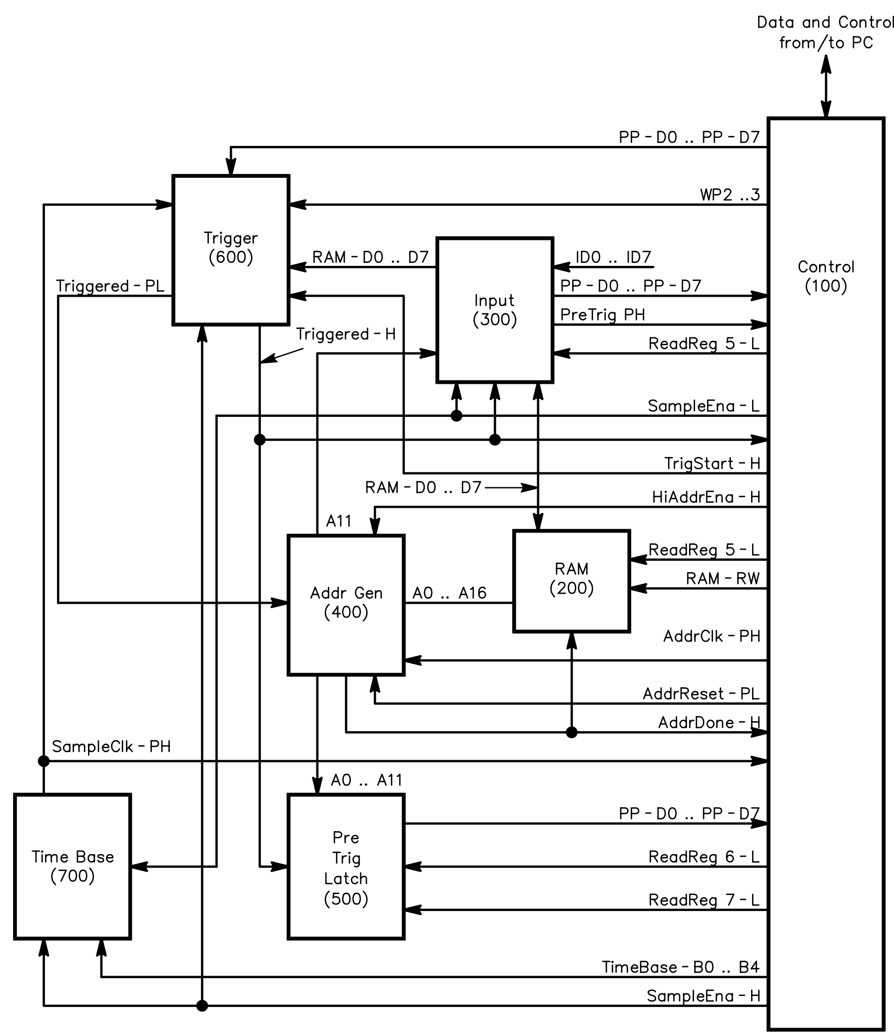


Fig 1—Logic Analyzer block diagram.

When the system is not sampling input data, the clock signal for the counter chain is derived from the Strobe. The AddrReset-PL signal is delayed by an RC delay circuit in order to ensure that the Load signal of U405 is still low when the clock signal goes high. This is required because the 74HC161 has a synchronous load circuit, meaning that the counter performs its load operation only while the load input is low and when the clock goes high.

Control

This section (see Fig 3) contains several different functions:

1. U104 is used to select whether the PC needs to read from, or write to, the system. It generates four Read Enable signals (only three of which are used) and four Write Pulses, which are used to latch control data from the PC.
2. U101 and U102 latch the LA control signals Time Base Select, Sample Enable and Address Reset
3. U105 controls the clock signals for the Address Generator and the RAM as well as the Count Enable for U404 and U405
4. U106A allows triggering to begin. This insures that the Trigger circuit will not be enabled until the Pre-Trigger RAM is full.

Input

Refer to Fig 4. The 74HC14 is a Schmitt trigger. As such, it gives the LA some degree of immunity from noise on the input signals. These signals are delivered to both the RAM and the Trigger circuits. Since the RAM data ports are bi-directional, this circuit also makes the RAM data available to the PC.

Please notice that this input circuit is valid for 5-V logic only! If you need any other logic levels, you *must* condition the signals before applying them to this circuit, or the ICs may be damaged. The input resistors (1 kΩ) give some degree of protection, since there are diodes within the 74HC14 that are connected from the input to both ground and power. The resistors are placed close to the input pins to minimize the effect of input (specified as 10 pF, maximum) and wiring capacitances. There is also a set of pull-down resistors to ensure that the HCMOS inputs never float.

Pre-Trigger

Refer to Fig 5. When a trigger signal occurs, the value of the next address will be stored in the Pre-Trigger address latch by the Triggered-H signal. This value is made available to the LA program so that it can determine the address—in the range $00000 to $00FFF (0 to 4095_{10})—at which the trigger event occurred. The

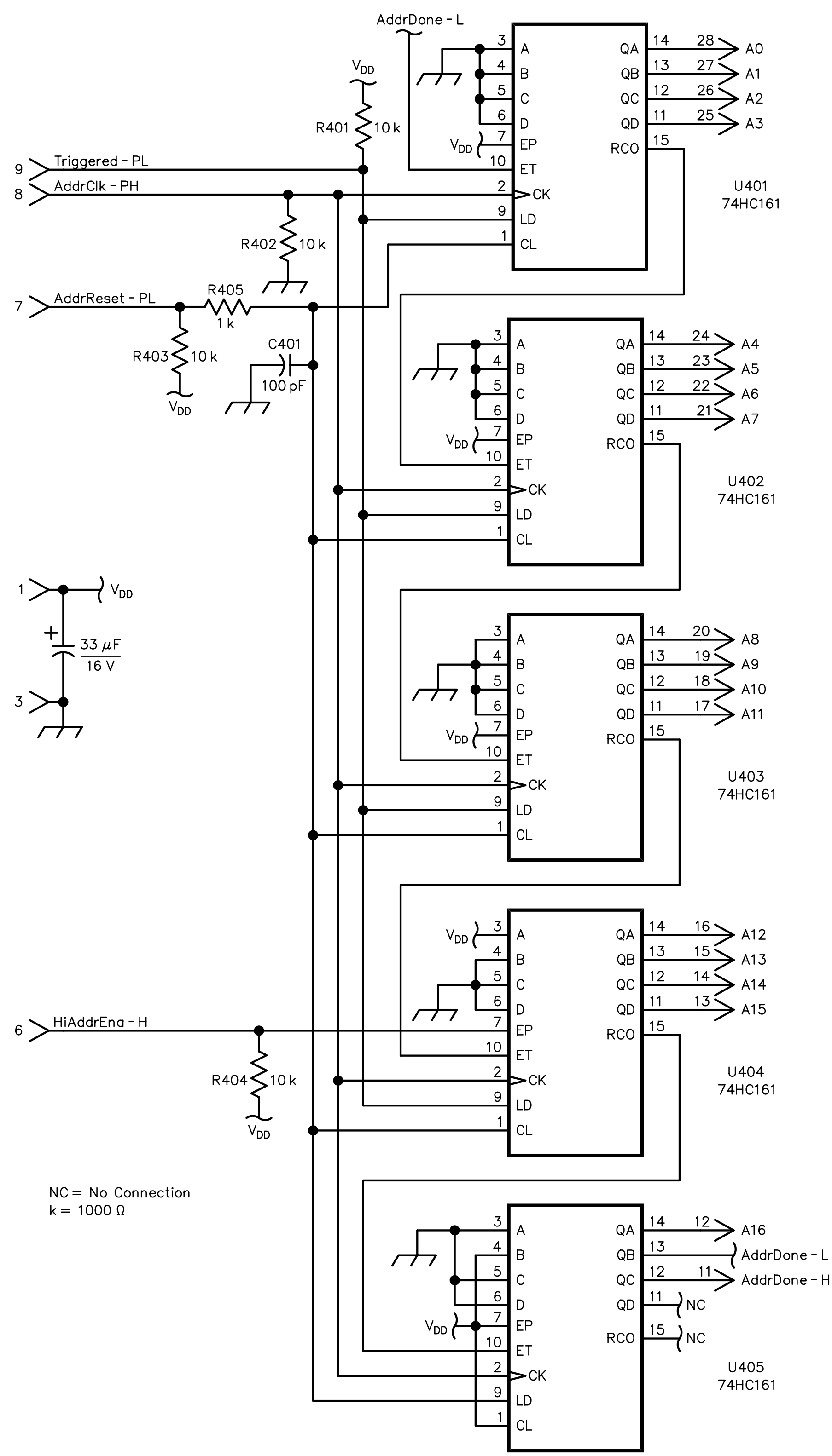


Fig 2—Address Generator schematic diagram.

program can then display the data values in the correct sequence for the Pre-Trigger display.

RAM

The RAM device (see Fig 6) was chosen mainly because it is available in a DIP. This device has a minimum clock-pulse-width requirement of 60 ns. This does not allow the system to operate at its fastest sample rate of 10 MHz (100 ns). It will work at 5 MHz (200 ns), which is fast enough for my present requirements. A very simple circuit modification would accommodate another memory device having a faster response time. It may be as simple as just plugging it in—with the signals routed appropriately.

When sampling the input data, the SampleClk-PH signal is applied to the R/W line. When reading the data back to the PC, the R/W signal is at a static level generated by the control logic.

Time Base

When the system is sampling the input data, this circuit (Fig 7, built on a daughter board) generates the sample clock used to advance the address counter, as well as the signal that writes the data into the RAM. Although only three bits are currently implemented to select the sample rate, there are five bits available. This allows up to 32 selections. The .INI file allows the user to define the sample period for each implemented selection. The primary reason for allowing two more bits, is that multiplier values of 1, 2 or 5 may be implemented. This gives sample periods that are similar to those of an oscilloscope.

Trigger

The Trigger circuit (Fig 8, built on a

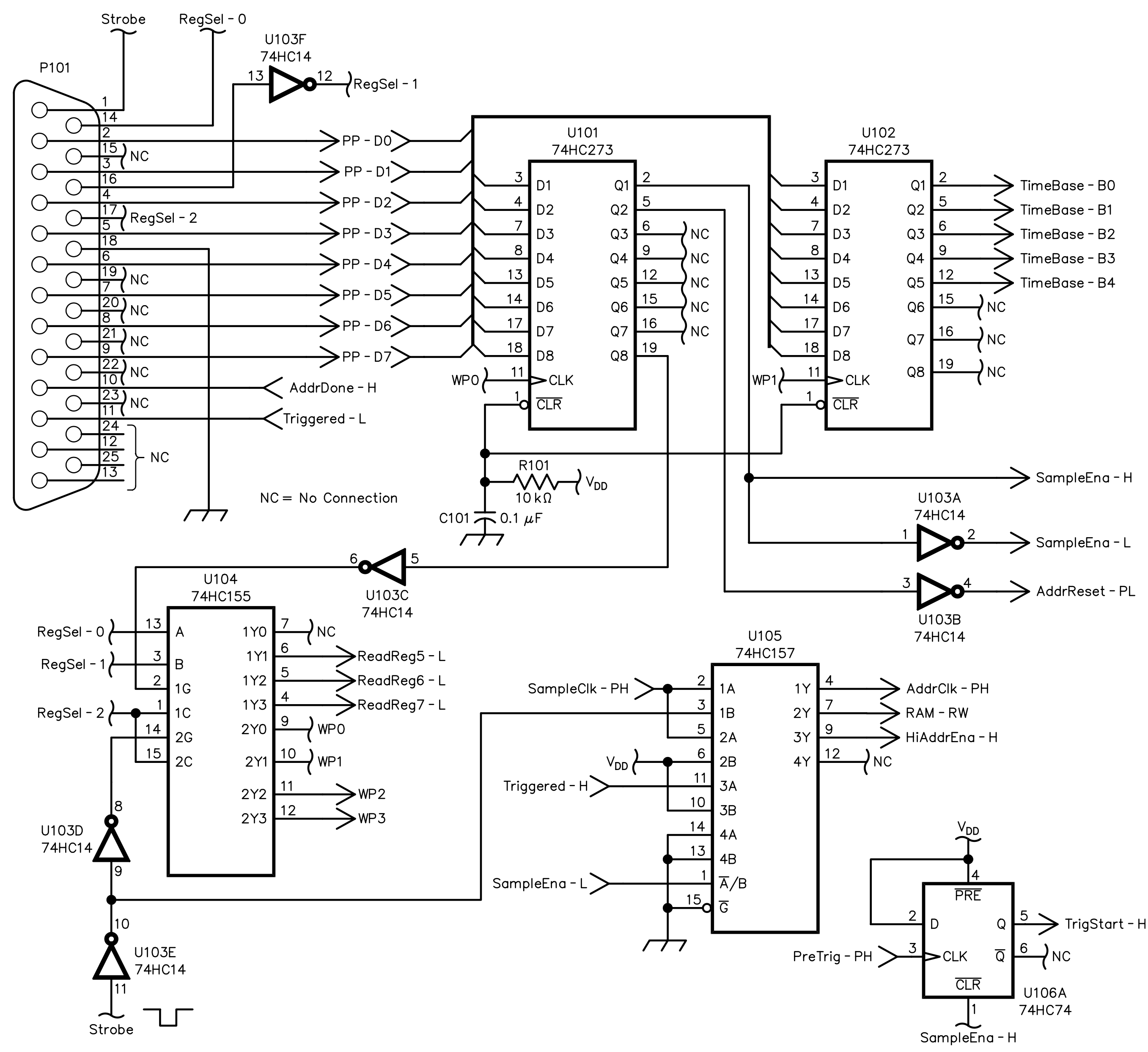


Fig 3—Control circuit schematic diagram.

daughter board) is used to generate the signals that enable the system to begin storing the input data beyond the Pre-Trigger address range. It allows the user to specify the Trigger State for each of the eight input lines individually. Any one of three values is allowed: 0, 1 or "don't care." The operation is such that when the system is enabled by the user, the Address Generator will generate only those addresses that are defined as the Pre-Trigger range. When a trigger event occurs, the Address Generator will still increment to the next address. This address is stored in the Pre-Trigger latch circuit. The trigger event will always be displayed at sample 4095_{10}. The next sample will be stored at address \$01000 (4096_{10}).

Table 1—Trigger States

Bit 1 0	*Trigger State*
0 0	always
0 1	never
1 0	0
1 1	1

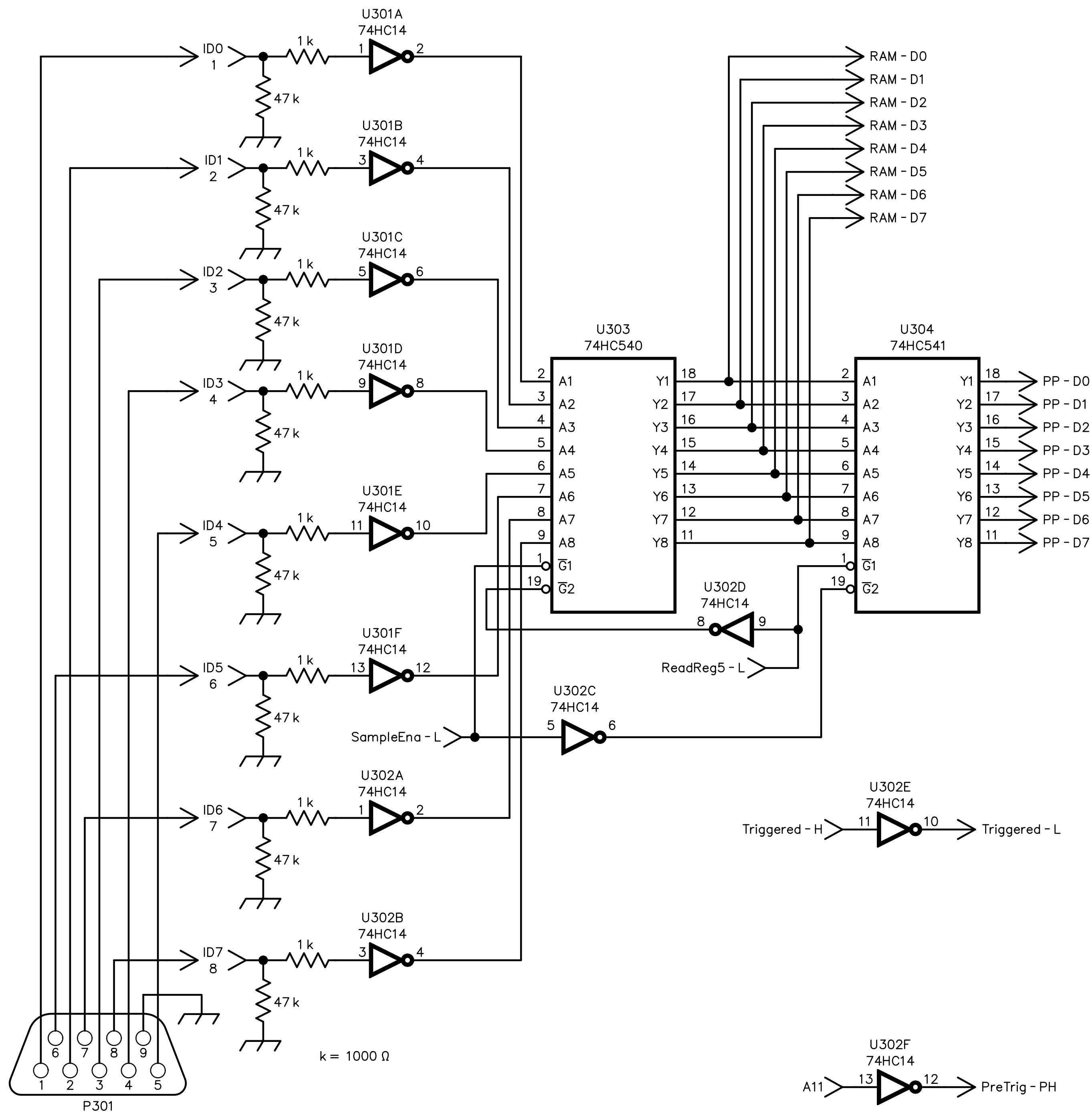


Fig 4—Input circuit schematic diagram.

See the timing diagram, Fig 9.

The Trigger State selection is accomplished using two control bits for each input. The least-significant bit, bit 0, is used to define the trigger input level as being either a “1” or a “0.” The most-significant bit, bit 1, is used to enable or disable the input signal to the comparator. Table 1 is a truth table describing the operation. Please note that the Trigger circuit is synchronous. This means that the trigger event will only be detected at the rising edge of the sample clock.

Construction

I have not presented construction details because individual builders may choose among several construction methods. Nonetheless, here is some general information on the circuit as I built it: Three of the circuits were built on daughter boards that plug into the main board via pin-strip headers and sockets. I chose this method for two reasons. (1) The largest circuit board on hand was not large enough to hold all the ICs. The one in the parts list would also require the use of daughter boards. Larger ones are available from the same vendor. (2) Those circuits that are on plug-in boards can be easily replaced in order to experiment with modified designs.

Several construction details have been deliberately omitted, such as the power supply, cable to the PC, probe cable, mounting methods, etc. Builders should determine the best methods for their own needs.

LA Front Panel Controls

Refer to Fig 10 while reading these topics.

Starting Sample Number

Enter the position of the desired sample number at the left-hand edge of the display window.

Ending Sample Number

Enter the position of the desired sample number at the right-hand edge of the display window. Note: The number of samples viewed at any one time is limited to the range: $50_{10} \leq$ number of samples $\leq 30{,}000_{10}$.

Cursor

Move the cursor to the desired sample. The cursor may also be moved by using the mouse to click on a location within the display area.

The above controls allow the user to type the desired number into the appropriate input box. However, the

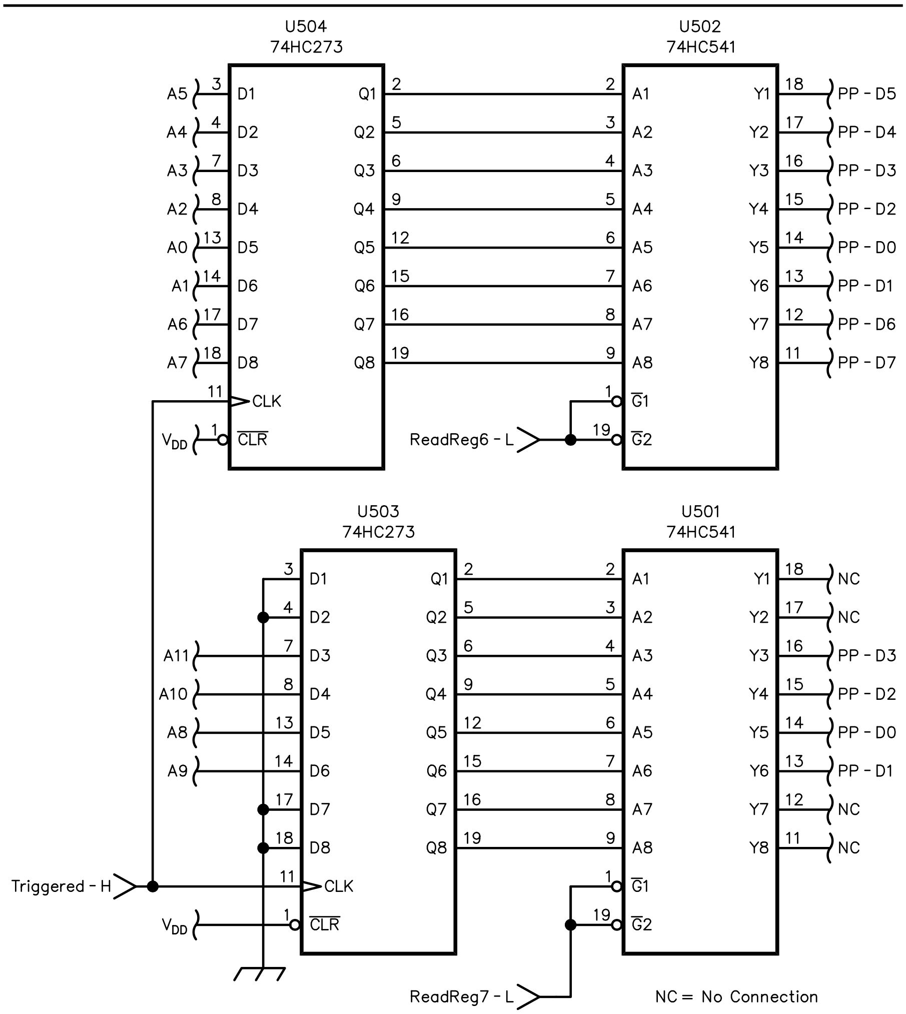


Fig 5—Pre-trigger circuit schematic diagram.

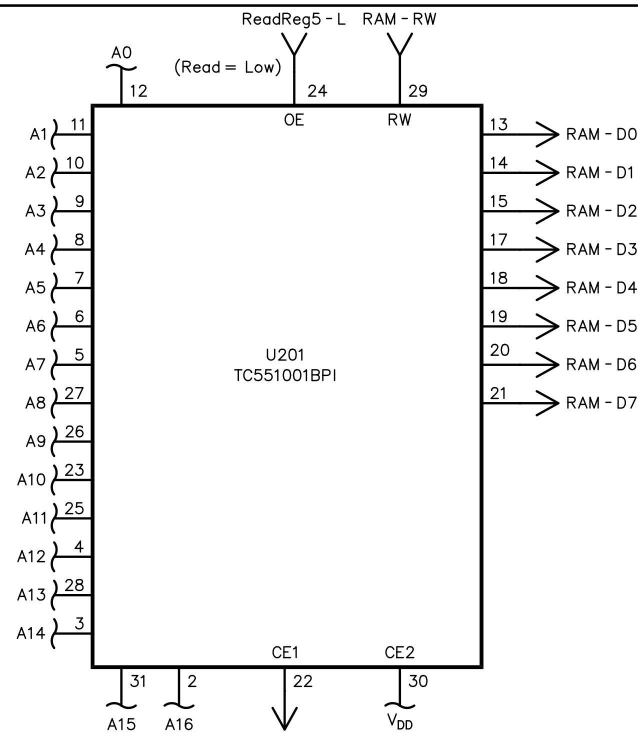


Fig 6—RAM circuit schematic diagram.

entered value does not take effect until the **ENTER** key is pressed. This mode of operation was chosen in order to disallow what might be illegal values generated by partial entries. In order to give the operator some visual feedback, the background color of the box changes until the **ENTER** key is pressed.

Slide Up

Move the display window to the set of samples that begin at 90% of the ending sample number, keeping a constant number of samples in the display window.

Slide Down

Move the display window to the set of samples which begin at 10% of the ending sample number, keeping a constant number of samples in the display window.

Cursor Memory Check Boxes

These boxes allow the user to store the current cursor location. As the cursor is moved, the Distance-to-Cursor values are updated. This allows the user to make time-difference measurements.

Draw

Under most circumstances, the program knows when to draw the waveforms. However, the waveforms develop "holes" when the cursor is moved. This control is then used to redraw the waveforms. I do not currently know of any method within *Visual Basic* that will allow the cursor to move within the display window without leaving these holes. *Warning*—The time it takes to display the window is directly proportional to the number of active channels and the number of samples to be displayed. It can take several seconds to draw one signal if there are 30,000 samples. This is processor dependent—I currently have a 166-MHz '686.

Trigger

This signals the hardware to acquire another sample set.

Notes on the Free Version of the Program

Only two channels are enabled: D0 and D1 of the printer port. No external hardware is supported. Triggering occurs after the Trigger control has been pressed and a change is detected in either or both signals.

Larry Cicchinelli, K3PTO, was first licensed in 1961. He received a BSEE

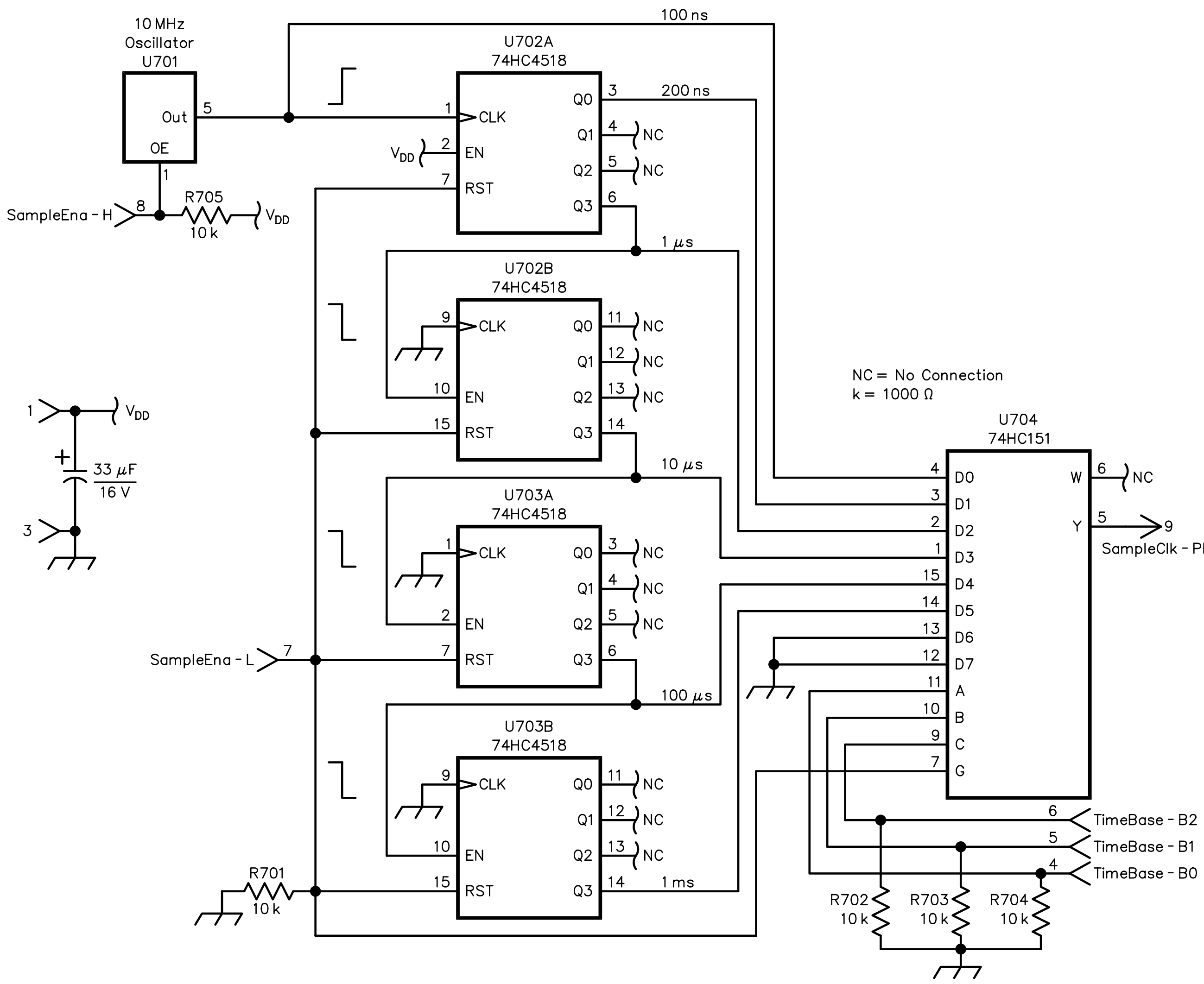


Fig 7—Time Base schematic diagram.

from Drexel Institute of Technology in 1969 and a MS in engineering science from Penn State in 1981. Larry has worked at Visteon Automotive Systems (Ford Motor Co.) since 1967. Most of his work (the first 25 years) has been the design and development of automatic test equipment and automotive electronic subsystems. His other activities include church and reading. Larry also wrote "A Dip Meter with Digital Display," QEX, Oct 1993, pp 19-21.

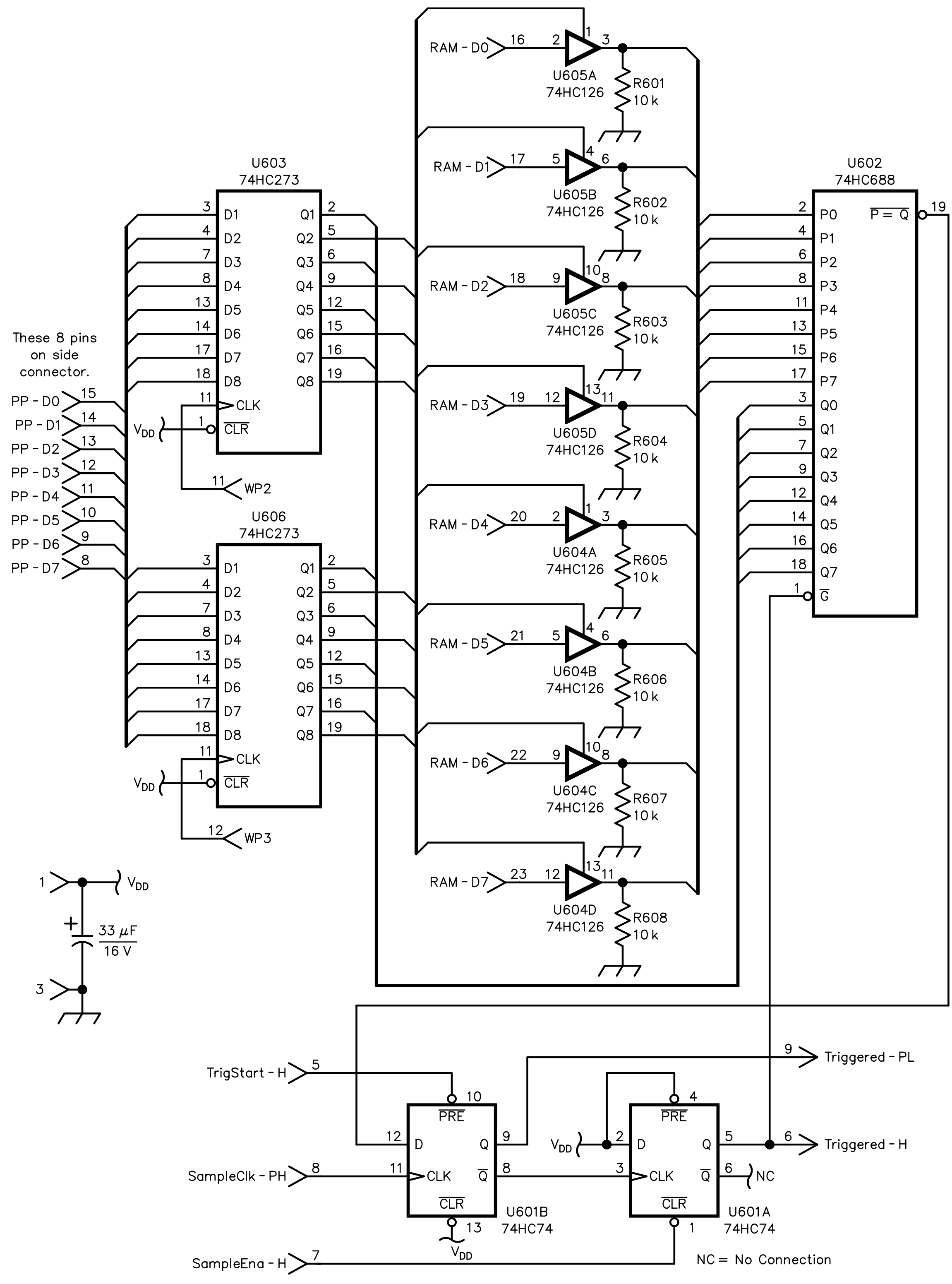


Fig 8—Trigger circuit schematic diagram.

```
Addr Count        0   1          4095            N >< PreTrigCount >< 4096
SampleClk      _.._____========___...._________========___...._________========__________===
SampleEna-H    ______=====================================================
AddrReset-PL   ==_====================================================
PreTrig-PH     ________________________________========___________________________________
TrigStart-H    _________________________________===================================
Trigger Event  XXXXXXXXXXXXXXXXXXXXXXXXXXXXXXXXXXXX____XXXXXXXXXXXXXXXXX
Triggered-PL   ======================================________==========
Triggered-H    ________________________________________________________=================
```

Fig 9—Trigger-circuit timing diagram.

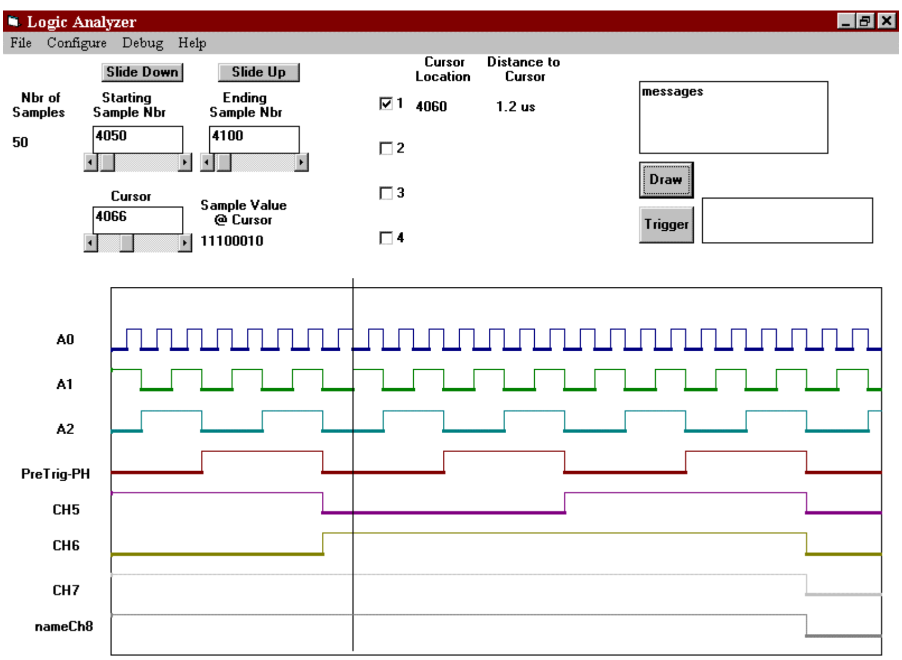


Fig 10—Logic Analyzer main screen capture.
The symbols used in the diagram: "_" is 0 V, "=" is 5 V, "X" is "don't care;" ".." indicates time duration.

Appendix: EPP-Mode Printer Port Usage for the Logic Analyzer

Data Port (read and write): Base + 0
  bits 0 .. 7: PPDR-Bit0 .. 7 = pins 2 .. 9
Status Port (read only): Base + 1
  bit 6: AddressDone-H = pin 10
  bit 7: Triggered-L = pin 11
Control Port (write only): Base + 2
  bit 0: strobe = pin 1
  bit 1: RegSel-0 = pin 14
  bit 2: RegSel-1 = pin 16
  bit 3: RegSel-2 = pin 17
  bit 5: Enable bi-directional port
(0 = write,1 = read)—this signal is *not* available on the printer connector.
The design implements two sets of hardware registers that allow the Data Port to be used to both read and write, based on the Register Select bits:
Register 0 Write: Control
  bit 0: SampleEna-L, SampleEna-H
  bit 1: AddrReset-PL
Register 1 Write:
  TimeBaseSel-0 .. TimeBaseSel-4
Register 2 Write:
  TrigSel-00 .. TrigSel-31
Register 3 Write:
  TrigSel-40 .. TrigSel-71
Register 4 Read: not used
Register 5 Read:
  ReadReg5-L:
  stored data from RAM
  (strobe is used to advance the address)
Register 6 Read:
  ReadReg6-L:
  low eight bits of Pre-Trigger address
Register 7 Read:
  ReadReg7-L:
  bits 0 .. 3: upper four bits of
Pre-Trigger address

Parts List

Vendors: M = Mouser; RS = Radio Shack; DK = DigiKey; J = Jameco

Part	*Qty*	*Vendor*	*Part Number*	*Part*	*Qty*	*Vendor*	*Part Number*
9-Pin D subminiature, chassis male	1	DK	A2043-ND	74HC126	2	M	511-M74HC126
9-pin D subminiature, female cable	1	M	156-1309	74HC14	3	M	511-M74HC14
9-pin D subminiature hood	1	RS	276-1539	74HC74	1	M	511-M74HC74
25-pin D subminiature, male chassis	1	M	571-7479122	74HC151	1	M	511-M74HC151
8-pin socket	7	M	571-26403573	74HC155	1	M	511-M74HC155
14-pin socket	1	M	571-26404633	74HC157	1	M	511-M74HC157
16-pin socket	10	M	571-26403583	74HC161	5	M	511-M74HC161
20-pin socket	11	M	571-26404643	74HC4518	2	M	511-M74HC4518
32-pin socket	1	M	571-26440183	74HC273	6	M	511-M74HC273
Housing, 36 pin	3	J	SCH36	74HC540	1	M	511-M74HC540
Resistor Net, 47 kΩ	1	M	266-47K	74HC541	3	M	511-M74HC541
Resistors, assorted	1	RS	271-312	74HC688	1	M	511-M74HC688
SW, DPDT-2P	1	J	21979	47-µF, 10-V capacitor	4	M	140-XRL10V47
50' wrapping wire	1	RS	278-501	0.2-µF capacitor	20	M	21RZ320
Box, Aluminum, 10×6×3.5 inches	1	DK	L108-ND	PC Board, 4.5 × 8"	1	M	574-3677-6
Test Clips, set of 6	2	DK	923848	PC board	3	RS	276-168B
TC551001BPI (1MB RAM)	1	DK	TC551001BPL-70-ND	Pin-strip header, right angle	3	M	517-5111TN
10 MHz oscillator	1	DK	SE1208-ND	Pin-strip header, straight	1	J	SMH36
				Pins for SCH36	100	J	FCH1

Notes

1. Use of headers, housings and PC boards depend on the construction techniques chosen by the builder.
2. Many parts can be obtained from several vendors

A High-Performance Homebrew Transceiver: Part 4

Here is a complete description of the AF board.

By Mark Mandelkern, K5AM

Designing and constructing the AF board of a radio is a relatively easy task. Op-amp circuits are perfectly predictable, special features can be added with little effort; signal levels are high and easily measured with simple equipment. The AF board in this radio contains the BFO amplifier, product detector, noise limiter, speech amplifier, balanced modulator and CW sidetone oscillator.

Part 1 gave a general description of the K5AM homebrew transceiver, built for serious DX work and contest operating.[1] Parts 2 and 3 described the IF and RF boards.[2, 3] This article gives circuit details for the AF board. A general description of this board was given in Part 1; the board is shown here in Fig 1. The block diagram in Fig 2 shows the arrangement of the various stages.

[1]Notes appear on page 56.

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Las Cruces, NM 88005
k5am@roadrunner.com

Features

Some of the special features of the AF board are:

- High-pass transmit audio filter to eliminate hum.
- 60-Hz receive audio filter to eliminate hum.
- Electronic attenuators in all circuits with front-panel controls to avoid routing audio signals to the panel.
- Class-A audio output stage with only 0.3% total harmonic distortion.
- Noise limiter for reduction of atmospheric static.

Circuit Description: BFO and Carrier Amplifiers

Fig 3 gives the schematic for the stages used to provide LO injection power for the product detector and balanced modulator and that provide a carrier for CW operation. Special effort was made to prevent carrier leakage into the IF board during SSB operation. Both a switched MOSFET and a diode switch are used to gate the carrier. The total residual hum, noise and carrier on the transmitted SSB signal are more than 65 dB down.

Receiver Circuits

Fig 4 is the schematic of the stages used to process received audio signals. Signals arrive from the IF board at terminal σ9R1 and are fed directly to the product detector. Measuring LO injection at a DBM (doubly balanced mixer) is difficult with simple equipment. The diodes in the DBM cause distorted scope patterns at the LO port that are not easy to interpret as to power level. The 47-Ω resistor at the product detector's LO port helps isolate the port and allows meaningful readings at the test point indicated. The 2-V(P-P) level specified at the test point is estimated to correspond to a +4 dBm LO injection level. This is appropriate, although the device is rated at +7 dBm (see the discussion in Part 3).

A special low-noise op amp was selected for the first AF amplifier. Care was taken to install this low-noise amplifier immediately adjacent to the product detector to avoid hum pick-up on a connecting cable. The gain of the low-noise op amp is adjusted for 300 mV (P-P) at terminal σPD. The AM and FM detector circuits are adjusted similarly and the three JFET gates are used to select the appropriate detector.

Noise Limiter

As hams painfully know, noise blankers do not blank static caused by atmospheric thunderstorms or corona noise produced by high-tension power lines. In the eternal struggle to copy DX signals (no matter how weak) under noisy conditions (no matter how severe) hams have always been eager to try any conceivable noise-reduction method.

In times past, when AM was the primary mode for voice work on the ham bands, receivers had noise limiters. One common type was the shunt limiter, a diode clipper applied to the detected AF signal. In some radios, the clipping level was adjustable. Other radios had an automatic noise limiter that used the detected carrier—a dc signal sometimes also used to drive the S-meter—to establish an appropriate bias level for the clipping diodes.

Modern commercial receivers have no noise limiters. Many hams, especially 160-meter DX hounds, have used diodes on the headphone line to clip the noise. This can be very effective. A disadvantage is that the degree of clipping depends on the AF voltage level at the headphone jack, the headphone impedance and the AF gain (AFG) setting selected by the operator. The main disadvantage is that to obtain a fair amount of clipping, the receiver AF output level must be much higher than normal; this is likely to cause distortion in the receiver AF circuits, degrading signal intelligibility.

These problems are avoided by putting the noise limiter inside the radio, ahead of the **AF GAIN** control. Clipping at low AF levels is a simple matter. With no AM carrier to set the bias level, we cannot expect an automatic noise limiter, but we do want an adjustable noise limiter. We need a panel control to set the clipping level according to conditions. In Part 1, Fig 3, the **NOISE LIMITER** control is just below the meters, third knob from the left. The limiter is quite effective, especially with atmospheric static for weak low-band DX signals.

"Limiter" is a good term. It reminds us that the circuit cannot eliminate the noise, as a good blanker can under certain conditions, but that it only limits the noise to a level set by the operator. This level is usually the peak level of the desired signal, so that the noise and the signal are evenly matched in a fair contest to register in the operator's brain.

Receiver sensitivity measurements are based on the minimum discernible signal (MDS) specification. This means that the (S + N)/N is approximately 3 dB. Experienced DX operators know, however, that this ratio is no "minimum signal"—it is "arm-chair copy," or in other words, "loud." DX operators can copy CW signals that are well below the MDS. Thus, the limiter can turn barely detectable signals into reasonably good copy.

Don't expect the noise limiter to eliminate thunderstorms! Its effectiveness is only apparent on a narro range of weak DX signals, those within a few decibels of the ambient noise level. When signals fade below that level, nothing will help; when signals build up a few decibels, a noise limiter won't be needed. Still, most new countries added to an oper-ator's 160-meter DXCC list involve signals within that narrow range.

The noise limiter is normally used with the AGC on. This is the safest operating method (see Part 1, pp 21-22). Receiver gain is enough that ambient antenna noise on any band activates the AGC system. This insures that the AF level into the noise limiter is essentially constant. The resulting headphone volume is thus also constant. Signals do not "jump out the noise," as is often said about receivers reported to be "quiet." Such receivers merely have inadequate

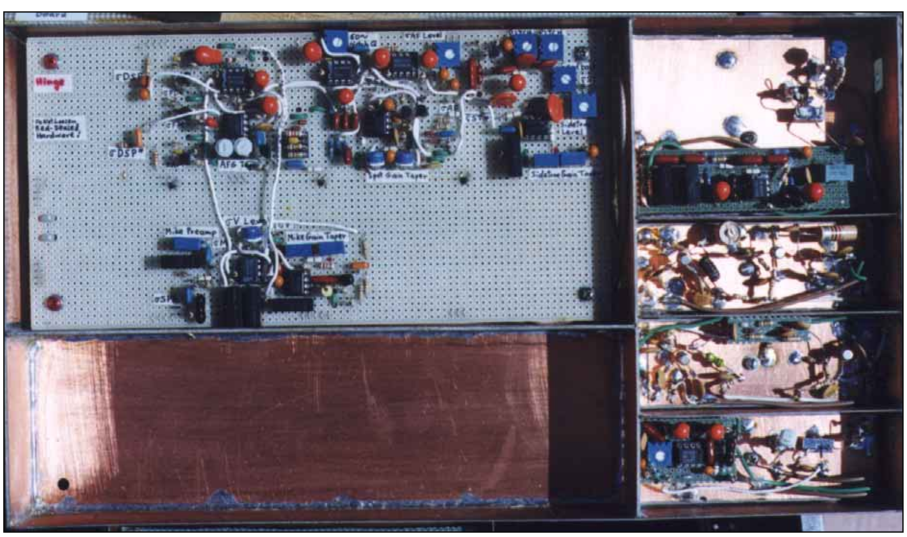
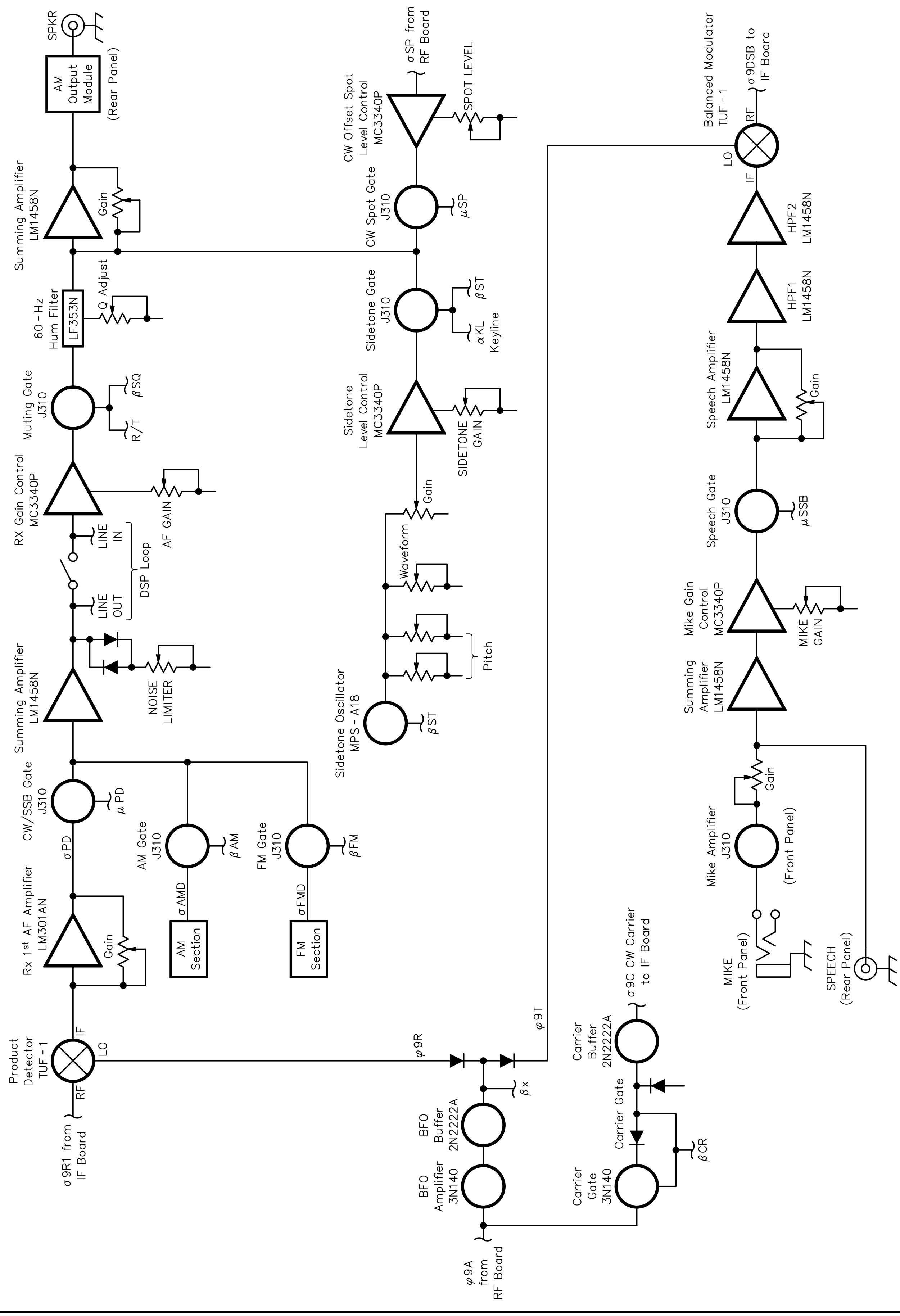


Fig 1—Top view of the AF board in the K5AM homebrew transceiver. At right, from the top, are the balanced modulator, the BFO amplifier, the carrier amplifier for CW and the product detector. The two-stage transmit high-pass AF filter is included in the balanced-modulator compartment, eliminating any possibility of hum pick-up on a connecting cable. Similarly, the first receiver AF amplifier, a special low-noise op amp, is included in the product-detector compartment. All other AF circuits are in the large compartment at the upper left; the lower left compartment is for the AM/FM circuits. (See Note 8.)

Fig 2 (see right)—AF board block diagram. Signals from the IF board arrive at terminal σ9R1. After detection, amplification and filtering, the audio signal travels from output terminal σAF to the audio-output module on the rear panel. For transmitting SSB, there are dual speech inputs: a MIKE jack on the front panel and a line-level jack on the rear panel. The DSB output is at terminal σ9DSB leading to the IF board, the SSB filters and RF speech clipper. Potentiometers labeled in all capital letters are front-panel controls; others are circuit-board trimpots for internal adjustment. An explanation of the terminal designators is given in Part 2, Table 1. The control lines are provided by the logic board.



SPKR
AM Output Module
(Rear Panel)
Summing Amplifier LM1458N
Gain
60 - Hz Hum Filter
LF353N
Q Adjust
Muting Gate J310
βSQ
R/T
RX Gain Control MC3340P
AF GAIN
LINE IN
LINE OUT
DSP Loop
Summing Amplifier LM1458N
NOISE LIMITER
CW/SSB Gate J310
μPD
σPD
AM Gate J310
βAM
σAMD
AM Section
FM Gate J310
βFM
σFMD
FM Section
Rx 1st AF Amplifier LM301AN
Gain
Product Detector TUF - 1
IF
LO
RF
σ9R1 from IF Board
φ9R
βx
BFO Buffer 2N2222A
BFO Amplifier 3N140
φ9A from RF Board
φ9T
Carrier Buffer 2N2222A
σ9C CW Carrier to IF Board
Carrier Gate
βCR
Carrier Gate 3N140
σSP from RF Board
CW Offset Spot Level Control MC3340P
SPOT LEVEL
CW Spot Gate J310
μSP
Sidetone Gate J310
βST
αKL Keyline
Sidetone Level Control MC3340P
SIDETONE GAIN
Gain
Waveform
Pitch
Sidetone Oscillator MPS - A18
βST
Balanced Modulator TUF - 1
σ9DSB to IF Board
RF
LO
IF
HPF2 LM1458N
HPF1 LM1458N
Speech Amplifier LM1458N
Gain
Speech Gate J310
μSSB
Mike Gain Control MC3340P
MIKE GAIN
Summing Amplifier LM1458N
Gain
Mike Amplifier J310
(Front Panel)
MIKE (Front Panel)
SPEECH (Rear Panel)

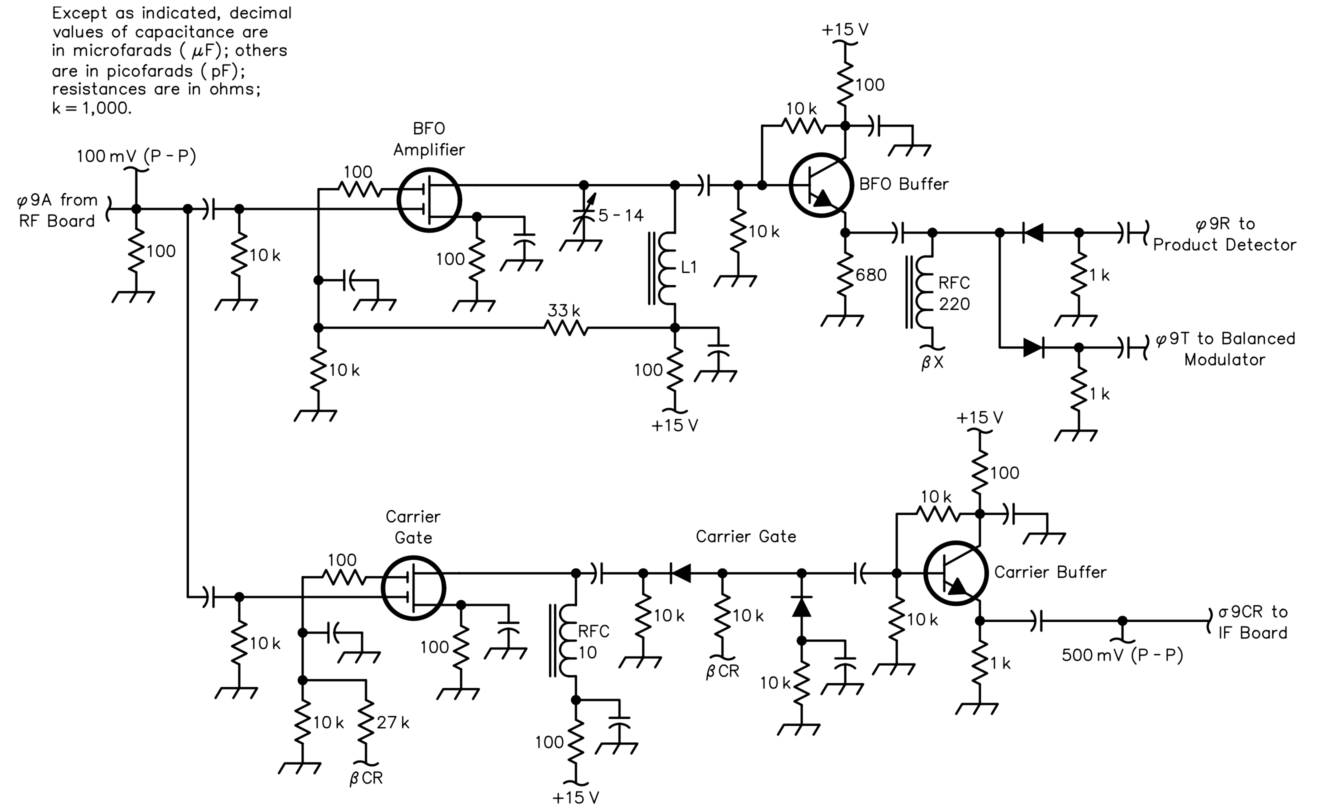


Fig 3—BFO and carrier-amplifier schematic diagram. Except as noted, each resistor is a 1/4-W, carbon-film type. All trimpots are one-turn miniature types, such as Bourns type 3386; Digi-Key #3386F-nnn. The unmarked coupling and bypass capacitors are all 10 nF disc ceramic types. Also, each control and power terminal has a bypass capacitor that is not shown. Electrolytic capacitors have 25-V ratings and values given in microfarads; those less than 100 µF are tantalum types. Capacitors labeled "s.m." are silver micas, with values given in picofarads. The values of RF chokes (RFC) are in microhenries. Part-supplier contact information appears in Notes 9, 10, 11 and 12.

The MOSFETs are all small-signal, VHF dual-gate types. Type 3N140 is used here, but any similar type may be substituted. Replacement-type NTE 221 is also available. The JFETs are J310s. Except where otherwise indicated, the diodes are small-signal silicon types, such as 1N4148; the bipolar transistors are type 2N2222A. Except as noted, the op amps are sections of LM1458Ns. Unless shown otherwise, each op amp is powered by the +15 V and –15 V rails. Not shown is the filtering at each op amp power terminal. This varies with the application. The minimum is a 100-nF, monolithic ceramic bypass. More sensitive and low-level circuits also use a 100-Ω isolation resistor and a 10-47 µF tantalum electrolytic at the terminal.

Potentiometers labeled in all-capital letters are front-panel controls; others are circuit-board trimpots for internal adjustment. The control lines are provided by the logic board. The signal levels indicated at various test points are scope readings.

L1—FT-37-61 ferrite toroidal core, 14 turns of #26 AWG enameled wire.

AGC systems; they are not necessarily "quiet" with regard to sensitivity or noise figure. Listening for weak signals in the noise on such receivers requires one to set the gain too high, creating an aural hazard when a strong signal suddenly appears. The flat AGC characteristic in this receiver allows the noise-limiter clipping level to be easily set. If "noise-jumping" signals are desired at other times, the IF GAIN control may be used to raise the AGC threshold (see Part 1, pp 21-22).

The noise-limiter circuit is shown in Fig 5. Op amp U1 raises the receiver AF level from 3 V to 14 V (P-P), while op amp U2 reduces it to the former level after the clipper. When clipping is used, however, the AFG control will need to be increased. With both controls on the

Fig 4—(see right) Receiver AF circuit schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. Capacitors labeled simply "1" are 1-µF monolithic ceramics (see Note 10). The DBM used as a product detector may be obtained in small quantities directly from the manufacturer (see Note 12). Signal levels indicated on the diagram are given with the AF GAIN set for 2 W output with an 8-Ω resistive load at the speaker jack.

C—Polypropylene capacitor, 2.7 nF, 2% tolerance. Panasonic #ECQ-P1H272GZ, Digi-Key #P3272. Where 2C is indicated, use two of these in parallel; this ensures the closest match.

C1—Polypropylene capacitor, 1 nF. Panasonic #ECQ-P1H102GZ, Digi-Key #P3102.

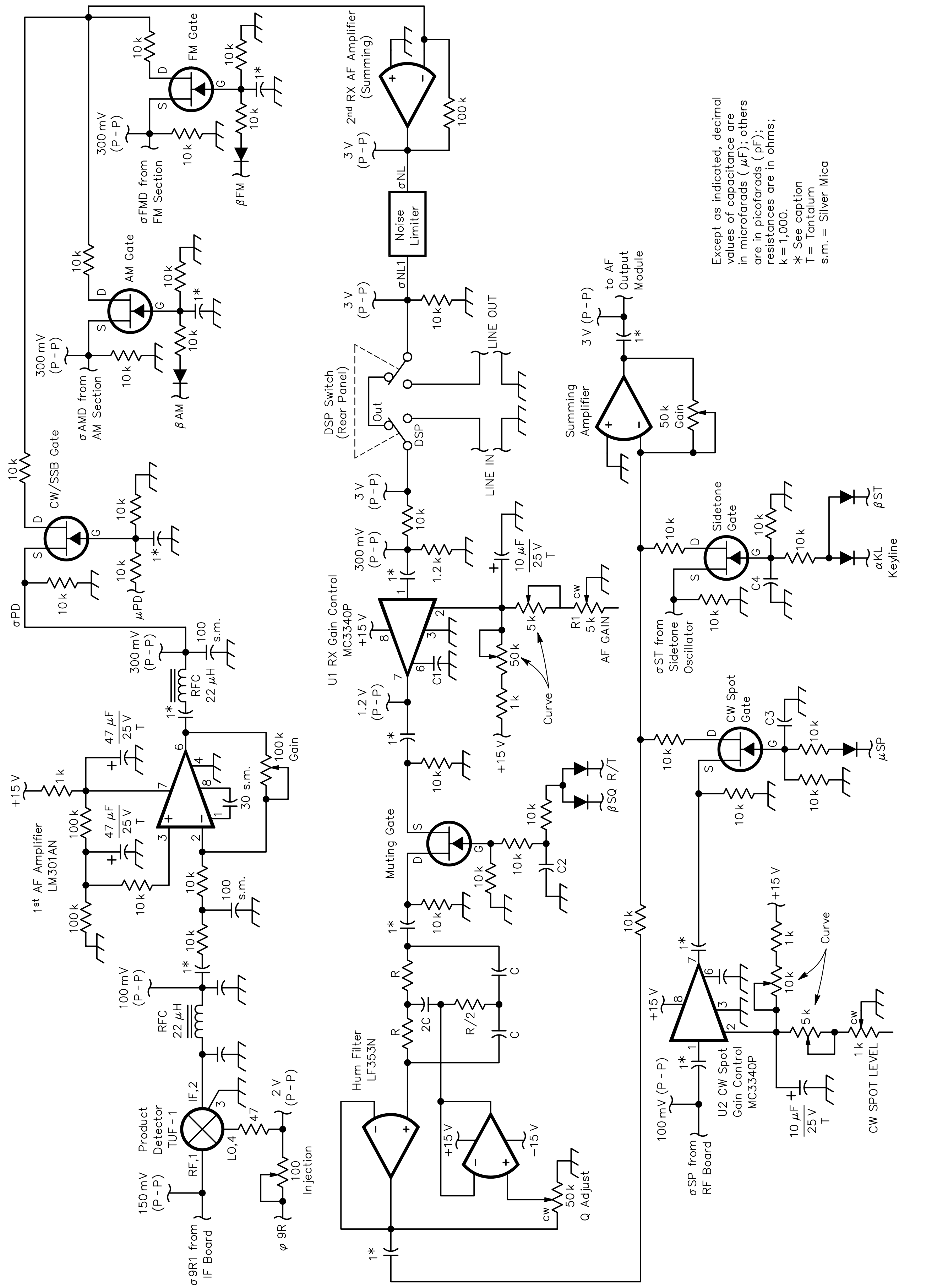
C2—Polypropylene capacitor, 470 nF. Panasonic #ECQ-P1H474GZ, Digi-Key #P3474.

C3-C4—Polypropylene capacitor, 220 nF. Panasonic #ECQ-P1H224GZ, Digi-Key #P3224.

R—Metal-film resistor, 1 MΩ, 1% tolerance. Digi-Key #1.00MXBK. Where R/2 is indicated, use two of these in parallel; this ensures the closest match.

R1—AFG potentiometer, 5 kΩ. This has been replaced five times in the last eight years. The current resident, a 2-W wirewound type found at a hamfest, seems headed for a long life.

U1-U2—Electronic attenuator, Motorola MC3340P; discontinued. Available as replacement component NTE #829, Mouser #526-NTE829 (Mouser). A datasheet is available on the Web: http://www.mot-sps.com/books/dl128/pdf/mc3340rev6.pdf. Omitted from the schematic is the filtering at the +15-V supply terminal: a 100-Ω isolation resistor, a 100-nF monolithic ceramic capacitor and a 47-µF tantalum electrolytic capacitor at pin 8.


σ9R1 from IF Board
150 mV (P - P)
Product Detector TUF-1
RF,1
LO,4
IF,2
φ 9R
100 Injection
2 V (P - P)
RFC 22 μH
100 mV (P - P)
1st AF Amplifier LM301AN
+15 V
47 μF 25 V T
100 k Gain
30 s.m.
100 s.m.
300 mV (P - P)
σPD
μPD
CW/SSB Gate
σAMD from AM Section
AM Gate
βAM
σFMD from FM Section
FM Gate
βFM
Hum Filter LF353N
Q Adjust
Muting Gate
βSQ R/T
U1 RX Gain Control MC3340P
Curve
R1 5 k AF GAIN
10 μF 25 V T
1.2 V (P - P)
3 V (P - P)
DSP Switch (Rear Panel)
Out
DSP
LINE IN
LINE OUT
Noise Limiter
σNL1
σNL
2nd RX AF Amplifier (Summing)
σSP from RF Board
U2 CW Spot Gain Control MC3340P
CW SPOT LEVEL
CW Spot Gate
μSP
σST from Sidetone Oscillator
Sidetone Gate
αKL Keyline
βST
Summing Amplifier
50 k Gain
3 V (P - P) to AF Output Module
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
* See caption
T = Tantalum
s.m. = Silver Mica

front panel, this is convenient; the clipping level is normally set according to conditions and not changed until the thunderstorm is over.

The clipping level is set by the **NOISE LIMITER** panel control. Voltage follower U3 is required to provide a low-impedance negative bias source for the negative-clipping diode. Inverter U4 tracks U3 and provides reverse bias to the positive-clipping diode of equal magnitude but opposite polarity. This insures that the positive and negative halves of the signal waveform are clipped equally. When the panel control is fully clockwise, the diodes have no reverse bias; each clips to 0.7 V. Thus the 14-V (P-P) level allows a maximum clipping depth of 20 dB. With the control fully counterclockwise, the diodes are completely reverse-biased, and the noise limiter is effectively off.

When noise limiting is needed for CW, the control is usually set near maximum; distortion on CW signals is of no consequence. For SSB, the full setting causes noticeable distortion, but signals are very intelligible. For less distortion, retard the control somewhat. With 6 dB of clipping, the distortion is barely noticeable. At those times when signals can be heard through the static without the limiter, use of the limiter makes listening far more pleasant and reduces fatigue. Also, the limiter can be left at 3-6 dB at all times; it won't noticeably effect the signals, but will reduce the noise from key clicks and splatter.

When a DX signal is too weak to hear during a static crash, the noise limiter still serves a valuable function and may enable an otherwise impossible contact. It saves the operator's ears! This is both a long-term safety feature and a short-term operating aid. An operator's hearing ability is depressed after some minutes of listening to loud noise. With the AFG high enough to hear a weak station between static crashes, one's hearing threshold may be degraded within ten minutes. On the other hand, if the gain is low enough so that static crashes don't cause desensitization, a weak station won't be heard. The solution to this dilemma is an adjustable noise limiter.

The noise-limiter schematic diagram is separated from the main diagram so that it might be built and installed in other receivers. Details and pin-outs are given. If the receiver AF level at the point of insertion is other than that specified the gain of the input and output amplifiers may be changed easily.

This noise limiter has been very helpful in DX work. One can only wonder: "Why don't all receivers have noise limiters?"

DSP Loop for External Filter

The DSP loop was added only a few years ago to accommodate a Timewave Model DSP-9+ filter. The loop is positioned before the **AF GAIN** control. This is done so that the AFG knob on the radio can be used independently of the filter. This avoids the usual cumbersome arrangement wherein the AFG on the radio must be continually readjusted for the proper input level and the AFG control on the external filter used to control headphone volume. This is the main reason for adding the DSP loop. There are also other problems:

- There is a 6-dB loss in the external filter between the line input and line output jacks.
- Signal levels are low and hum can be introduced.
- External devices installed on the speaker line can suffer from annoying RFI during transmissions.

The solution is to raise the AF level 20 dB for the loop and use an external amplifier to compensate for the loss in the filter. A simple 20-dB pad in the radio at the loop's return jack restores the AF signal to the original level.

The external amplifier is shown in Fig 6. It has controls to set the proper levels both to and from the filter. These need be set only once: The filter can be switched in or out at any time with no readjustment of receiver, amplifier or filter controls. This procedure eliminates the need for any modifications to the external DSP filter and allows adjustment for a variety of different external units. A switch on the rear panel of the radio cuts the loop out of the circuit. This avoids the need for a patch cable when the external filter is not connected. When connected, the filter is switched in and out by its own controls.

Electronic Attenuators

To avoid routing audio signals to front-panel controls and possible hum pickup, electronic attenuators are used for the **AFGAIN** function and all other front-panel audio level controls. The attenuators [MC3340Ps —*Ed.*] provide a 60-dB range, with a pin 2 control voltage range of 3-6 V. The maximum

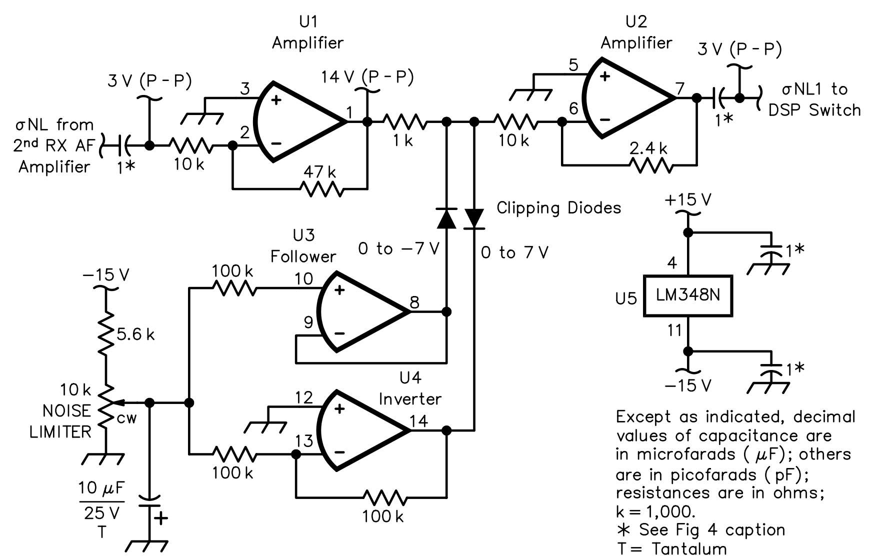


Fig 5—Noise-limiter schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. The 1-kΩ resistor at the output of U1 is needed to prevent excessive current drain on U1 when the diodes conduct. The gain-setting resistors for U4, while they might be of any (equal) value to set the gain at –1, should be of no lower value than shown. The circuit of U4 establishes a virtual ground at the inverting input; a low-value input resistor would unduly load the potentiometer circuit, resulting in less available reverse bias. At the full counterclockwise position of the control, a reverse bias of 9 V is presented to the diodes. This is enough to cut off the diodes, since the peak AF voltage is 7 V in either direction. U—Quad op amp, LM348N. Each of the four op amps is shown separately. A fifth symbol shows the power connections, but all is contained in a single, 14-pin DIP package. Note that while type LM324N is used in other parts of the radio, it is a single-supply type vulnerable to crossover distortion. It is useful for dc-control circuits because it has a wider output voltage swing, but it should not be used for AF. The dual LM1458N and quad LM348N types used on the AF board are standard low-distortion 741 types.

voltage gain of the attenuators is four. The circuits are configured for 300-mV (P-P) input, with output varying from 1.2 to 1200 mV (P-P). This same setup is used for the **MIKE GAIN**, **SIDETONE LEVEL** and **CW OFFSET SPOT LEVEL** circuits. (One attenuator for each circuit.) Each of these circuits requires only a single, unshielded dc lead to the front panel. Simplicity of wiring and avoidance of hum was well worth the extra trouble of installing the ICs.

I like the normal operating positions of all controls to be between 10 and 11 o'clock, with 40-dB of gain reduction at the full counterclockwise position and 20-dB increase when fully clockwise. Satisfying this particular demand was more difficult than expected—the characteristics of the devices varied from sample to sample. Each circuit requires two trimpots to set the desired curve, and the adjustments interact greatly. Alignment was a bit tedious, but the attenuator circuits have functioned perfectly and have required no further attention.

Muting Gate

After the attenuator is the JFET muting gate, used for TR switching and squelching. The deceptively innocent-looking capacitor, C2, in the source circuit deserves attention. Since all other circuits in the radio are quite fast, this capacitor determines receiver recovery time. Some delay is needed, or the radio will squeal in protest for a few milliseconds at the end of each transmission, or after each dit or dah when operating CW QSK. Too much capacity here will cause excessive delay and destroy the radio's break-in ability. On the other hand, too little capacity will allow clicks to be heard in the headphones. The divider for the blocking voltage also affects the timing. The circuit shown here mutes the receiver smoothly and quietly. Stations can be heard breaking in while the radio is transmitting CW at 50 wpm.

60-Hz Notch Filter

The next stage is an adjustable-Q 60-Hz notch filter.[4] It eliminates any hum, whether from internal circuits or included on received signals. The component values specified result in a calculated notch frequency of 58.9 Hz. The component manufacturers' tolerances result in a worst-case error of 1.7 Hz. This possible error is one reason for using an adjustable-Q filter: to ensure that the notch includes 60 Hz. The trimpot is adjusted to obtain a 3-dB band-reject range of about 55-65 Hz. Measured

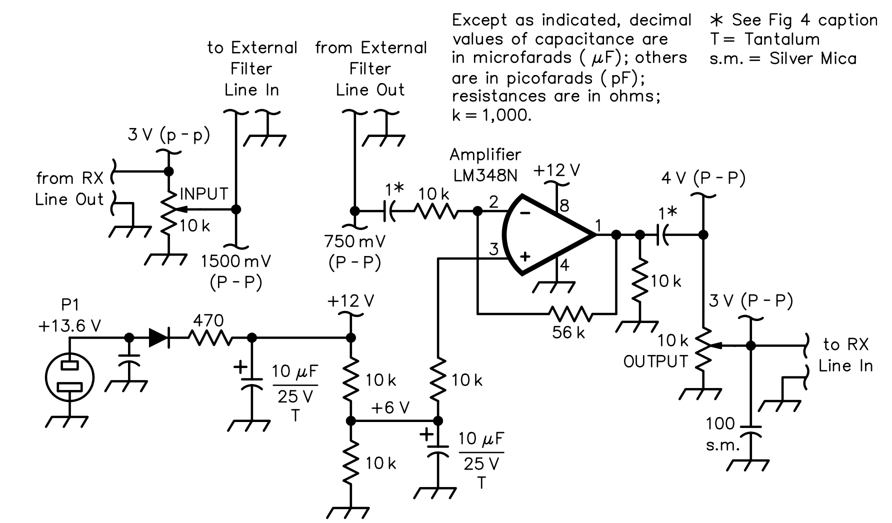


Fig 6—External-DSP-loop-amplifier schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. The amplifier may be housed in a small aluminum box. Defense against RFI from the transmitting antenna is not critical, since the amplifier and filter are positioned before the receiver muting gate.
The amplifier is powered by the same 13.6-V dc power supply used for the numerous operating-bench auxiliary gadgets. The amplifier is fitted with a 6-foot shielded power cable terminated in a two-pin Jones plug. The 13.6-V dc fused distribution boxes on the operating bench contain two-pin Jones sockets. These connectors are solid, reliable and easy to use. Since the shack has them only for this purpose, this system is safer than one that uses other common connectors, such as phono plugs.
P1—Jones plug, two-pin, Cinch #P-302-CCT, Digi-Key #CJ102P. Mating socket, Cinch #S-302-AB, Digi-Key #CJ302S.

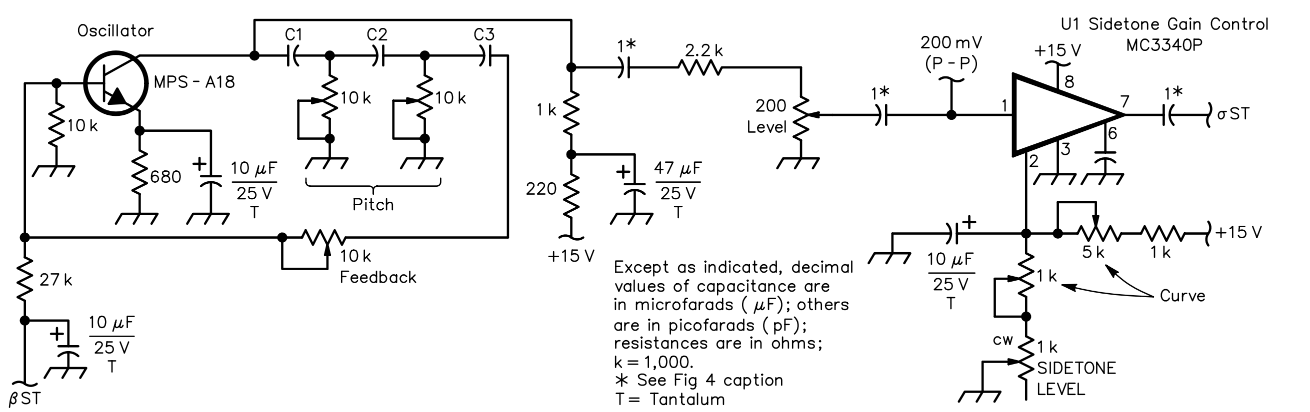


Fig 7—Sidetone-oscillator schematic diagram. For general notes on the schematics, refer to the caption for Fig 3.

C1-C3—Polypropylene capacitor, 47 nF. Panasonic #ECQ-P1H473GZ, Digi-Key #P3473.

U1—MC3340P. See caption for Fig 4.

notch depth is 50 dB.

The notch filter circuit is a modern version of the traditional "twin-T" network. The network alone has a Q of only 0.3, but with feedback derived from the upper voltage-follower op amp, the Q can be adjusted from 0.3 to more than 50. The trimpot determines the amount of feedback. The lower voltage follower is needed to drive the network from a low-impedance source so the circuit is unaffected by the resistance of the trimpot. Notch depth depends on component match (see Note 4).

The last stage is a summing op amp that combines the signals from the detectors, sidetone oscillator and CW-offset spot mixer on the RF board. The gain of the summing-amplifier stage is adjusted to provide the correct audio signal level at terminal σAF, which leads to the AF output module on the rear panel.

Sidetone Oscillator

The sidetone oscillator is shown in Fig 7. Adjustments are provided for pitch, feedback and output level. The feedback control was originally intended to allow adjustment for the minimum amount needed to sustain oscillation, thus to obtain the purest-possible sine wave. However, a pure sine wave may produce fatigue over long operating periods. The feedback control may be used to introduce some harmonic distortion, which may result in a more pleasant sound. Recall that a good violin produces many nice-sounding harmonics. When in either **TUNE** or **PULSE TUNE** mode, the sidetone oscillator is disabled.

AF Output Module

The AF-output module is mounted on the transceiver rear panel, with the output transistors using the panel as a heat sink. The circuit is shown in Fig 8. Two stages of low-pass filtering are used, each with a 3-dB cut-off frequency of 3000 Hz. The measured composite response is –3 dB at 2500 Hz and –6 dB at 3000 Hz.

Each filter stage is a unity-gain, maximally-flat Butterworth filter with:

$$Q = \frac{\sqrt{2}}{2} \qquad \text{(Eq 1)}$$

in a voltage-follower configuration.[5] This configuration is the simplest to apply with respect to component selection. It also involves the simplest formulas. One first chooses the cut-off frequency, f (in Hz), and the capacity, C (in farads). Then the resistance R (in ohms) is simply:

$$R = \frac{1}{2\sqrt{2}\,\pi f C} \qquad \text{(Eq 2)}$$

The 10-kΩ resistor at the non-inverting input does not affect the response; it serves merely to protect the op amp (see Note 5, p 126).

The output stage operates class A. The biasing is chosen for 50 mA idling current. Total harmonic distortion (including hum and noise) of the AF-output module measures 0.3% at normal levels and 1.5% at the full 2-W output. In a test involving the entire receiver, an RF signal on the 20-meter band produced total harmonic distortion (including hum and noise) of 0.6% at normal levels.

Transmit Circuits

Fig 9 is the schematic of the stages used for transmitting. The speech amplifier has two alternative inputs. The front-panel **MIKE** jack is rarely used. I don't like a clutter of cables on the operating bench or protruding plugs that impede control-knob access. I always use the rear panel **SPEECH** jack.[6] This jack provides a high-level, 600-Ω line input and accepts speech signals from the station's speech-distribution system and the digital voice keyer.[7] Running this input at a high-level minimizes hum pick-up—often a problem when digital voice recorders are used. The nominal input levels are 30 mV (P-P) at the front-panel **MIKE** jack and 300 mV (P-P) at the rear-panel **SPEECH** jack. The inverting op-amp summing stage combines the two inputs with no interaction.

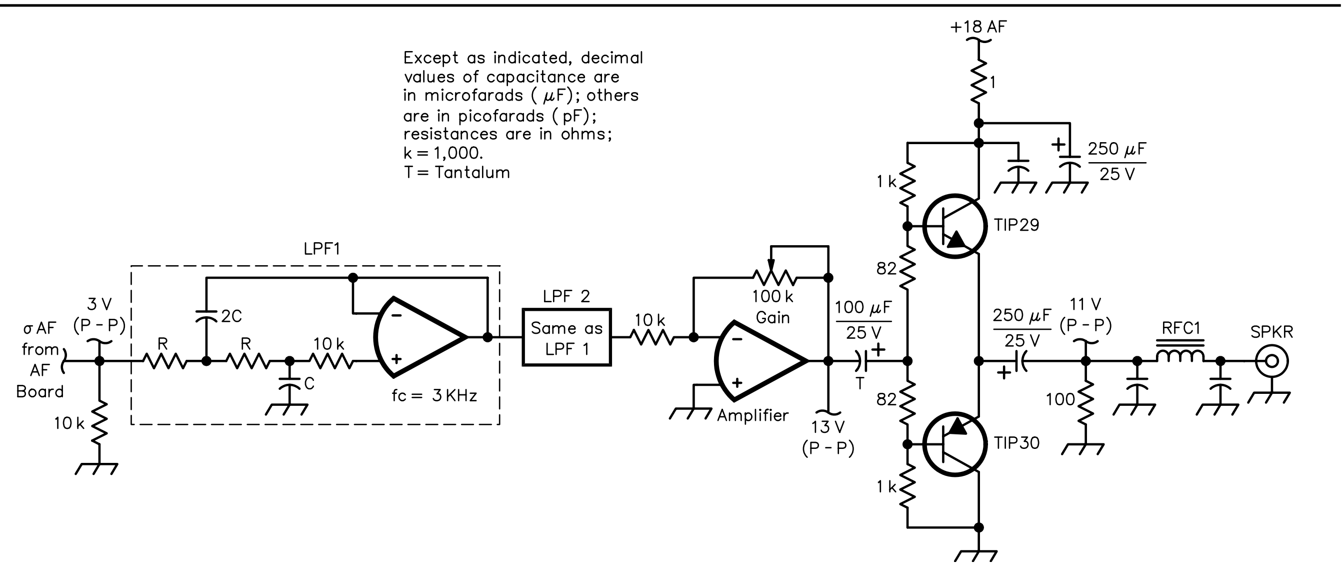


Fig 8—AF-output-module schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. The 1-Ω resistor in the +18-V supply line to the output stage is a shunt for current metering during tests. A DMM set to the millivolt range will read out in milliamps directly. These 1-Ω shunts are also included at many other points in the radio where a current measurement is required. The complementary-output transistors are available at Hosfelt.

C—Polypropylene capacitor, 1 nF, 2% tolerance. Panasonic #ECQ-P1H102GZ, Digi-Key #P3102. For 2C use two of these in parallel; this insures the closest match.

R—Metal-film resistor, 37.5 kΩ, 1% tolerance. Digi-Key #37.5KXBK.

RFC1—RF choke, 10 μH. The choke must have a dc resistance of 1 Ω or less.

The JFET microphone amplifier is located directly at the front panel on a very small circuit board attached to the jack. This minimizes hum pick-up. A single shielded lead runs from this board to the AF board. The microphone amplifier will accept high-impedance mikes; for low-impedance mikes, an appropriate load resistor must be provided externally. For my Heil HC-5 mikes, I install a 2.2-kΩ resistor in the mike plug. This arrangement provides the greatest flexibility (see Note 7). The RC network at the mike jack filters RF. Silver-mica and polypropylene capacitors are used for their low-noise characteristics.

After the summing amplifier, there is an electronic attenuator for **MIKE GAIN** control. It is adjusted as described above in the "Receiver Circuits" section. The electronic attenuator avoids the need to route speech signals to the front panel, again minimizing hum pick-up. Capacitor C2 at pin 6 of the attenuator is for high-frequency roll-off; it may be altered as desired. The value given for C2 results in a 3-dB roll-off at 5 kHz. After the attenuator is a JFET gate. This, along with switched stages on the IF board, eliminates any possibility of modulation during **CW** or **TUNE** operation. The next op amp drives the high-pass filter.

High-Pass Filter

Two stages of high-pass filtering are used, each have a 3-dB cut-off frequency of 300 Hz. The measured composite response is –3 dB at 380 Hz and –50 dB at 60 Hz. Care was taken to install the high-pass filter immediately adjacent to the balanced modulator, thus eliminating any possibility of hum pick-up on a connecting cable.

Each stage is a unity-gain, maximally-flat Butterworth filter with

$$Q = \frac{\sqrt{2}}{2} \qquad \text{(Eq 3)}$$

in a voltage-follower configuration (see Note 5). This configuration is the simplest to apply in regard to component selection. It also involves the simplest formulas. One first chooses the cut-off frequency f (in Hz) and the capacity C (in farads). Then the

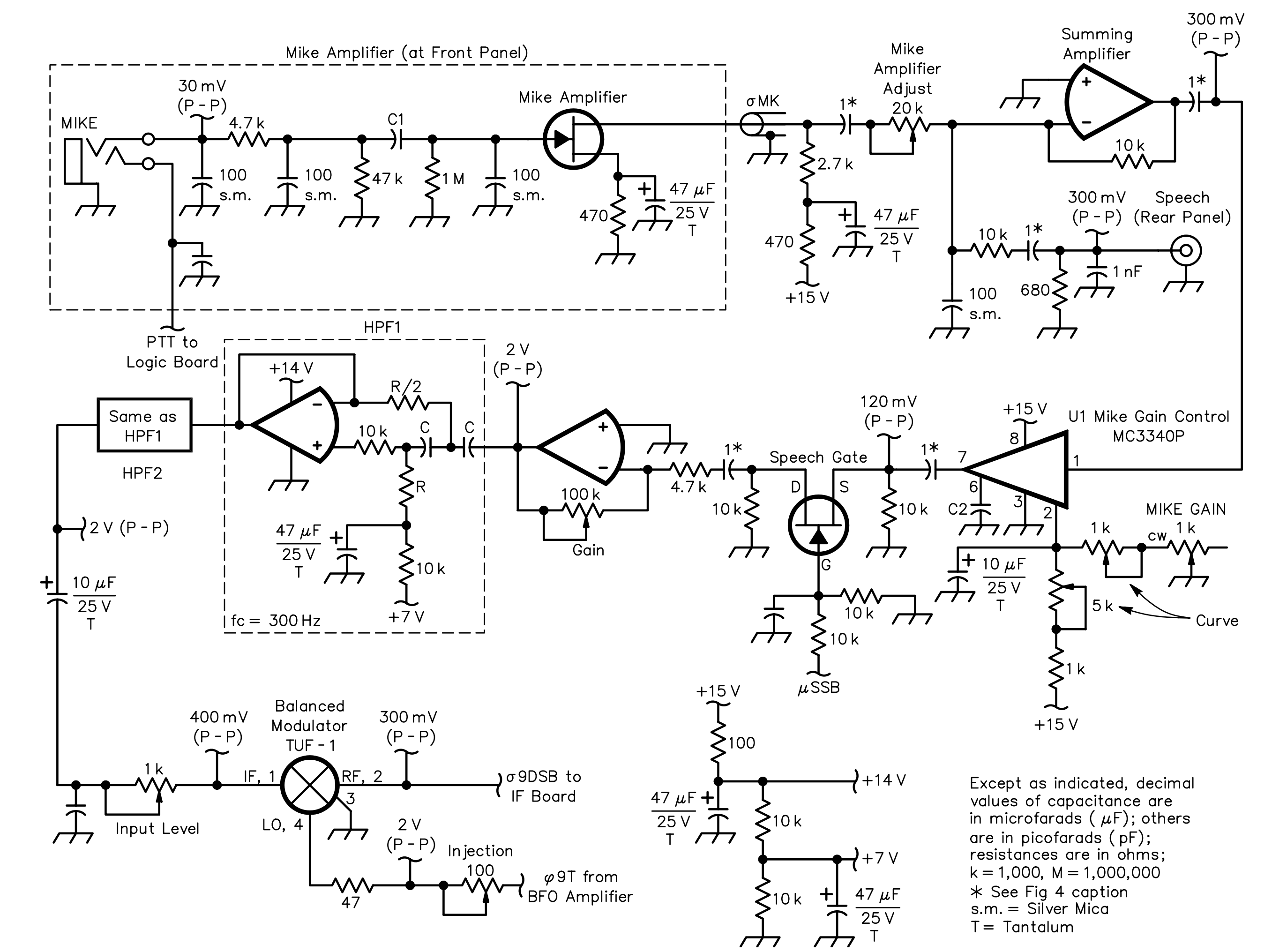


Fig 9—Transmitter AF-circuit schematic diagram. For general notes on the schematics, refer to the caption for Fig 3. The signal levels after the mike-gain attenuator are given for the normal control-knob position of 11 o'clock (see Part 1, p 21 for information about microphone calibration). The DBM used as a balanced modulator may be obtained in small quantities directly from Mini Circuits.

C, C1—Polypropylene capacitor, 10 nF, 2% tolerance. Panasonic #ECQ-P1H103GZ, Digi-Key #P3103.

C2—Polypropylene capacitor, 5.6 nF. Panasonic #ECQ-P1H562GZ, Digi-Key #P3562.

R—Metal-film resistor, 75 kΩ, 1% tolerance. Digi-Key #75.0KXBK. For R/2 use two of these in parallel; this insures the closest match.

U1—MC3340P. See caption for Fig 4.

resistance R (in ohms) is simply:

$$R = \frac{1}{\sqrt{2}\pi\, f\, C} \qquad \text{(Eq 4)}$$

With a fixed value of C, the cut-off frequency may be changed (within limits) by simply using the inverse relation between f and R. For example, with the value shown for C, a cut-off frequency of f = 200 Hz can be obtained with R = 112.5 kΩ. The 10-kΩ resistor at the non-inverting input does not affect the response; it simply protects the op amp as previously described. The 10-kΩ filtering resistor at the +7-V bias supply point also does not affect the response; the 47-µF capacitor provides a virtual AF ground for the network.

Construction

The AF board is built as shown in Fig 1. The general method of construction was described in Part 1, where the need for careful shielding and lead filtering was discussed. The power and control leads are filtered as described in Part 2. Most of the purely-AF circuits are hand-wired on perf boards. Plain copper board and truly "ugly" construction is used for all circuits containing 9-MHz signals. The board's underside is shown in Fig 10.

Test and Alignment

Normal operating levels at various points of the circuit are indicated on the schematic diagrams. The associated trimpots are adjusted to obtain the specified oscilloscope readings. There are quite a few trimpots in the radio. When building a single unit, it is often easier to install a trimpot than to individually select a component. This minor expense is justified by saving of hours of work in selecting and changing components. This is especially true when a radio is designed and built stage-by-stage. Often, a section is built without full knowledge of what lies ahead. Desired levels are apt to change in the light of final testing. None of the trimpots has required readjustment since the radio was built. Exceptions occur when modifications are introduced; then the trimpots become especially useful.

Summary

This article gives a complete description of the AF board in a high-performance homebrew transceiver. Double-balanced mixers are used for both the product detector and the balanced modulator. Special filters are included to avoid hum both in reception and in the transmitted signal. As in the remainder of the radio, flexibility and operating convenience are prime design factors.

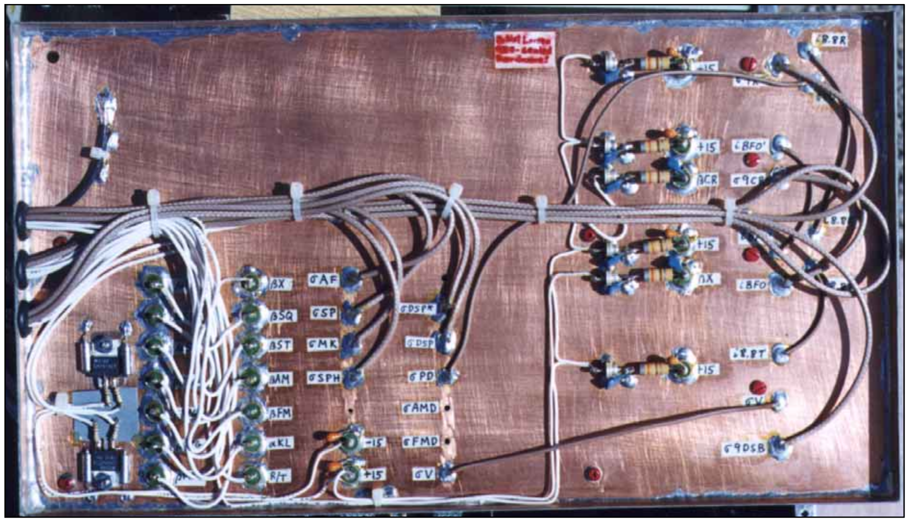


Fig 10—Bottom view of the AF board. Effective filters are installed at each terminal and coax cables are soldered directly to the double-sided circuit board (see the discussion in Part 2). To minimize connector troubles, the board is hard-wired to the radio; a 12-inch-long bundle of wires and cables allows the board to be easily lifted and serviced.

Notes

[1]M. Mandelkern, K5AM, "A High-Performance Homebrew Transceiver: Part 1," *QEX*, Mar/Apr 1999, pp 16-24.

[2]M. Mandelkern, K5AM, "A High-Performance Homebrew Transceiver: Part 2," *QEX*, Sep/Oct 1999, pp 3-8.

[3]M. Mandelkern, K5AM, "A High-Performance Homebrew Transceiver: Part 3," *QEX*, Nov/Dec 1999, pp 41-51.

[4]R. C. Dobkin, "High Q Notch Filter," Linear Brief #LB-5, in *Linear Applications Handbook*, National Semiconductor Corp, Santa Clara, California, 1980.

[5]W. Jung, *IC Op-Amp Cookbook* (Indianapolis: Howard W. Sams and Co, 1974), pp 331-333.

[6]Similarly, I don't use the front-panel **HEADPHONE** jack, but only the rear-panel **SPEAKER** jack. On the operating bench is a 40-year-old speaker/headphone switch box with a headphone-level control for equalization. This provides instant switching without fussing with the headphone plug or readjusting the AFG.

[7]M. Mandelkern, K5AM, "The AMSAFID: An Automatic Microphone Switcher Amplifier Filter Integrator Distributor," *QST*, Nov 1995, pp 47-49.

[8]Sharp-eyed readers may notice that the AM/FM compartment is empty. This only proves that this shack follows true ham-radio tradition: Nothing is ever really finished.

[9]Digi-Key Corporation, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677; tel 800-344-4539 (800-DIGI-KEY), fax 218-681-3380; **http://www.digikey.com/**.

[10]Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952; tel 800-524-6464, fax 800-524-5414; E-mail **hosfelt@clover.net**; **http://www.hosfelt.com/**.

[11]Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76063, tel 800-346-6873, fax 817-483-0931; E-mail **sales@mouser.com**; **http://www.mouser.com/**.

[12]Mini Circuits Labs, PO Box 350166, Brooklyn, NY 11235-0003; tel 800-654-7949, 718-934-4500, fax 718-332-4661; **http://www.minicircuits.com/**.

See Part 5 in March/April 2000 *QEX*

RF

By Zack Lau, W1VT

Feeding Open-Wire Line at VHF and UHF

It can be quite challenging to make a high-power balun that works well at VHF and UHF by scaling ferrite-core designs used at HF. The upper fre quency response limit can be increased by making the wires shorter, but smaller cores must be used to get a sufficient number of turns for effective coupling between the turns. Obviously, smaller cores won't handle much power. Thus, while ferrite cores may work for lossy TV baluns, I wouldn't recommend them for transmit applications at VHF and UHF. A loss of 0.5 dB may not be significant in casual receiving applications like cable TV, but it is quite substantial at the 1000-W level—it is a loss of 10.9% or 109 W![1]

Fortunately, it is seldom necessary to cover a wide frequency range—most VHF antennas cover only one amateur band. Thus, one can use a narrow-bandwidth design, and avoid lossy ferrite materials.

A balun that works well at VHF is shown in Fig 1, using a λ/2 of coax. It is often used to feed Yagi antennas. It is important to use an electrical λ/2 and shorten the physical coax length by its velocity factor. (Sometimes, people use a section of copper-jacketed semi-rigid coax such as UT-141A to reduce the difficulty of weather-proofing.) This has a 4:1 ratio, typically transforming a balanced 200-Ω load to an unbalanced 50-Ω load. However, it could also be used to transform a 300-Ω twin lead to 75-Ω coax. The velocity factor is 0.70, so a 71.4-cm length has an electrical length of 102 cm. Solid-dielectric cables are often preferred; as foam or partial air dielectrics may show more variation in measured velocity factors.

A λ/2 of coax does two things—it reflects the input impedance to the output and introduces a 180° phase shift. Thus, if the coax is terminated in 100 Ω, it reflects the 100-Ω load as

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[1]Notes appear on page 61.

100 Ω, no matter what the impedance of the coax. The two 100 Ω loads in parallel provide the desired 50-Ω input impedance. For a single frequency, the impedance of the coax doesn't matter very much. However, the impedance does affect the bandwidth of the balun. For resistive terminations, raising the coax impedance improved the bandwidth of the input match, as shown in Fig 2. Thus, 75- or 100-Ω coax might be used to improve balun bandwidth. RG-62 and RG-133A are two types of 93-Ω coax that are commercially available, the latter being more suitable for high-power work.

Since RG-133A may be difficult to find, I've made 100-Ω coax in the past out of 0.5-inch brass tubing and #12 copper wire.

The equation for the characteristic impedance (Z_0) of air-dielectric coax is

$$Z_0 = 138 \log\left(\frac{D}{d}\right) \quad \text{(Eq 1)}$$

Where D is the inside diameter of the outer conductor and d is the outside diameter of the center conductor. Thus, with a wall thickness of 14 mils,

$$Z_0 = 138 \log\left(\frac{0.5-(2\times 0.014)}{0.0808}\right) = 106\,\Omega \quad \text{(Eq 2)}$$

I used thin Teflon spacers to hold the center conductor in place. Thus, instead of searching for rarely manufactured RG-125/U or 150-Ω coax, you might make your own from tubing and wire.

Typical line impedances for commercial twin-lead or open wire are 300 and 450 Ω, not 200 Ω. Thus, a matching transformer is needed to go from 200 Ω to the desired line impedance. Charles Emil Ruckstuhl, W1ZJD, showed that a λ/4 of 300-Ω twin-lead could be used to match a 200-Ω balun to 450-Ω homebrew ladder-line.[2] He used a λ/2 balun of RG-213 to feed a 500-foot run of ladder line on 2 meters. This is shown schematically in Fig 3.

Transmission-line impedances other than 450 Ω can also be matched with a λ/4 line section. The impedance of the section is:

$$Z_0 = \sqrt{Z_{out} \times Z_{in}} \quad \text{(Eq 3)}$$

For 300 Ω, the impedance is:

$$\sqrt{300 \times 200} = 245\,\Omega \quad \text{(Eq 4)}$$

This impedance is non-standard, but this can be generated at home or modified from commercially available transmission lines. Andrew Griffith, W4ULD, describes a technique for modifying ladder line with a propane torch and a wooden jig.[3] Andrew lowers the impedance by reducing the spacing and reforming the polyethylene insulation.

Alternatively, the equation for open wire line is:

$$Z_0 = 276 \log\left(\frac{2s}{d}\right) \quad \text{(Eq 5)}$$

Where Z_0 is the line impedance
s is the spacing between the centers of the lines d is the diameter of the wires

Or,

$$s = \left(\frac{d}{2}\right) 10^{\left(\frac{Z_0}{276}\right)} \quad \text{(Eq 6)}$$

Thus, if the impedance is 245 Ω and #12 wire (with a diameter of 80.8 mils)

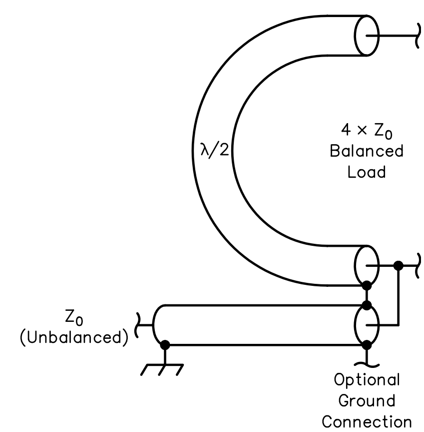

Fig 1—A 4:1 balun using λ/2 of coax.

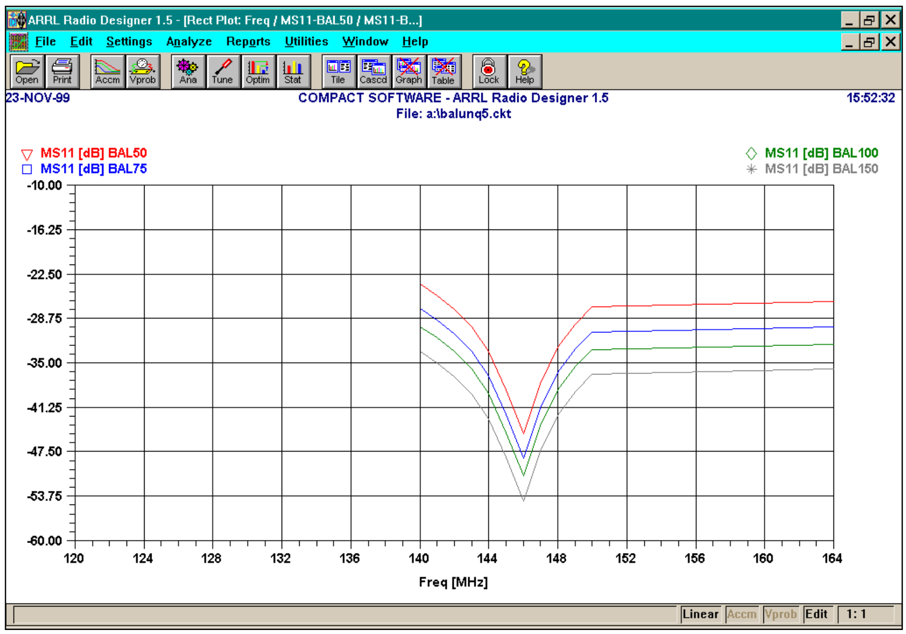

Figure 2—Raising the coax impedance to from 50 to 150 Ω improves the bandwidth of a 50:200 Ω λ/2 coaxial balun.

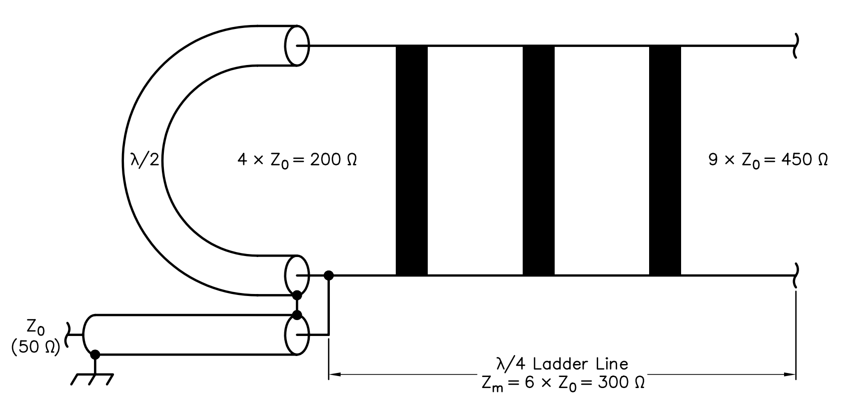

Fig 3—A 9:1 balun using λ/2 of coax and a 300-Ω λ/4 matching section.

is available; the required spacing is 312 mils.

Baluns can also be modeled on a computer—I've modeled the both the 50 to 200 and 50 to 450 Ω baluns using *Amateur Radio Designer* (ARD).[4] This software has a three-port model. The open-wire line is modeled as two separate ports each having one half of the line impedance. Thus, a 450-Ω open-wire is modeled as two 225-Ω coaxial lines (see Fig 4). When creating the ARD report form, don't forget to click on **Terms**... to set the proper terminations of 50, 225 and 225 Ω.

The performance of the balun can be determined by looking at the phase and power relationships between the ports. Ideally, there would be exactly 3 dB of loss from the input of the balun to each of the balanced outputs and a 180° phase difference between the transmission coefficients. Thus, if port 1 is the input and ports 2 and 3 are the outputs, you want to see 180° phase differences between PS21 and PS31. It is erroneous to expect a 180° phase shift for PS32, the phase of the transmission

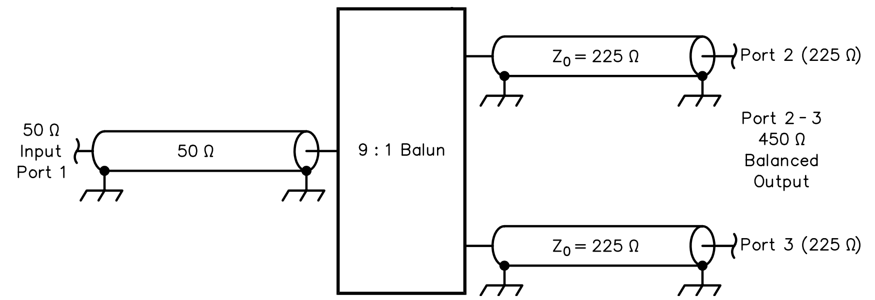

Fig 4—Modeling a 450-Ω balanced line as two 225-Ω coaxial lines with ARD's three-port model.

Table 1—Improved performance of a 9:1 balun using 75-Ω coaxial λ/2 balun (50TO450B) versus one using 50-Ω coax (50TO450L). Both use a 300-Ω λ/4 matching stub to go from 200 to 450 Ω.

Compact Software - ARRL Radio Designer 1.5 14-JUL-99 11:54:58
File: C:\ARD\BALUNQ5.CKT

Freq (MHz)	*MS11 (dB) 50TO 450B*	*MS21 (dB) 50TO 450B*	*MS31 (dB) 50TO 450B*	*PS21 (deg) 50TO 450B*	*PS31 (deg) 50TO 450B*	*MS11 (dB) 50TO 450L*	*MS21 (dB) 50TO 450L*	*MS31 (dB) 50TO 450L*	*PS21 (deg) 50TO 450L*	*PS31 (deg) 50TO 450L*
120.000	–27.60	–3.08	–3.10	–66.6	134.9	–18.19	–3.45	–2.89	–59.7	135.6
121.000	–28.06	–3.06	–3.11	–67.6	133.2	–18.68	–3.42	–2.91	–60.9	133.8
122.000	–28.54	–3.06	–3.11	–68.5	131.5	–19.19	–3.38	–2.93	–62.1	132.0
123.000	–29.03	–3.05	–3.12	–69.5	129.8	–19.72	–3.35	–2.94	–63.4	130.2
124.000	–29.52	–3.04	–3.13	–70.5	128.1	–20.26	–3.32	–2.96	–64.6	128.5
125.000	–30.03	–3.04	–3.13	–71.4	126.5	–20.81	–3.29	–2.97	–65.8	126.7
126.000	–30.56	–3.03	–3.13	–72.4	124.8	–21.39	–3.27	–2.99	–67.0	124.9
127.000	–31.10	–3.03	–3.13	–73.3	123.1	–21.99	–3.24	–3.00	–68.2	123.2
128.000	–31.65	–3.03	–3.14	–74.3	121.3	–22.61	–3.22	–3.01	–69.4	121.4
129.000	–32.23	–3.03	–3.14	–75.2	119.6	–23.25	–3.20	–3.03	–70.6	119.6
130.000	–32.83	–3.03	–3.14	–76.1	117.9	–23.92	–3.19	–3.04	–71.8	117.9
131.000	–33.46	–3.03	–3.14	–77.0	116.2	–24.63	–3.17	–3.05	–73.0	116.1
132.000	–34.11	–3.03	–3.13	–77.9	114.5	–25.37	–3.15	–3.06	–74.1	114.4
133.000	–34.80	–3.03	–3.13	–78.8	112.7	–26.14	–3.14	–3.06	–75.3	112.6
134.000	–35.53	–3.03	–3.13	–79.7	111.0	–26.97	–3.13	–3.07	–76.5	110.9
135.000	–36.30	–3.03	–3.13	–80.6	109.3	–27.86	–3.12	–3.08	–77.6	109.1
136.000	–37.13	–3.03	–3.13	–81.5	107.5	–28.81	–3.11	–3.09	–78.8	107.4
137.000	–38.03	–3.03	–3.13	–82.4	105.8	–29.84	–3.10	–3.09	–79.9	105.6
138.000	–39.02	–3.04	–3.12	–83.2	104.0	–30.98	–3.09	–3.10	–81.0	103.9
139.000	–40.10	–3.04	–3.12	–84.1	102.2	–32.26	–3.09	–3.10	–82.2	102.1
140.000	–41.33	–3.04	–3.12	–85.0	100.5	–33.70	–3.08	–3.10	–83.3	100.4
141.000	–42.74	–3.04	–3.12	–85.8	98.7	–35.40	–3.08	–3.11	–84.5	98.6
142.000	–44.41	–3.04	–3.12	–86.7	96.9	–37.44	–3.07	–3.11	–85.6	96.9
143.000	–46.45	–3.04	–3.12	–87.5	95.2	–40.00	–3.07	–3.11	–86.7	95.1
144.000	–49.08	–3.04	–3.12	–88.4	93.4	–43.40	–3.07	–3.11	–87.8	93.4
145.000	–52.79	–3.04	–3.12	–89.2	91.6	–47.76	–3.07	–3.12	–89.0	91.6
146.000	–58.68	–3.04	–3.12	–90.1	89.9	–48.96	–3.07	–3.12	–90.1	89.9
147.000	–61.69	–3.04	–3.12	–90.9	88.1	–44.78	–3.07	–3.12	–91.2	88.1
148.000	–54.84	–3.04	–3.12	–91.8	86.3	–41.03	–3.07	–3.12	–92.4	86.4
149.000	–50.34	–3.04	–3.12	–92.6	84.5	–38.24	–3.07	–3.11	–93.5	84.6
150.000	–47.29	–3.04	–3.12	–93.5	82.8	–36.07	–3.07	–3.11	–94.6	82.9
151.000	–44.98	–3.04	–3.12	–94.3	81.0	–34.28	–3.08	–3.11	–95.7	81.1
152.000	–43.13	–3.04	–3.13	–95.2	79.2	–32.77	–3.08	–3.11	–96.9	79.4
153.000	–41.57	–3.04	–3.13	–96.1	77.5	–31.46	–3.09	–3.10	–98.0	77.6
154.000	–40.23	–3.03	–3.13	–96.9	75.7	–30.28	–3.09	–3.10	–99.2	75.9
155.000	–39.04	–3.03	–3.13	–97.8	74.0	–29.23	–3.10	–3.10	–100.3	74.1
156.000	–37.97	–3.03	–3.14	–98.7	72.2	–28.26	–3.11	–3.09	–101.4	72.4
157.000	–37.00	–3.03	–3.14	–99.6	70.5	–27.36	–3.12	–3.08	–102.6	70.6
158.000	–36.11	–3.03	–3.14	–100.4	68.7	–26.52	–3.13	–3.08	–103.8	68.8
159.000	–35.27	–3.03	–3.14	–101.3	67.0	–25.73	–3.14	–3.07	–104.9	67.1
160.000	–34.49	–3.03	–3.15	–102.2	65.3	–24.98	–3.16	–3.06	–106.1	65.3
161.000	–33.76	–3.02	–3.15	–103.1	63.5	–24.27	–3.17	–3.05	–107.2	63.6
162.000	–33.06	–3.02	–3.15	–104.1	61.8	–23.60	–3.19	–3.04	–108.4	61.8
163.000	–32.40	–3.03	–3.15	–105.0	60.1	–22.95	–3.20	–3.03	–109.6	60.1
164.000	–31.77	–3.03	–3.15	–105.9	58.4	–22.32	–3.22	–3.02	–110.8	58.3

```
*Model 2-meter 50-ohm to 450-ohm balun
*Half wave balun and quarter wave matching section
*Model of 100 ft of UT-141A semi-rigid coax
BLK
CAB 1 2 Z=50 P=1200000MIL V=0.70 C1=.348913 C2=.024644
141LINE:2POR 1 2
END

*Model of 100 ft of 300 ohm transmitting
*tubular parallel line.
*Data from page 19.2 1999 ARRL Handbook
BLK
CAB 1 2 Z=300 P=1200IN V=0.80 C1=.105641 C2=.022346
LINE:2POR 1 2
END

*Verify that the attenuation constants
*are the same when modeled as two 150 ohm
*transmission lines
blk
cab 1 3 z=150 p=1200in v=0.80 c1=0.105641 c2=0.022346
cab 2 4 z=150 p=1200in v=0.80 c1=0.105641 c2=0.022346
trf 100 1 0 2 r1=50 r2=300
trf 200 3 0 4 r1=50 r2=300
300line:2por 100 200
end

*Model 2-meter balun—50 ohms unbalanced to 200 ohms balanced
*Note: ports 2 and 3 are 100 ohms with respect to ground
BLK
CAB 1 2 Z=50 P=.719M V=0.7 C1=.348913 C2=0.024644
BALUN:3POR 1 1 2
END

****************************************************
*Look at the effect of coax impedance on a 4:1 balun

*Balun with 50 ohm coax
BLK
CAB 1 2 Z=50 P=0.719M V=0.7 C1=.348913 C2=0.024644
TRF 1 1 2 R1=200 R2=50
BAL50:2POR 1 2
END

*Balun with 75 ohm coax
BLK
CAB 1 2 Z=75 P=0.719M V=0.7 C1=.348913 C2=0.024644
TRF 1 1 2 R1=200 R2=50
BAL75:2POR 1 2
END

*Balun with 100 ohm coax
BLK
CAB 1 2 Z=100 P=0.719M V=0.7 C1=.348913 C2=0.024644
TRF 1 1 2 R1=200 R2=50
BAL100:2POR 1 2
END

*Balun with 150 ohm coax
BLK
CAB 1 2 Z=150 P=0.719M V=0.7 C1=.348913 C2=0.024644
TRF 1 1 2 R1=200 R2=50
BAL150:2POR 1 2
END
****************************************************

*Balun with 200 to 450 ohm transmission line transformer
*Quarter wave of 300 ohm line as matching section
*Zero loss case
blk
cab 1 2 z=50 p=0.719m v=0.7 c1=0 c2=0
trl 1 3 z=150 p=0.488m v=0.95
trl 2 4 z=150 p=0.488m v=0.95
50to450:3por 1 3 4
end
*

*Balun with 200 to 450 ohm transmission line transformer
*Quarter wave of 300 ohm line as matching section
*lossy line case with UT-141A 50 ohm semi-rigid balun
blk
cab 1 2 z=50 p=0.719m v=0.7 c1=0.34913 c2=0.02464
CAB 1 3 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
CAB 2 4 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
50to450l:3por 1 3 4
end

*Balun with 200 to 450 ohm transmission line transformer
*Quarter wave of 300 ohm line as matching section
*lossy line case with 75 ohm coax for the balun
*loss set equal to that for UT-141A
blk
cab 1 2 z=75 p=0.719m v=0.7 c1=0.34913 c2=0.02464
CAB 1 3 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
CAB 2 4 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
50to450b:3por 1 3 4
end

*Balun with 200 to 450 ohm transmission line transformer
*Quarter wave of 300 ohm line as matching section
*lossy line case with 100 ohm coax for the balun
*loss set equal to that for UT-141A
blk
cab 1 2 z=100 p=0.719m v=0.7 c1=0.34913 c2=0.02464
CAB 1 3 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
CAB 2 4 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
50to450c:3por 1 3 4
end

*Balun with 200 to 450 ohm transmission line transformer
*Quarter wave of 300 ohm line as matching section
*lossy line case with 150 ohm coax for the balun
*loss set equal to that for UT-141A
blk
cab 1 2 z=150 p=0.719m v=0.7 c1=0.34913 c2=0.02464
CAB 1 3 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
CAB 2 4 Z=150 P=16.18IN V=0.80 C1=.105641 C2=.022346
50to450d:3por 1 3 4
end

*Lengthen line lengths to optimize for
*both 144 and 432 MHz.
*Balun with 200 to 450 ohm transmission line transformer
*Quarter wave of 300 ohm line as matching section
*lossy line case with 75 ohm coax for the balun
*loss set equal to that for UT-141A
blk
cab 1 2 z=75 p=0.729m v=0.7 c1=0.34913 c2=0.02464
CAB 1 3 Z=150 P=0.416m V=0.80 C1=.105641 C2=.022346
CAB 2 4 Z=150 P=0.416m V=0.80 C1=.105641 C2=.022346
50to450x:3por 1 3 4
end
freq
*step 120mhz 164mhz 1mhz
*1mhz 10mhz 100mhz 1000mhz
step 140mhz 150mhz 1mhz
step 420mhz 450mhz 1mhz
end

opt
*Optimizer used to determine attenuation constants for 300 ohm twin
  lead
line f=1mhz ms21=-0.09dB
line f=10mhz ms21=-0.3dB
line f=100mhz ms21=-1.1dB
line f=1000mhz ms21=-3.9dB
end
```

Figure 5—The *Amateur Radio Designer* file used to model baluns.

Table 2—Performance of a Dual-Band 432/144 MHz Balun

Compact Software—*ARRL Radio Designer* 1.5 15-JUL-99 09:37:30
File: balunq5.ckt

Freq (MHz)	*MS11 (dB) 50TO 450B*	*MS21 (dB) 50TO 450B*	*MS31 (dB) 50TO 450B*	*PS21 (deg) 50TO 450B*	*PS31 (deg) 50TO 450B*	*MS11 (dB) 50TO 450L*	*MS21 (dB) 50TO 450L*
140.000	–41.33	–3.04	–3.12	–85.0	100.5	–33.70	–3.08
141.000	–42.74	–3.04	–3.12	–85.8	98.7	–35.40	–3.08
142.000	–44.41	–3.04	–3.12	–86.7	96.9	–37.44	–3.07
143.000	–46.45	–3.04	–3.12	–87.5	95.2	–40.00	–3.07
144.000	–49.08	–3.04	–3.12	–88.4	93.4	–43.40	–3.07
145.000	–52.79	–3.04	–3.12	–89.2	91.6	–47.76	–3.07
146.000	–58.68	–3.04	–3.12	–90.1	89.9	–48.96	–3.07
147.000	–61.69	–3.04	–3.12	–90.9	88.1	–44.78	–3.07
148.000	–54.84	–3.04	–3.12	–91.8	86.3	–41.03	–3.07
149.000	–50.34	–3.04	–3.12	–92.6	84.5	–38.24	–3.07
150.000	–47.29	–3.04	–3.12	–93.5	82.8	–36.07	–3.07
420.000	–32.28	–3.06	–3.22	105.5	–58.9	–22.70	–3.27
421.000	–32.89	–3.06	–3.22	104.6	–60.6	–23.34	–3.25
422.000	–33.53	–3.06	–3.22	103.7	–62.3	–24.00	–3.23
423.000	–34.19	–3.06	–3.22	102.8	–64.1	–24.69	–3.21
424.000	–34.89	–3.06	–3.22	101.9	–65.8	–25.42	–3.20
425.000	–35.63	–3.06	–3.22	101.0	–67.5	–26.18	–3.19
426.000	–36.41	–3.06	–3.22	100.1	–69.3	–26.99	–3.17
427.000	–37.24	–3.06	–3.21	99.2	–71.0	–27.85	–3.16
428.000	–38.14	–3.06	–3.21	98.3	–72.8	–28.78	–3.15
429.000	–39.12	–3.06	–3.21	97.5	–74.5	–29.78	–3.14
430.000	–40.19	–3.06	–3.21	96.6	–76.3	–30.87	–3.14
431.000	–41.39	–3.07	–3.21	95.7	–78.0	–32.08	–3.13
432.000	–42.76	–3.07	–3.21	94.9	–79.8	–33.43	–3.12
433.000	–44.34	–3.07	–3.20	94.0	–81.6	–34.98	–3.12
434.000	–46.23	–3.07	–3.20	93.2	–83.3	–36.78	–3.12
435.000	–48.57	–3.07	–3.20	92.3	–85.1	–38.90	–3.11
436.000	–51.57	–3.07	–3.20	91.4	–86.9	–41.37	–3.11
437.000	–55.33	–3.07	–3.20	90.6	–88.7	–43.79	–3.11
438.000	–57.45	–3.07	–3.20	89.8	–90.4	–44.59	–3.11
439.000	–54.50	–3.07	–3.20	88.9	–92.2	–42.86	–3.11
440.000	–50.84	–3.07	–3.20	88.1	–94.0	–40.31	–3.11
441.000	–47.97	–3.07	–3.20	87.2	–95.7	–37.98	–3.11
442.000	–45.72	–3.07	–3.21	86.4	–97.5	–35.99	–3.12
443.000	–43.88	–3.07	–3.21	85.5	–99.3	–34.30	–3.12
444.000	–42.33	–3.07	–3.21	84.6	–101.0	–32.84	–3.13
445.000	–40.99	–3.06	–3.21	83.8	–102.8	–31.54	–3.13
446.000	–39.80	–3.06	–3.21	82.9	–104.6	–30.38	–3.14
447.000	–38.73	–3.06	–3.22	82.0	–106.3	–29.33	–3.15
448.000	–37.75	–3.06	–3.22	81.2	–108.1	–28.36	–3.15
449.000	–36.85	–3.06	–3.22	80.3	–109.8	–27.46	–3.16
450.000	–36.01	–3.06	–3.22	79.4	–111.6	–26.62	–3.18

coefficient going from port 2 to port 3.

Table 1 shows the results of modeling the balun with 50- and 75-Ω coax. Surprisingly, simulations showed that the 75-Ω coax worked better than the 50, 100 and 150-Ω varieties, assuming the coax loss are constant.

The balun will also operate on odd harmonics. Remember that the λ/2 line for the balun must supply a 180° phase shift. At odd harmonics, the coax adds $(2{\times}w) + 1$ 180° shifts, where w is a whole number. Alternatively, the coax adds $w{\times}360°$ plus an additional 180° phase shift. Of course, the $w{\times}360°$ phase shifts equal a 0° phase shift, so the net effect of the longer line is still a 180° phase shift. In practice, the longer line will narrow the bandwidth of the balun, but the balun may still be useful for narrow-band work. Similarly, the coax is still a multiple of λ/2, which is required to reflect the proper impedance between the ends of the coax. Table 2 shows the 144- and 432-MHz performance of a 450:50 Ω balun using λ/2 and λ/4 transmission lines cut for 144 MHz. I used 75-Ω coax, as it offers a better bandwidth than 50-Ω coax.

Notes

[1]Percentage loss= $100\left(1-10^{\left(\frac{-\text{loss}}{10}\right)}\right)$

[2]C. Ruckstuhl, W1JZD,"A Simple Dirt-Cheap No-Loss Transmission Line, Almost," *Proceedings of the 19th Eastern VHF/UHF Conference of the Eastern VHF/UHF Society* (Newington: ARRL, 1993), pp 65-75.

[3]A. Griffith, W4ULD, "The 1/3-Wavelength Multiband Dipole," *QST*, September 1993, pp 33-35.

[4]ARRL products are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the World Wide Web at **http://www.arrl.org/catalog**.

Next Issue in QEX

Mark Mandelkern, K5AM, brings us the fifth and final segment on his homebrew transceiver. As we wrap up the series, the focus is on the control system, PTOs, frequency counter and construction details. Mark is also working on a piece for us about the "front-end" designs—the parts that change from band to band.

We begin a look at another high-performance homebrew rig with the first part of John Stephensen, KD6OZH's series on his *ATR-2000*. It's a synthesized, 20-meter mono-bander designed with due regard for actual conditions found on 14 MHz at John's QTH. He sets forth his goals and neatly explains the rationale behind his architectural choices. He gives circuit examples that address the needs for low phase noise and high receiver dynamic range. Check it out.

Jim Kocsis, WA9PYH, has been experimenting with 1691-MHz weather-satellite imaging and contributes his down-converter design. This synthesized unit may also cover several Amateur Radio bands with slight modifications. Jim uses convenient 50-Ω blocks, where possible. The result is an efficient and compact design: There is only one alignment point!

Johan Van de Velde, ON4ANT, presents his homebrew, multi-band Yagi design. He examines tradeoffs in element spacing and boom length and weighs his computer-modeling results before beginning construction. If you want to cover several HF bands with very good performance, look at his interlaced designs.

Letters to the Editor

Center-Loaded Whip Antenna Loss (Jul/Aug '99)

I have studied the article "Center Loaded Whip Antenna Loss" in the July/August '99 *QEX* and question a few of the statements. In the paragraph below Table 1, you state "The 'lossy whip' is just one of many types of compact, but poor, antennas available to the unsuspecting ham." I presume you meant the combination of low-Q loading coil and whip.

In several places, you refer to piano wire. I have a number of different commercial mobile antennas and they all use stainless-steel whips, some of which are machined to provide a taper. Piano wire has a completely different material composition and a constant diameter.

I do not understand the premise of Table 1. I gather the idea is to show that a coil with a Q of 33 is some 3 dB worse than some others, but we don't know the configuration of the other coils, the length of the antenna or the location of the coil in the antenna. The last listing is apparently for a whip without a coil, but of what dimension and for what frequency?

You have apparently ignored the losses contributed by the vehicle ground. Some references were made to losses of 3 and 6 dB in the coil/resonator assemblies. I cannot believe this is entirely true because 45 W (in my case) of dissipation in the low-Q Hustler inductor would probably result in degradation of insulation and plastic covering. This has not been the case. The temperature rise is discernable, but not excessive by any means. I do not know of anyone who has destroyed a coil unless running high power on 160 meters to a coil wound on PVC.

I do not dispute the possible reduction in field strength with these low-Q configurations, but I don't think all of the loss is in the coil. Some of the losses can apparently be offset by increasing the length of the base section. Raising or lowering the input resistance by changing coil parameters does not necessarily result in field strength changes. Is there such a thing as an infinite-Q and a loss-less coil?

The article by Bruce Brown, W6TWW, (*ARRL Antenna Compendium, Vol. 1*) has been widely quoted in ARRL publications. Most of the subsequent analyses have been based, I believe, on his conclusions. I am not knowledgeable enough to disagree with Brown's conclusions, but I have some trouble with a couple of the statements that seem to be taken as fact. On p. 109, for example: "Thus the coil forces a much higher current into the top section than would flow in the equivalent part of a full 90°-high antenna." If this is so, then wouldn't a base-loaded antenna coil "force" more current in the top section?

From another view, Jerry Sevick, W2FMI, in a series of articles some years ago concluded that 75% loading was somewhat better than mid-point loading. His conclusions however, were based on measurement of input resistance over a known ground system, not field-strength measurements. It may be that W2FMI's conclusions are correct for antennas of 45° and larger, yet may not hold for antennas of 30° and less, which Brown addressed.

Incidentally, Sevick appended some "Loading Coil Investigations" to his *Building and Using Baluns and Ununs* book. Among other things, he compares Hustler coils for 80/40/20 meters to homebrew coils and coil stock. While the Hustler coils, as we know, suffer from pretty low Q, his conclusions are curious: "On 20 meters, even the Hustler whips are only poorer by about 5 dB from the ideal quarter-wave condition. When looking at the other numbers, the question comes up as to whether it's worth replacing the Hustler resonators with high-Q coils." For 40 meters, his table shows an efficiency of 11.3% for the Hustler and 21.5% for a larger-diameter B&W coil. On 20 meters, the Hustler yields 35.5% and the B&W coil 41%. That equates to near 3 dB on 40 and almost nil on 20. His efficiencies are based on antenna input resistance changes and not field strength, so they may not be completely valid.

The AA6GL *MOBILE* program included with *Compendium* 4 is an informative tool, but it is also based largely on the conclusions drawn by Brown. My own experience with Hustler coils shows that the efficiency can apparently be improved by using a longer base section. In a recent Florida QSO Party, I operated with Hustler coils on three bands (40/20/15) using the multi-coil adapter and an 8-ft base section. This puts the coil at about 75% of the antenna height. My contact total was about 25% better this year than last, when I used the small bug-catcher, a 6-ft base and a 5-ft top. Band conditions, by the way, were much poorer this year.

Without confirming A/B field-strength tests, that proves nothing except that the low-Q Hustler configurations are certainly competitive. Having three coils on one base is also a very nice feature. The arrangement is not good for highway speeds, however. I do not favor operating while in motion, from a safety standpoint.

Perhaps this fall I will set up a small test range and check out some of these theories, such as center versus 75% loading, high-Q versus medium- and low-Q coils and any other possibilities that occur to me.—*Arlan Bowen, N4OO, 349 Persimmon Rd, Sopchoppy, FL 32358;* **abowen@nettally.com**

Hi Arlan,

There is a simple technique to determine the relative field from any wire antenna; it will show how top loading, center-coil loading and other methods for changing current distribution affect antenna performance. The field from an antenna is directly proportional to the area under its current-distribution curve. You can plot the current as a function of position on the antenna. The area under the curve is then described in ampere-feet or ampere-meters. You will find that the area often remains nearly constant for a fixed-length antenna, regardless of the method used to change the current distribution. This means that top loading, for example, may not increase gain *per se*, although it will change the input impedance. Top loading with horizontal elements atop a very electrically short antenna does however significantly increase the area under the curve.

Keep in mind that the RF current reverses phase every half wave along a wire, and this will create a negative (opposite phase) ampere-feet area under part of the curve that must subtracted from the positive ampere-feet area. In addition, if some of the antenna wires are oriented such that their current direction opposes that of the main radiator, the ampere-feet area associated with these wires must also be subtracted.

My recent article on mobile-whip antennas assumed a good conductor beneath the whip, because I have an aluminum plate on the top of my van. Whip performance from installations using steel as the ground plane may be different, in which case roof losses may be more significant than coil losses. Some vehicles have a plastic roof, in which case I highly recommend that a large aluminum ground plane be installed and carefully bonded to the vehicle ground. This may also help to reduce EM field intensity within the vehicle, but that will vary from one instal-lation to the next.—*Grant Bingeman, KM5KG, 1908 Paris Ave, Plano, TX, 75025;* **DrBingo@compuserve.com**

On Antenna-Mounting Effects

Dear Doug,

Since about the middle 1970s, I have maintained a particular interest in the performance of antennas in their operational environments.[1] My studies have included effects of the platform structure that supports the antenna, such as automobiles, ships and sailboats with aluminum masts and conductive rigging; the effects of the environment, such as high-rise buildings and high-voltage power lines; the effect of Yagi antennas on masts used to support half-slopers, or the mast itself as a shunt-fed, grounded radiator on the lower bands; and the effect of the support structure and towers on the performance of antennas they support, such as dipoles and loops.[2, 3] I tell you this because I wish to comment on some aspects of articles and remarks published recently in *QEX*.

Peter Madle, KE6RBV, (*QEX*, May/June 1999, pp 32-42) numerically modeled HF and VHF antennas on cars, campers and RVs. I have carried out similar studies, comparing experiments with modeling.[4] I have a particular interest in HF coil-loaded antennas, previously on 4×4 SUVs (GMC Jimmy and Yukon trucks) and currently on a Ford F-250 pick-up truck. I use a version of D. K. Johnson, W6AAQ's screwdriver an-tenna, as fabricated and modified by Larry Parker, VE3EDY. I have his short-coil version for 80 meters and shorter wavelengths and a long-coil version for 160 meters. This is an excellent, versatile antenna. For camp application, I use a segmented 4.9-meter (16-ft) military-style whip (stainless-steel copper-coated) over the base section with a tuning inductor, instead of the 1.8-meter (6-ft) whip for mobile use. I prefer to mount the antenna on the left-rear bumper of the truck. This location provides gain in the direction away from the antenna toward the right-front corner of the vehicle. In a campground, one can always face the vehicle in the right direction. Particularly for the lower bands, it is best to have the full length of the truck working for you, providing enhance-ment in the desired direction.

My main purpose here is to comment on modeling VHF antennas. In this case, we are not interested in the space-wave patterns, which Madle has calculated for perfectly conducting and finite ground, except to ensure that we have an at-the-horizon lobe. For VHF, we are concerned with the ground-wave pattern, which generally is similar but can be different. In addition, to compare modeled performance with measured field strengths, we are definitely concerned with ground-wave field strengths (at HF as well as at VHF).

A final comment on VHF whips mounted in the center of the roof: One might expect that this would be the desired location, but it is not necessarily. Currents induced on the body of the vehicle can significantly control the radiation pattern, particularly if the vehicular structure happens to exhibit a resonance for the frequency in use. I have numerically modeled (using *NEC*-4D and a detailed wire-grid model) short whips mounted in the center of the vehicle roof over a wide range of frequencies (30-100 MHz, unpublished). I found that there are clear resonant effects, which can result in deep azimuthal nulls over narrow bands of frequencies. We much earlier found similar effects for HF marine whips on metallic ships (see Reference 1).

To minimize the effect of the vehicle on the pattern at VHF, it might be better to minimize the currents induced on the vehicle as a result of directly feeding the antenna against the vehicle. We can do this by using a vertical-J or Slim-Jim antenna. In this case, from a structural point of view, the best place to mount the antenna is on a short stub mast attached to the left (or right) rear bumper. Therefore, a rear-bumper-mount 2-meter antenna might not be a waste of time.

On a different topic, Rudy Severns, N6LF (*QEX*, September/October '99, p 59) has commented briefly on shunt-fed grounded verticals. He referenced a 1937 classic paper by Morrison and Smith. Personally, I like Laport,[5] an antenna book that overviews many different types of antennas as well as various versions of shunt-fed grounded towers. Broadcasters employ isolated towers, the impedance characteristics of which are well known. In general, amateurs use a mast as a vertical radiator for the 160-, 80- or even the 40-meter band that also supports one or more Yagi antennas. John True, W4OQ's article entitled "How to Design Shunt-Feed Systems for Grounded Radiators," published in a 1975 *ham radio* article and reprinted in *Communications Quarterly*,[6] has discussed the effect of this "top loading" on the resonant frequency of the mast and hence its impedance, which is important knowledge if we are going to shunt feed the tower. I well remember John, having corres-ponded with him in past years. He is now a silent key. He did not have the tools to numerically model the tower and Yagi antenna that we have now to determine the effect that the Yagi might have on the directional pattern of the shunt-fed mast. The effect can be significant.[7]

The Yagi's elements can carry appreciable current and so are clearly parts of the radiating system. If the Yagi is large, for example a 40-meter Yagi on a shunt-fed mast being operated on the 80-meter band or a 20-meter Yagi on a mast for 40 meters, one can rotate the direction of fire of the tower by rotating the Yagi!

Induced currents in support structures that arise because the antenna is fed against the structure, or currents induced to flow by parasitic excitation, whether intentional or unintentional,[8] are an important consideration when analyzing the performance of antennas. These effects range from insignificant to dominating and may control the direction of fire. Re-radiation by structures in the near field of the antenna is an important consideration that is generally ignored by most amateurs who write articles about antennas. For example, we can read in our literature that such-and-such an antenna is decibels better than an antenna previously used, but the evaluation was made with both the new and reference antennas in close proximity. Both antennas can contribute to the radiation field, no matter which antenna is driven.—*John S. (Jack) Belrose, VE2CV, 17 Rue de Tadoussac, Aylmer, QC J9J 1G1;* **john.belrose@crc.ca**

References

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[2]J. S. Belrose, "The Effect of Supporting Structures on Simple Wire Antennas," *QST*, Dec 1982, pp 32-35.

[3]J. S. Belrose, "Loops for 80-m DX," *QEX*, Aug 1997, pp 3-16.

[4]J. S. Belrose and L.Parker, "A Tunable All Bands Camp/Mobile Antenna (TABA): Radiation Characteristics determined by Experiment and Modelling," *Communications Quarterly*, Fall 1998, pp 47-57.

[5]E. A. Laport, *Radio Antenna Engineering*, McGraw-Hill, 1952, pp 104-111.

[6]J. True, "How to Design Shunt-Feed Systems for Grounded Vertical Radiators," *Communications Quarterly*, Fall 1991, pp 90-95.

[7]J. S. Belrose, "Comments on Gamma-Matching Grounded Towers," *Communications Quarterly*, Spring 1994, pp 74-79.

[8]J. S.Belrose, "Elevated Radial Wire Systems for Ground-Plane Type Antennas-+ Part 3: A monopole antenna with parasitic wire elements," *Communications Quarterly*, Spring 1999, pp 37-40.