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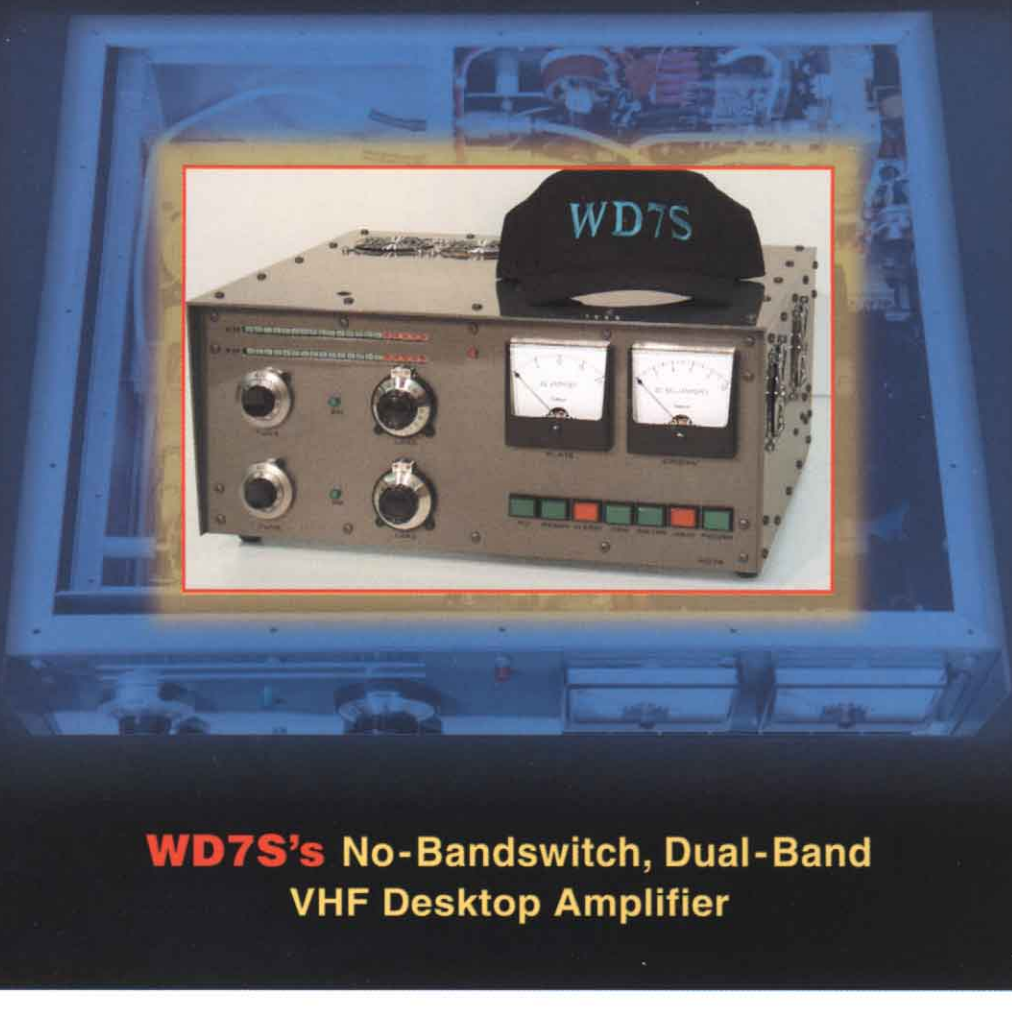
QEX

INCLUDING:
COMMUNICATIONS QUARTERLY

Forum for Communications Experimenters

July/August 2000



WD7S's No-Bandswitch, Dual-Band VHF Desktop Amplifier

ARRL The national association for AMATEUR RADIO
225 Main Street
Newington, CT USA 06111-1494

QEX

INCLUDING: COMMUNICATIONS QUARTERLY

QEX (ISSN: 0886-8093) is published bimonthly in January, March, May, July, September, and November by the American Radio Relay League, 225 Main Street, Newington CT 06111-1494. Yearly subscription rate to ARRL members is $22; nonmembers $34. Other rates are listed below. Periodicals postage paid at Hartford, CT and at additional mailing offices.

POSTMASTER: Form 3579 requested. Send address changes to: QEX, 225 Main St, Newington, CT 06111-1494

Issue No 201

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
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Subscription rate for 6 issues:

In the US: ARRL Member $22, nonmember $34;

US, Canada and Mexico by First Class Mail: ARRL member $35, nonmember $47;

Elsewhere by Surface Mail (4-8 week delivery): ARRL member $27, nonmember $39;

Elsewhere by Airmail: ARRL member $55, nonmember $67.

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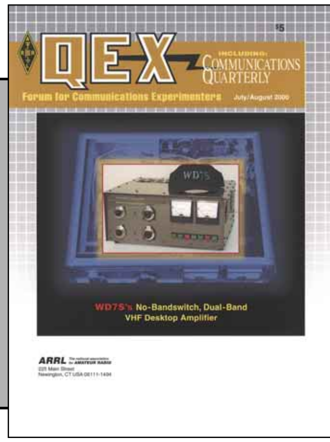

About the Cover

Check out WD7S's amplifier on page 3.

Features

Columns

July/Aug 2000 *QEX* Advertising Index

Empirically Speaking

We're pleased to announce that former *Communications Quarterly* staffers Peter Bertini, K1ZJH (**k1zjh@arrl.org**) and Douglas Page (**dpage@arrl.org**) will continue their columns here in our forum. Peter was Senior Technical Editor at *Communications Quarterly* and Associate Editor at *ham radio* before that. He also writes the "Radio Connection" column for *Popular Communications*. His "Tech Notes" column will carry on starting with the next issue. Doug brings us details of timely technology in his "Science in the News" and "New Products" items. Be sure to check his offering in this issue.

For more information, visit our newly updated Web page at **www.arrl.org/qex/**. We've added some historical information and new sample articles. Many of you have asked about back issues and reprints of *Communications Quarterly*, as well as the possibility of a CD ROM. Some issues are still available for $5 each from Publication Sales (**pubsales@arrl.org**). Given enough demand, a CD might appear: Let us know if you'd like that to happen.

It was delightful to see so many of you at Dayton this year. We think attendance broke the all-time record and some exciting new products and information were on tap. As the banquet speaker, Special Counsel for Amateur Radio Enforcement Riley Hollingsworth, K4ZDH, was a big hit. The FCC's Bill Cross, W3TN, said that it gets quiet on the air when Riley checks into a net. Reported Bill, "Even the static disappears."

Jim Haynie, W5JBP, attended the Hamvention in his new role as the League's president. Publisher Dave Sumner, K1ZZ, commented that Haynie's "The Big Project" initiative requires Amateur Radio to actively find its own way into school curriculums. "We can't look to educators to solve our problems for us," he said. Dave also reiterated his belief that Amateur Radio's technical experimenters will continue to play a very significant role in communications science. Others asserted that genuine and lasting interest in technical pursuits must be engendered at quite a young age and encouraged a "from-the-ground-up" approach. Investment of time and money in young folks now will likely pay important dividends later as we're asking, "Who will follow us?"

In This Issue

Paul Hewitt, WD7S, is serious about his meteor-scatter and weak-signal operations. He's taken an elegant idea to the limit of US Amateur Radio power output with his no-bandswitch amplifier for six and two meters. This is the frequency range where parasitic impedances become critical and as little as 0.1-inch of asymmetry in a push-pull circuit may mean the difference between malfunction and success. Precise construction and tuning details are given. Alpha/Power Inc of Longmont, Colorado indicate they plan to offer a version of this desktop unit soon.

L. B. Cebik, W4RNL, returns with the second part of his LPDA investigations. Results indicate some interesting effects as boom length and number of elements are increased. L. B. also dispels a myth or two about the way these broadband giants work.

Sam Ulbing, N4UAU, espouses auxiliary-station operation—even if it's just to cross a room. He begins a series on UHF remote-control transceivers that use off-the-shelf subsystems, showing that modular radio design for the experimenter need not be terribly complex. Recently, we've seen that designing with devices having high levels of circuit integration seems to require a slightly different set of skills than those used when trying for minimum production costs. Sam brings that out nicely.

Rick Peterson, WA6NUT, has breathed new life into his HW-101 with a DDS-driven, summing-loop PLL. In a summing-loop synthesizer, a mixer takes the place of one or more dividers. Computer control through a parallel port also brings CW keying and decoding to Rick's station. The system allows some extension of the Heathkit receiver's frequency range, too.

Wayne Stanley, W4RDG, presents a voltage standard for checking dc and ac voltmeters. He includes a bit about how to calibrate the calibrator.

In his column, Doug Page discusses an esoteric, high-speed wired communications mode that has some folks "buzzing." Zack Lau, W1VT, contributes a mathematical proof of conjugate matching and discusses its application to RF amplifiers.—*73, Doug Smith, KF6DX,* **kf6dx@arrl.org.**

A No-Bandswitch, Dual-Band VHF Desktop Amplifier

An old design idea together with new tubes yields legal-limit output on 6 and 2 meters from a small box.

By Paul Hewitt, WD7S

Dual-band VHF amplifiers are by no means new in Amateur Radio. Several designs have appeared over the years, both commercial and homebrew.[1, 2] The most recent commercial design was the Henry Tempo 6N2, now out of production for more than 20 years. There has never been a desktop capable of the new 1500-W PEP amateur power limit for both the 6- and 2-meter bands, until now. The amplifier I describe was designed primarily for meteor-scatter and weak-signal work, but the conservative design of the power supply and cooling system makes it well suited for EME (moonbounce) also. The amplifier uses a pair of Svetlana 3CX800A7s, operating in class AB2-push-pull on 2 meters and AB2-parallel on 6 meters. Features include: legal-limit output, no key-down time limit, no band switching, either-or operation, 50-60 W of drive for 1500 W PEP output on both bands, small size and light weight. In an effort to make this design easier to duplicate, off-the-shelf components and materials were used wherever possible. If you prefer not to "roll your own," Alpha Power should have a commercial version available soon.

[1]Notes appear on page 16.

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Bandon, OR 97411
wd7s@arrl.net

Output Tank Circuit and Construction Details

Since this was a new tank design, I decided to use a box-in-a-box approach for the layout to make adjusting and debugging much easier. This should also make duplication of the tank circuits much easier for those of you who wish to use a separate, outboard power supply. The tank compartment (See Figs 1, 2 and 7) starts with a 7×7×2-inch Bud chassis with two surplus 11-pin ceramic sockets[3] mounted three inches apart (center-to-center) and four inches from the rear of the chassis. The sockets are mounted 1/4 inch below the chassis in 2 1/2-inch-square holes with each socket oriented at 45° to the hole (Fig 3). This is similar to an Eimac 2216 socket for the 8877. The Eimac #1906 chimneys were used in the normal manner. This is probably the upper frequency limit for this mounting scheme due to possible instability problems at UHF. One problem encountered while mounting the sockets this way is the small amount of tube exposed above the chimney to grasp for removal. This is easily solved by connecting several 1/4-inch-wide hose clamps together end to end, then clamping them to the anode cooler to serve as a temporary handle.

The front, top and sides of the tank compartment are all made from 1/16-inch aluminum with 1/2×1/2×1/8-inch aluminum angle brackets at all corners. The completed compartment measures 7×6 1/4×14 1/2 inches (WHD). The front and right-side compartment panels extend to the top of the main cabinet cover to keep inlet and outlet air separated. The compartment cover (not shown) has 176 0.2-inch holes above the tubes for cooling exhaust. When this box is attached to the floor of the main cabinet, it leaves a 4×7-inch hole at the back before the rear panel is attached. This opening serves as access for a dip meter when checking the 2-meter tank resonance and tuning range.

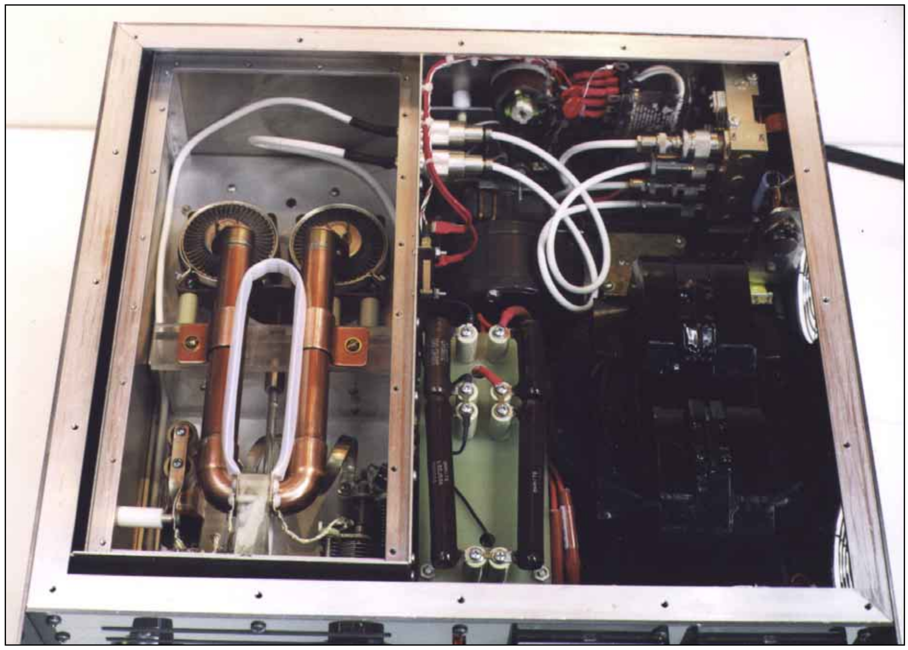


Fig 1—A top view of the amplifier. The plate transformer in the front right and high-voltage power assembly next to it occupy the majority of the control side with the blower, filament and control transformers sitting just behind. The tank compartment with the cover removed shows the plate line on its Plexiglas stand.

The 2-meter tank circuit consists of a shorted λ/4 balanced line section, tuned by a homebrew split-stator capacitor at the tube end of the line. Designing this type of tank circuit is made very painless with information and design examples provided in references.[4] There are several advantages of using a push-pull tank circuit instead of a parallel arrangement at 144 MHz. In the push-pull arrangement, C_{out} and C_{in} of the tubes are in series, allowing a total C_{out} and C_{in} that is one quarter of that with the parallel arrangement. This allows lower values of loaded Q, resulting in higher efficiency and reduced component heating, minimizing thermal drift. The low value of C_{out} in this case allows the use of a λ/4 line in place of a λ/2 line, increasing bandwidth and decreasing size.

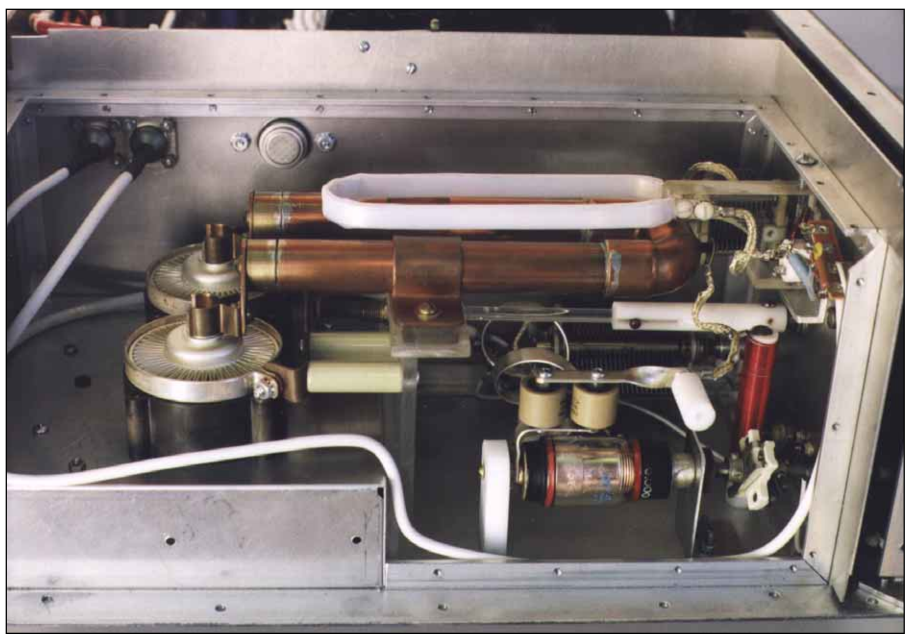


Fig 2—Side view of tank compartment. The 2-meter load capacitor on its ceramic standoffs mounted to the Plexiglas stand is in the center with the 6-meter tank coil just visible below the spline shaft assembly. The 6-meter tune capacitor on its L bracket is in the lower right front.

The line section, L2 of Fig 4, is made from 3/4-inch type-K copper pipe; the shorted end is made from one standard 90° copper elbow and one "street" 90° copper elbow. The open ends of the line are closed with brass plugs of the same outside diameter as the pipe, with a portion machined to fit inside the pipe. Standard copper pipe caps could be substituted for these plugs. The total length of the line section is 7 3/4 inches, from the ends to the inside of the shorted end. The plate-to-anode connections are made with 1/16-inch copper plate fastened between copper fuse clips on the anode connectors and threaded holes in the ends of the brass plugs.

The line section is supported by a Plexiglas stand and held in place by two copper clamps. The stators of the 2-meter TUNE capacitor, C3, are supported by 1 1/2-inch ceramic standoffs attached to the Plexiglas stand (Fig 5). A 3/8×5/8-inch ear from each stator attaches to a 3/8-inch-wide strap around each of the tubes' anode coolers. The rotor of C3 is a two-inch disk of 1/16-inch copper, which is mounted to a piece of 1/4×28 brass all-thread rod. The threaded rod is held by a brass fitting (tapped for 1/4×28) that is attached to the Plexiglas stand. The threaded rod is coupled to a sliding spline shaft consisting of 1/4-inch Plexiglas rod inside a 1/2-inch piece of Teflon rod that has been drilled through the center. The Teflon tube is attached to a short piece of 1/4-inch stainless-steel rod that exits the tank compartment through a 1/4-inch panel bushing. This shaft is coupled to the turns counter with a synchronous gear belt (Fig 6). This shaft relocation via the gear belt serves only to improve the front-panel appearance; it may be omitted if desired.

The 2-meter tank circuit is coupled to the antenna by an adjustable resonant link. The link position is adjusted by a #10-24 screw through a threaded Plexiglas block, to which the link is attached (Fig 2). One end of the adjusting screw is supported by the

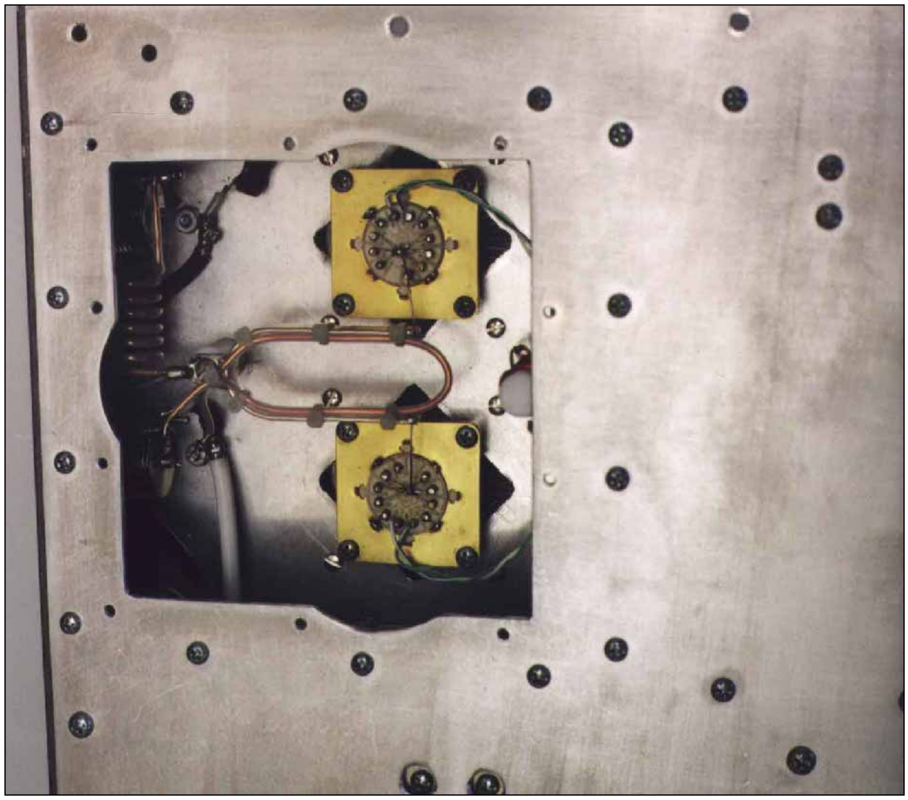


Fig 3—(right) Bottom view of amplifier with cathode compartment cover removed The filament choke is to the right with the 2-meter input network visible centered between the tube sockets. Both input tuning capacitors are mounted on the rear panel and one of the 6-meter input coils is just visible on the left. Note the use of coax for exciter lines to keep feedback paths minimized.

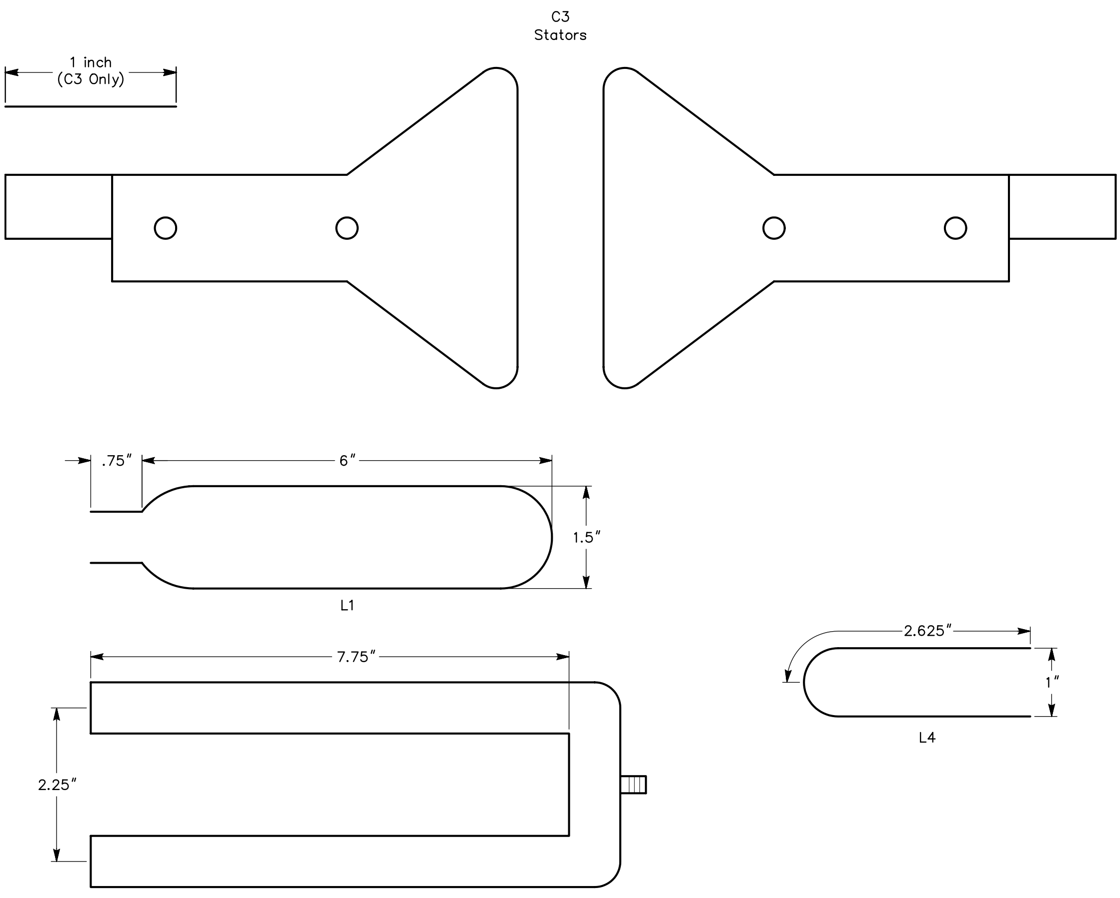


Fig 4—Details of 2-meter input and output tank inductors and the 2-meter TUNE capacitor stators.

top-cover angle bracket and the other end by a 2-inch-long piece of 1/2-inch angle mounted to the front panel of the compartment. The link is made from 3/8×1/16-inch silver-plated copper strap, covered with a 3/8-inch Teflon sleeve; it is attached to the Plexiglas block with four #10-32 nylon screws. Silver-plated coax braid attaches the link to the 2-meter **LOAD** capacitor, C2, and the isolation relay, K3. The link-adjustment screw is accessed through 3/8-inch hole in the main cabinet top cover during initial setup, then the hole is covered with a plastic plug.

As with all push-pull circuitry, the key to successful operation is symmetry. All stray capacitance must be divided equally between both sides of the line and tubes. Also, try to keep ferrous materials out of the tank compartment to prevent imbalance.

The 6-meter tank is a normal π network designed for a plate-load resistance of 1150 Ω with a loaded Q of approximately 22. The design parameters are shown in Table 1. The input of the π network is connected to the center of the shorted end of the balanced line section where a #10-32 brass stud has been soldered in place. This is the low-impedance point of the line section, and anything but a dead short can be attached here without affecting the performance of the line.[5] To verify that the physical center of the line is also the RF center, couple a dip meter to the line and tune for a dip around 144 MHz. Then without moving the dip meter (DM), touch the tip of a lead pencil at points along the line until you find the spot that has the least effect on the DM. This is the attachment point for the plate choke and the 6-meter tank.

To keep the loaded Q of the π network as low as possible, full 51-54 MHz operation was not attempted. The 6-meter **TUNE** capacitor, C1, operates very near its minimum capacitance because of strays and the C_{out} of both tubes. These strays account for a large portion of the 6-meter tune capacitance, which causes the majority of the tank circulating current to flow through the blocking caps, C7-C8. I used parallel Centralab 858 "door-knobs" for the blocking capacitors and I haven't experienced any problems. Capacitors with larger current ratings (such as HT-57s) would be a better choice for longer duty cycles.

The 3-30 pF vacuum-variable capacitor is mounted to the floor of the tank compartment with an **L**-bracket bent from 1/16-inch aluminum. Use the bearing retainer nut to attach the capacitor to the bracket, but be sure to add shims to the bearing to make up for space lost by the thickness of the bracket, or backlash will occur. The other end of the capacitor is supported by a #6-32 brass screw in the end of the capacitor that passes through a piece of Teflon bar stock mounted to the floor of the cabinet. A #6-32 hex nut holds the end of L3 and a piece of 3/8-inch strap (which supports C7 and C8) tight to the end of the vacuum cap. The other end of L3 is supported by a 1/2×1-inch Teflon standoff. Silver-plated braid ties this end to the 6-meter **LOAD** cap, which is mounted on the sidewall of the tank compartment.

Setting the Output Tanks

The 6-meter tank can be checked for tuning range with an SWR analyzer at the output connector. Connect a 1.1 kΩ, noninductive resistor from one of the tube anodes to ground to simulate the 1100-Ω plate-load resistance. You

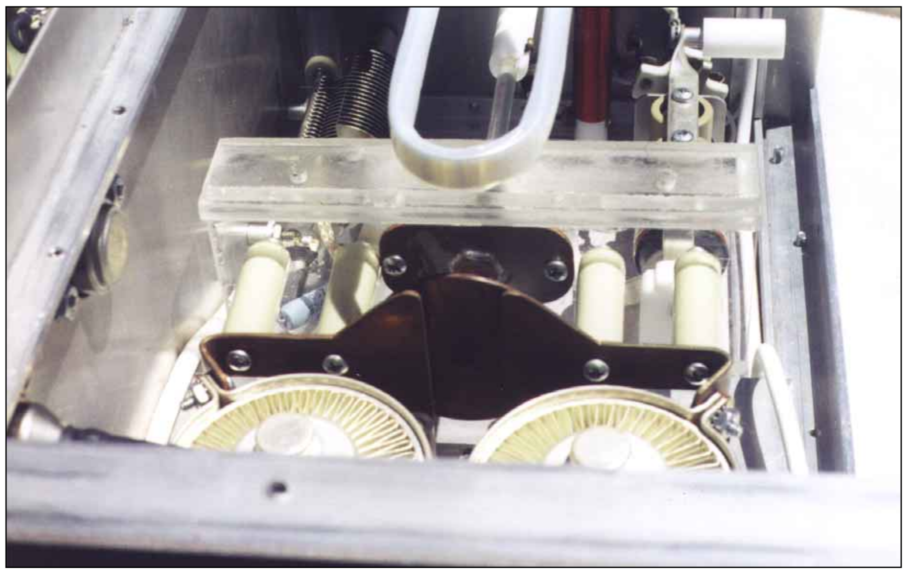


Fig 5—Rear view with the plate line removed showing the 2-meter TUNE capacitor.

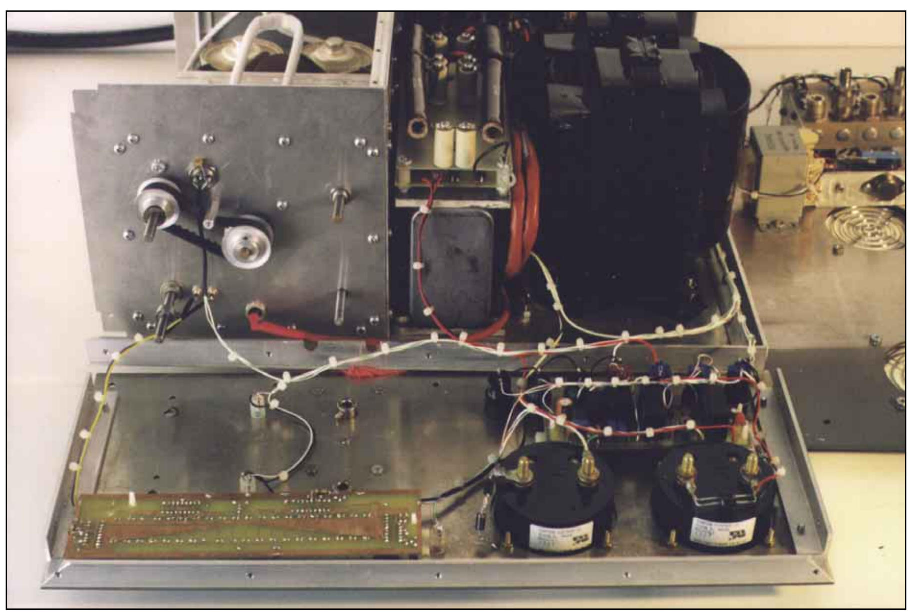


Fig 6—Front view of the amplifier with front panel folded down showing the location of high-voltage power-supply components at right and the gear-belt layout on the left. The bar-graph display PC board is visible in the lower left with the control board just beyond the two panel meters.

should be able to achieve a 1:1 match at both ends of the desired tuning range. Some stretching or compression of L3 may be required to bring the tuning range to 50-51 MHz. Be sure the top cover of the tank compartment is in place, as the stray capacitance it introduces accounts for a portion of the tuning capacitance for 6-meters.

The 2-meter tank was set up entirely with a dip meter inserted through the opening in the back of the compartment. Set the output link about 3/8-inch from the top plane of the line; this is close to the final position and adds stray capacitance. To keep loaded Q low, try to set the line length so it doesn't take very much tune capacitance to resonate the line at 144 MHz. Start with a line section a little longer than shown and carefully trim it to resonance. Remember that 1 inch of line length equals approximately 10 MHz of tuning range! The tank should resonate at 144 MHz with about 1/4 inch of space between the plates of C3. When the desired tuning range is found, install a 1/4×28 lock nut on the threaded shaft of C3 to stop travel at the upper frequency limit. This will prevent accidental tuning of the tank to the third harmonic of 50 MHz, which could potentially damage the front end of a 2-meter exciter. Remember that the isolation of most coax relays is very poor at 144 MHz: It was only 35 dB with the DowKey 260B used here. For this reason, a Jennings RJ1A vacuum relay, K3, was added in series with the 2-meter transfer relay. The vacuum relay disconnects the 2-meter link from the coax and terminates the coax in a resistive load during 6-meter operation. Make sure the electrical length of the coax between the relays is more than $\lambda/10$ and less than $3\lambda/8$ at 144 MHz. With these safety measures in place, third-harmonic energy at the 2-meter input port was measured at –17 dBm during 1500-W 6-meter operation.

Input Networks

The 2-meter input network consists of a single tuned-link, air-core transformer (Figs 3 and 4). This coupling method requires slightly more drive than a link-coupled, half-wavelength resonant line, but saves a lot of space. Two references[6, 7] give formulas and rules of thumb for designing this type of transformer, but it still required a lot of "cut and try" before suitable sizes for L4 and L5 were found. The loaded Q of the resonant input link is approximately 3.5 before the coupled-in resistance from the secondary modifies it. The degree of mutual coupling of air-core transformers is an elusive value that makes the final loaded Q difficult to calculate. The input tuning capacitor, C5, could be changed to a 25-pF unit since the one used here turned out to be much larger than needed. Both sides of C5 are above chassis ground; it requires a nonmetallic screwdriver to adjust.

Six-meter drive and B– are applied to the center tap of L4, providing parallel cathode drive for 50-MHz operation. A common T network, with a loaded Q of five, matches the 50-Ω line to the 24.6-Ω input impedance of the parallel tubes. All the cathode pins of the sockets are tied together with buss wire in a star pattern. One-inch-long bus-wire leads connect the ends of L4 to the center of the bus-wire stars on the sockets. Again, symmetry is all-important in balanced operation.

Both input networks were set up with an antenna analyzer and two 50-Ω carbon resistors to simulate the input load impedance. Tack-solder the resistors from cathode to ground on each socket with the shortest-possible leads. Short leads are very important on 2 meters because lead inductance and stray capacitance become quite significant. The resistors present 25 Ω to the parallel 6-meter network and 100 Ω to the 2-meter push-pull network. Connect the analyzer to the respective input ports and adjust each network for lowest SWR. Some adjustment of the

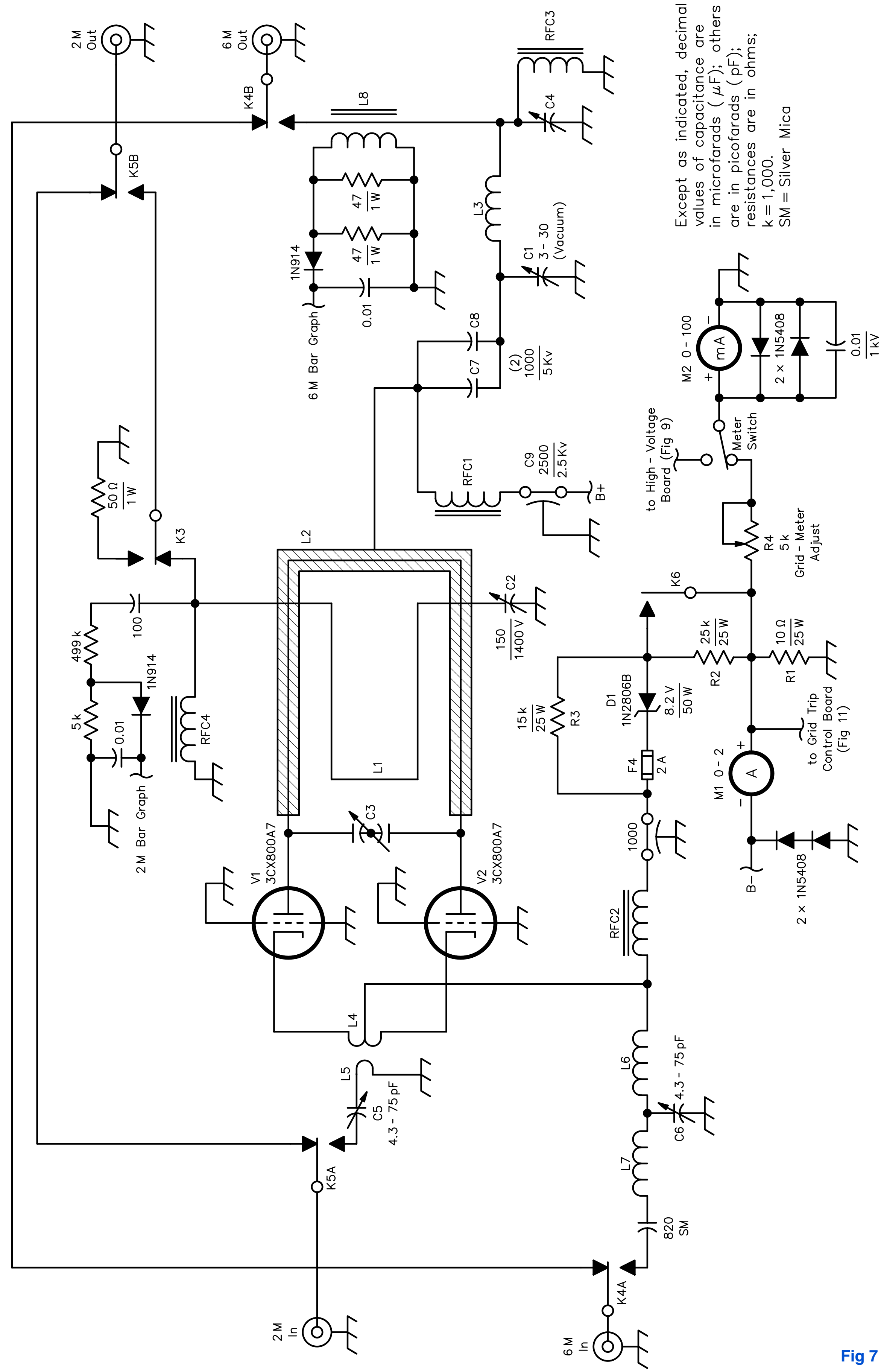
Table 1—Print out of operating parameters for the 6-meter input and output networks

The values for the PI matching network were calculated with Elmer (W5FD) Wingfield's new formulas found in the more recent ARRL handbooks.
FREEWARE Courtesy of KD9JQ
Triode Amplifier Program Version 2.0
For Grounded Grid Operation

(2) USER Biased 3CX800A7 at 50.0 MHz Rated for FORCED AIR

DC Plate Volts	=	2150.0	V (2500 V Max)
Max Plate Voltage	=	2150.0	V
Peak Plate Swing	=	1900.0	V
Min Plate Voltage	=	250.0	V
Plate Current Peak	=	3.368	A
Plate Current DC	=	1.119	A
Grid Current DC	=	0.063	A
Cath Current Peak	=	3.651	A
Design Plate RL	=	1128.1	Ω
RL for Matching	=	1154.8	Ω
Plate Dissipation	=	805.5	W (1600 W Max)
Grid Dissipation	=	3.2	W (8 W Max)
Cathode Bias	=	8.2	V
Conduction Angle	=	187.8	Degrees
Peak Grid Voltage	=	36.8	V
Zin @ Cathode	=	24.6	Ω
Pin Drive (PEP)	=	41.0	W
P0 @ Plate (PEP)	=	1600.0	W
P0 to Load (PEP)	=	1637.9	W
DC Power Input	=	2405.5	W
Efficiency	=	66.5	%
Power Gain	=	16.0	dB
Cath to Grid Cap	=	52.00	pF
Plate to Grid Cap	=	12.20	pF

T Match Input			***Pi Match Output***		
RS	=	50.0 Ω	RP	=	1154.8 Ω
L2	=	0.521 μH	CI	=	50.0 pF
CI	=	44.924 pF	L1	=	0.2377 μH
L1	=	0.422 μH	C2	=	232.0 pF
RFC	=	0.195 μH	RL	=	50.0 Ω
Zin	=	22.6 –*j* 9.9 Ω	QL	=	22.0
QL	=	5.0			



2 M Out
6 M Out
K4B
K5B
L8
RFC3
C4
47 1 W
47 1 W
1N914
L3
C1 3 - 30 (Vacuum)
0.01
6 M Bar Graph
C8
C7
(2) 1000 5 Kv
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
SM = Silver Mica
M2 0 - 100
mA
2 × 1N5408
0.01 1 kV
to High - Voltage Board (Fig 9)
Meter Switch
RFC1
C9 2500 2.5 Kv
B+
50 Ω 1 W
K3
L2
R4 5 k Grid - Meter Adjust
K6
C2
150 1400 V
100
499 k
1N914
5 k
0.01
RFC4
2 M Bar Graph
25 k 25 W
R2
10 Ω 25 W
R1
15 k 25 W
R3
D1 1N2806B
8.2 V 50 W
to Grid Trip Control Board (Fig 11)
M1 0 - 2
A
F4 2 A
L1
C3
1000
B–
2 × 1N5408
V1 3CX800A7
V2 3CX800A7
RFC2
L4
L5
C5 4.3 - 75 pF
L6
4.3 - 75 pF
C6
L7
820 SM
K5A
K4A
2 M In
6 M In

Fig 7

Fig 7—RF deck schematic. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. See Table 3 for part-supplier contact information.
C1—3-30 pF vacuum-variable capacitor
C2—150 pF air-variable, 1400 V capacitor
C3—Split-stator air-variable (see text and Fig 4)
C4—325 pF air-variable, 1400 V capacitor
C5, C6—4.3-75 pF APC-style air-variable trimmer capacitor
C7, C8—1000 pF, 5 kV doorknob capacitor
C9—2500 pF, 2.5 kV feedthrough capacitor
D1—8.2 V, 50 W stud-mount Zener diode, 1N2806B
K3—Jennings RJ1A SPDT vacuum relay, 26.5 V dc coil
K4, K5—DowKey model 260B with "C" option, coax relay 26.5 V coil
K6—SPDT relay, 10 A contacts, 24 V dc coil
L1—See Fig 4
L2—See Fig 4
L3—2 turns of 3/8×1/16-inch strap, 1 1/2-inch diameter, 2 inches long
L4—See Fig 4
L5—8.5 inches of #14 Teflon covered copper wire with 6 inches tied tightly inside L4
L6—7 turns, 1/2-inch ID×1 1/8-inch long Teflon covered #14 AWG copper wire
L7—9 turns 1/2-inch ID×1 1/2-inch long Teflon covered #14 AWG copper wire
L8—30 turns #26 AWG enameled wire on an Amidon T-37-17 powdered-iron core over center conductor of coax
M1—0-2 A, 3.5-inch Simpson panel meter
M2—0-100 mA, 3.5-inch Simpson panel meter
RFC1—36 turns #18 AWG enameled wire, tight-wound on 1/2-inch-diameter Teflon rod
RFC2, RFC3—Ohmite Z-50
RFC4—Ohmite Z-144
V1, V2—Svetlana 3CX800A7

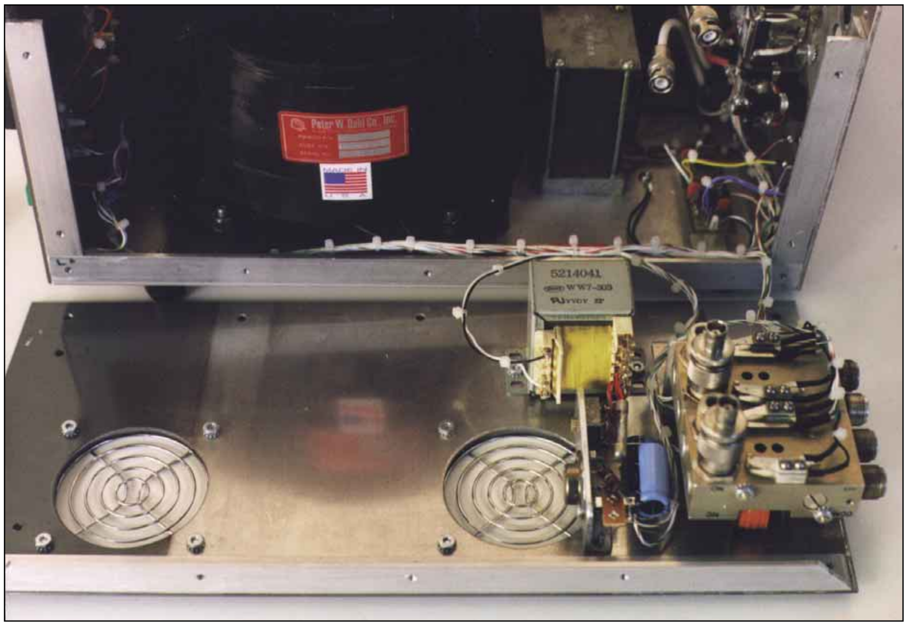


Fig 8—View of the right-hand side panel folded down showing the two transfer relays and the low-voltage power supply. The filament transformer is visible to the right of the plate transformer. Note the use of 80-mm fan covers for the cooling air inlets.

Table 2—Operating parameters of the amplifier

Plate voltage no load, 2350 V
Plate voltage @ 1 A, 2150 V
Zero signal plate current, 35 mA
6-Meter drive, single tone, 49 W
6-Meter power output,1420 W
6-Meter plate current, 980 mA
6-Meter grid current, 75 mA
Apparent efficiency, 67.4% (feed-through power not subtracted)
2-Meter drive, Single tone, 50 W
2-Meter power output, 1275 W
2-Meter plate current, 950 mA
2-Meter grid current, 68 mA
Apparent efficiency 62.4% (feed-through power not subtracted)

Table 3—Parts suppliers

Fair Radio Sales Co, Inc
1016 East Eureka St
PO Box 1105
Lima, OH 45804
tel 419-227-6573, 419-223-2196
fax 419-227-1313
e-mail **fairadio@wcoil.com**
URL **http://www.fairradio.com/**

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

Newark Electronics
4801 N. Ravenswood Ave
Chicago, IL 60640-4496
tel 800-463-9275, 773-784-5100
URL **http://www.newark.com/**

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Svetlana Electron Devices
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URL **http://www.svetlana.com/**

6-meter input coils (L6 and L7) may be required to achieve a 1:1 match. If the 2-meter network does not present a good match, some adjustment in length of L4 and L5 may be necessary. Again, the covers must be in place and tubes must be in their sockets when measuring the match. Both filament leads and the filament choke must be in their final positions because the cathode-to-filament capacitance affects input tuning at 144 MHz. Remember, this analyzer method only gets you close to the final tuning points; settings will be different under live conditions. Finally, don't forget to remove the temporary resistors from the cathode and tank circuits after the initial adjustments are done.

AC Mains and Low-Voltage Power Supplies

Several protective measures are built into the amplifier to protect the tubes and the operator. The ac mains are brought into the amplifier with four-conductor cord to keep the neutral and ground separated per the *NEC* (Fig 9). Three fuses are used to keep the blow-

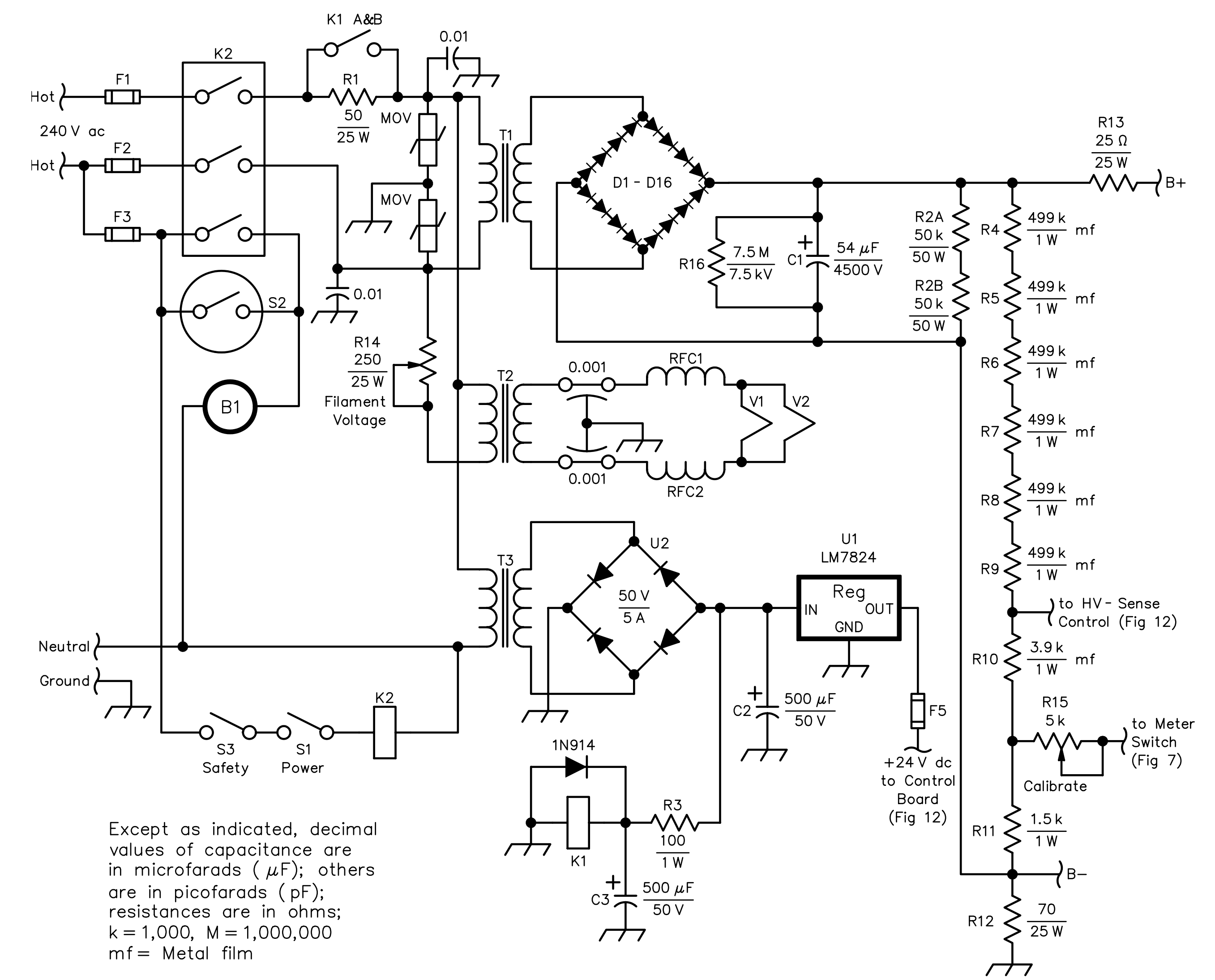


Fig 9—Power-supply schematic. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. See Table 3 for part-supplier contact information.

B1—Dayton #4C761 squirrel-cage blower
C1—54 µF, 4500 V oil-filled electrolytic capacitor
C2, C3—500 µF, 50 V electrolytic capacitor
D1-D16—1N5409 1200 PIV, 3 A Silicon rectifiers, four diodes per string
F1, F2—20 A, 250 V ceramic fast-blow fuse
F3—3 A, 250 V fuse
F5—2 A slow-blow fuse
K1—DPST relay, 25 A contacts, 24 V dc coil
K2—3PST relay, 25 A contacts, 120 V ac coil
MOV—130 V metal oxide varistor, V130LA5
R1—150Ω 25 W wire-wound resistor
R2A, R2B—50 kΩ 50 W wire-wound resistor
R3—100 Ω, 1 W carbon resistor
R4-R9—499 kΩ, 1 W metal film resistor
R10—3.9 kΩ, 1 W metal film resistor
R11—1.5 kΩ, 1 W metal film resistor
R12—70 Ω, 25 W wire-wound resistor
R13—25 Ω, 25 W wire-wound resistor
R14—250 Ω, 25 W rheostat
R15—5 KΩ, 10-turn potentiometer
RFC1, RFC2—12 bifilar turns of #18 AWG enameled wire on 1/2-inch-diameter Teflon rod
S1—SPST 5 A lighted panel switch
S2—Temperature snap-disc control Grainger # 2E245
S3—SPDT 5 A microswitch
T1—Peter Dahl plate transformer: 240 V primary, 1800 V, 1 A CCS secondary
T2—240 V primary, 16 V, 5 A secondary
T3—120 V primary, 20 V, 2 A secondary
U1—LM7824 TO-3 case
U2—5 A, 50 V rectifier bridge

er's ac source separate from the fused plate transformer. This was done so the blower's cool-down delay still functions after a high-voltage fault, which removes ac from the entire amplifier in dual-fuse designs. A thermal snap switch in the tank compartment keeps the blower running after shutdown or a high-voltage fault if the exhaust air has been above 110°F and is not yet below 90°. This only happens after several minutes of continuous operation. You may want to use a switch with slightly higher ratings (120°F on, 110°F off) if your shack is often warmer than the 90° off point used here. The 3CX800A7's filaments dissipate only 20 W each during cutoff and natural convection is more than enough to cool the tubes when shut off.

AC voltage for K2, the main control relay, is also taken from F3, the blower fuse, to make sure blower voltage is available before the amplifier can be powered up. A safety switch that closes when the top cover is in place supplies ac to S1, the main power switch. Power-supply inrush protection is provided by K1, which closes approximately one second after K2 and effectively removes R1 from the transformer primaries. The delay period is set by R3 and C3. There are two MOVs across the ac lines after the contacts of K2 and K1. Arcing at the relay contacts can and will produce voltage spikes and spike protection on the fuse side of K2 will not always save the high-voltage diode strings and other components. Filament voltage is adjusted at the primary of T2 by R14, a 250-Ω, 25-W rheostat. Filament voltage is measured at the filament choke via two leads brought to the back panel of the amplifier. Purists may want to measure voltage at the filament pins, but the low filament-current demand of these tubes makes the voltage drop between the sockets and the choke negligible. The rheostat also provides passive filament-inrush protection.

The low-voltage supply—consisting of T3, U1 and U2—supplies regulated 24 V at 1 A. The bridge rectifier and 24-V regulator are mounted to a piece of 3/4×1-inch aluminum angle that is placed directly in the cooling-air inlet path (Fig 8).

High-Voltage Power Supply

The high-voltage power supply consists of a Peter Dahl Hypersil plate transformer and a full-wave bridge rectifier with capacitor input filter. The supply is capable of producing 2150 V at 1 A CCS. The entire supply is assembled as one piece, then installed on the floor of the main cabinet next to the plate transformer (Fig 6). The oil-filled filter capacitor is sandwiched between two 3×6 1/2-inch pieces of Plexiglas, in turn held together with four #10-24×5-inch-long flat-head screws. The rectifier strings and high-voltage-meter multiplier resistors are mounted to a 3×6 1/2-inch PC board supported by 1/4-inch spacers on the same four screws. A 3×6 1/2-inch piece of fiberglass board holding the two bleeder resistors, B+ current limiting resistor and B– float resistor tops off the stack. All of the power resistors on the top board are mounted on one-inch ceramic standoffs. High voltage is routed to the tank compartment with test-lead wire and a high-voltage feed-through capacitor. The diode strings do not use equalizing resistors or capacitors as they are from the same batch and there is plenty of PIV headroom. Spike protection for the string is in the transformer primary, where it belongs. The high-voltage supply has three bleed-down paths for the filter capac-

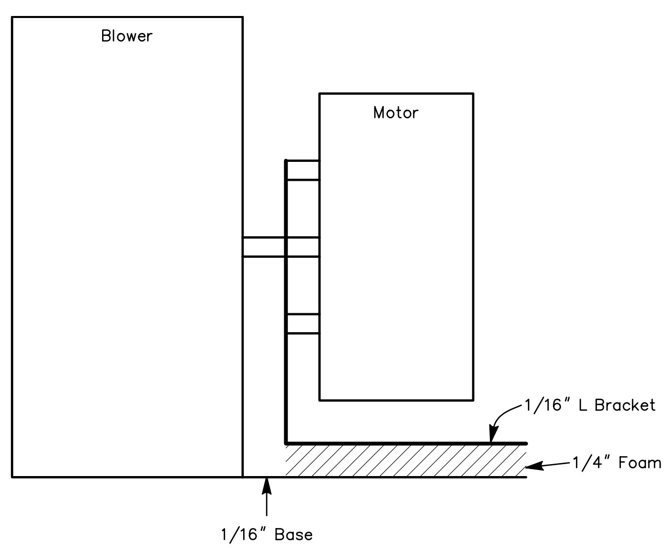


Fig 10—Layout of the blower modifications for sound reduction.

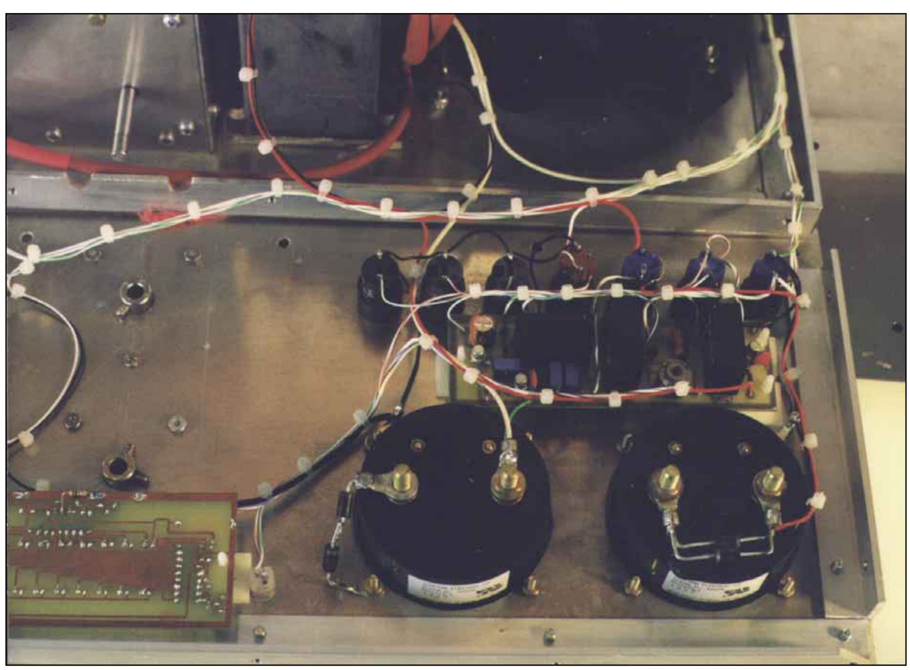
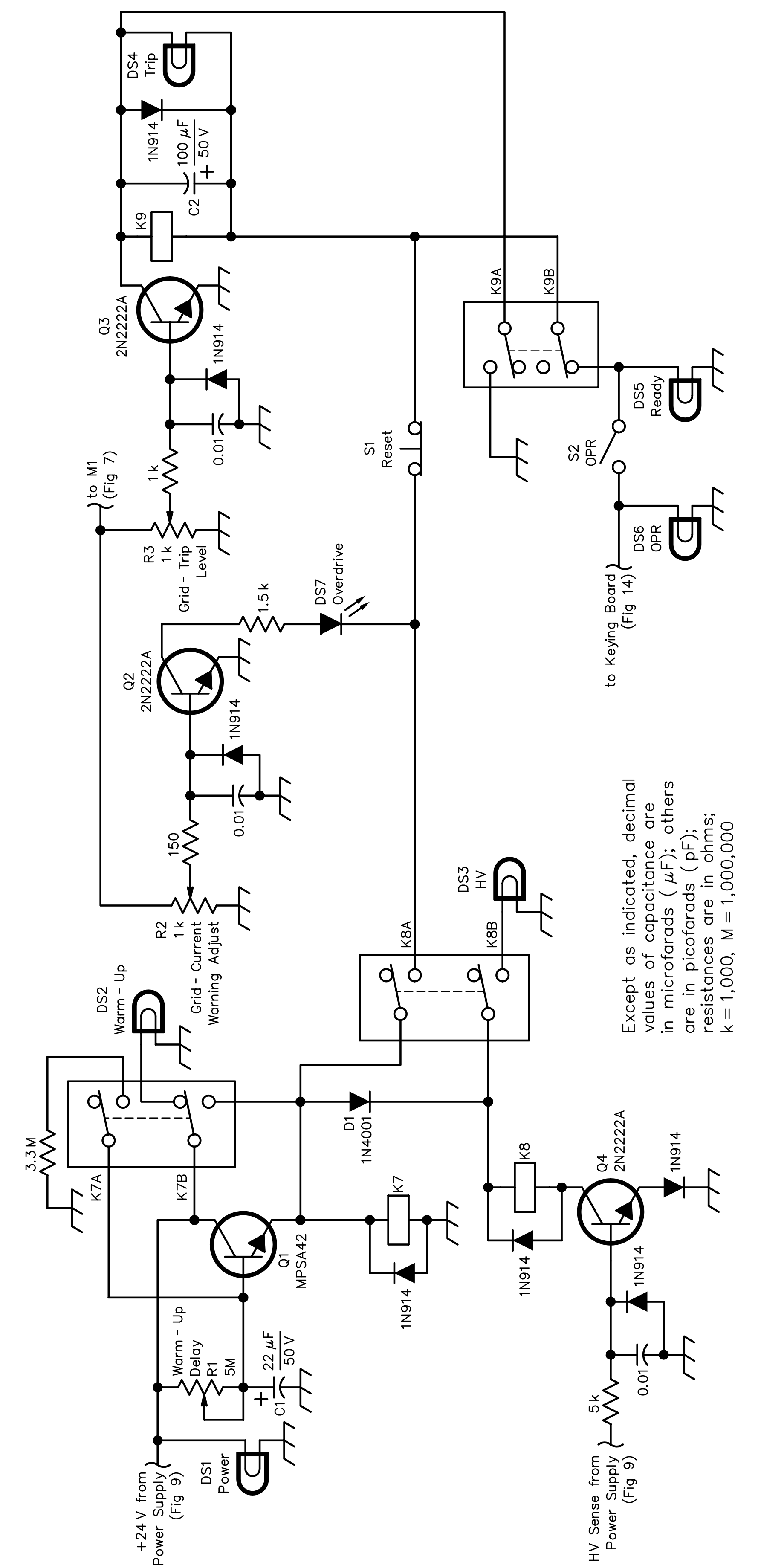


Fig 11—View of control board.



itor: the bleeder resistors, the high-voltage meter multipliers and a 7.5 MΩ, 7.5-kV resistor at the terminals of the capacitor itself. Even with these redundant safety measures, *never* assume that they are working. *Always follow standard safety procedures when working with any high voltage, including the ac mains: It can kill you.*

Cooling

Whole-cabinet cooling is accomplished via a Dayton model 4C761 squirrel-cage blower. Cooling air is drawn into the cabinet at the right-hand side through two $2^3/_4$-inch holes, one on each side of the plate transformer. This removes heat from the bleeder resistors and other components before it enters the blower inlet. The Svetlana 3CX800A7 datasheet recommends airflow of 11 cfm at a back pressure of 0.2 inches (of water), for 600 W dissipation at sea level and 25°C inlet air temperature per tube. For two tubes, this is 22 cfm at 0.2 inches for 1200 W of dissipation. At a 1500-W output level with 60% efficiency, the tubes only dissipate 960 W.

The socket sub-mounting method I used resulted in a back pressure of 0.2 inches with this blower, as measured on my bench with a home-brew manometer. Since the blower is rated for 43 cfm at 0.2 inches (sea level assumed), there should be plenty of headroom for different elevations and inlet air temperatures. Remember that the above calculations are for continuous dissipation, while SSB and CW operation rarely approaches 50% of these values.

The mounting flange at the bottom of the blower was cut off so the outlet can align with the 2-inch-tall cathode chassis. The blower motor was taken off its squirrel cage and mounted to an aluminum **L** bracket. The squirrel cage was mounted to a 4×4-inch piece of $^1/_{16}$-inch aluminum that supports the **L**

Fig 12—Control schematic. Unless otherwise specified, use $^1/_4$ W, 5%-tolerance carbon composition or film resistors. See Table 3 for part-supplier contact information.
C1—22 µF, 50 V electrolytic capacitor
C2—100 µF, 50 V electrolytic capacitor
DS1—24 V lamp part of S1 (see Fig 9)
DS2, DS3, DS5—24 V panel light
DS4—24 V lamp part of S1
DS6—24 V lamp part of S2
K7, K8, K9—DPDT 5 A DIP relay, 24 V dc coil
R1—5 MΩ PC-mount potentiometer
R2, R3—1 kΩ PC-mount potentiometer
S1—Normally closed momentary-contact lighted panel switch
S2—SPST lighted panel switch

bracket through a piece of 1/4-inch foam (Fig 10). This sound-isolation method was borrowed from Alpha Power Inc, and it works very well.

Control Board and Metering

The control PC board (Fig 11) is mounted on nylon standoffs just below the panel meters and above the row of switches and panel lights. As with most indirectly heated, oxide-coated cathodes, a warm-up period is required; for the 3CX800A7, a minimum of three minutes is recommended. When the amplifier is turned on, regulated 24 V is applied to the warm-up delay circuit consisting of Q1, R1 and C1 (Fig 12).

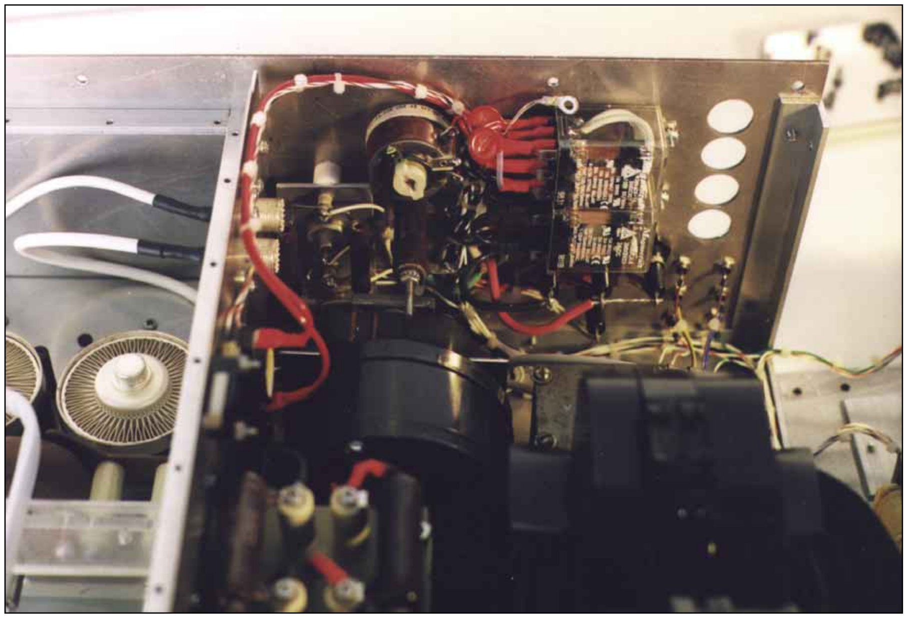


Fig 13—(right) View of rear panel showing K1 and K2. The bias heat sink and assembly is visible below the filament rheostat.

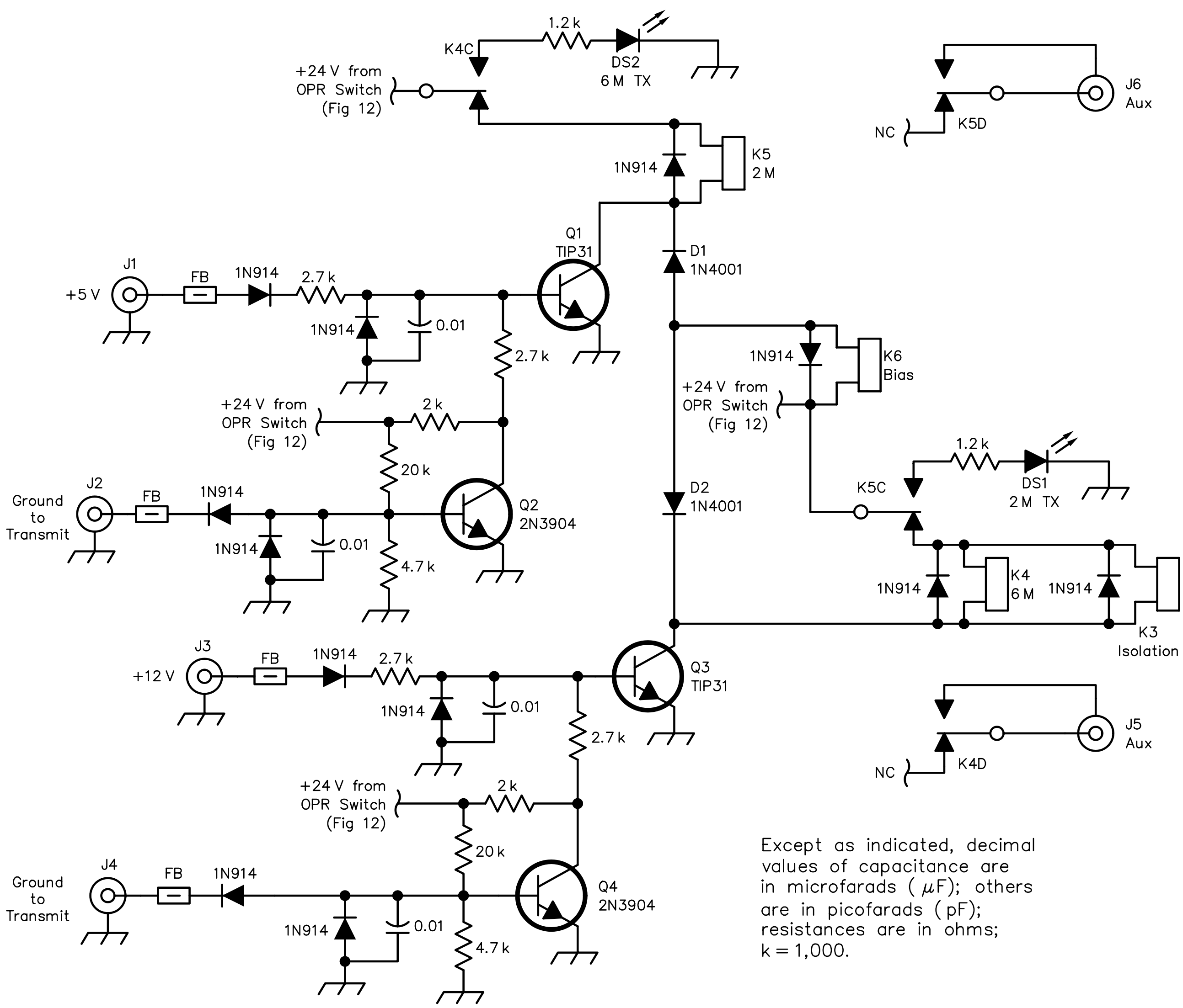


Fig 14—Schematic of the dual keying-circuit PC board. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. See Table 3 for part-supplier contact information.

When sufficient voltage appears at the base of Q1, relay K7 closes, with one set of contacts latching the relay on while the other set of contacts resets C1. If high-voltage is present, Q4 is switched on and K8 closes, lighting DS3 and supplying 24 V to the rest of the amplifier. Diode D1 was added between the field of K8 and the field of K7 to make sure K7 closes first. The proximity of the control board to the magnetic field of the plate transformer affects the timing of K7 slightly.

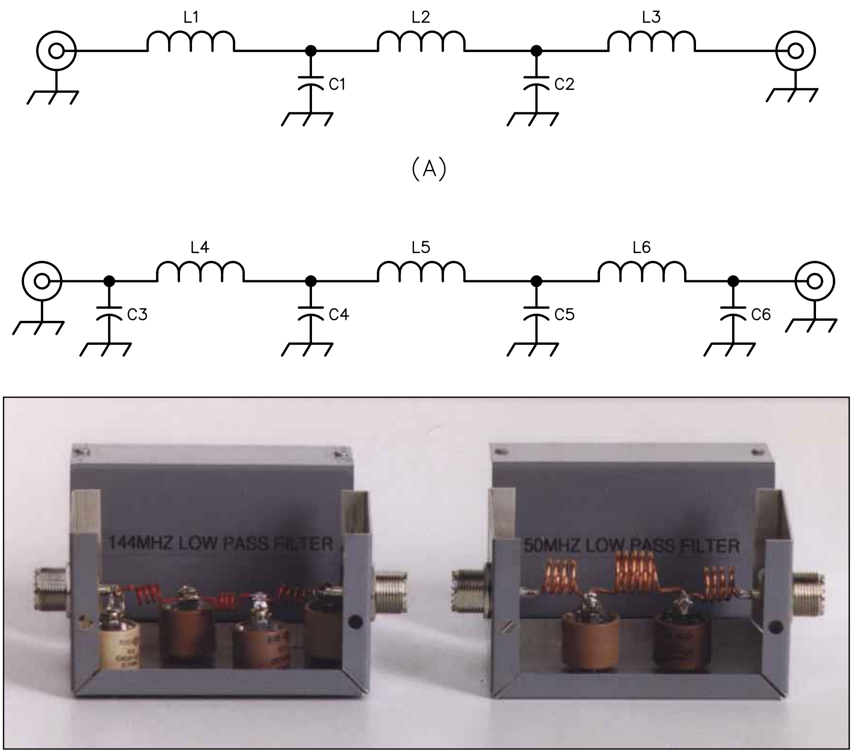
With the low grid-dissipation rating of these tubes, grid over-current protection is necessary. Grid current is measured as voltage drop across R1 of Fig 7, a 10-Ω, 25-W wire-wound resistor. R3 on the control board is set to fire K9 with Q3 at a grid current of 120 mA. One set of contacts latches the relay on; the other set interrupts the 24-V line to the **OPR** (operate) switch. Capacitor C2—across the field of K9—stops relay chatter on voice peaks. Lamp DS4 is part of the grid-reset switch S1, and it lights while K9 is energized. A second grid-current-sensing circuit, using Q2, is set to light DS7 on the front panel at 75 mA. This LED is used as a grid-current warning indicator and can be set at the grid-current level you prefer. The trip points of these two circuits and the grid-current meter reading are adjusted while the amplifier is disconnected from the ac supply.

To do so, connect a 24-V external supply to the 24-V bus of the amplifier after the timing circuit. Connect the positive lead of a variable-voltage dc supply to the ungrounded side of R1 of Fig 7 through an accurate dc milliammeter. Connect the external-supply negative lead to the chassis. With the grid-current panel meter disconnected, slowly increase the voltage of the bench supply until the milliammeter reaches the set points of 120 mA for grid trip and 75 mA for the warning LED. Adjust R3 and R2 (Fig 12), respectively, to set the trip points. After the trip points are set, reconnect the grid-current meter and adjust R4 (Fig 7) for a full-scale reading of 100 mA using the same method as for the trip points. This "cold" method of setting grid trip points is a lot safer for the tubes than removing B+ and then applying drive to induce grid current.

The panel meter, M2, serves as the high-voltage meter as well as the grid-current meter. The calibration of the high-voltage portion was "roughed in" with the amplifier off and all voltages removed. A 1.5-V dc supply was connected to the multiplier side of R15 on the high-voltage rectifier board, then R15 is adjusted for a full-scale reading on M2. This corresponds to a full-scale reading of 3000 V. Meter calibration was then checked with a high-voltage probe and an accurate DMM, after the amplifier was turned on.

Plate current is measured directly in the B– line of the high-voltage supply, which is held slightly above ground by R12 on the power-supply assembly. Both meters are protected by 1N5408 diodes; two series-connected diodes on the B– bus keep it below approximately 1.4 V and two more are connected back-to-back from the multimeter to ground.

There are relative-output bar-graph displays for each band above the tune and load controls. These displays use two LM3914 bar-graph drivers per 20-segment display. A current-transformer pickup unit is used for 6 meters and a voltage-sensing pickup for 2 meters. The dual-bar-graph-display schematic can be found in Forrest Mims' book[8] and the 6- and 2-meter pickups are shown in Fig 7. The displays look appealing but turned out to be more trouble than they are worth, since I still use external wattmeters while I tune. The bar-graph displays' rapid response time would be better suited for plate- and grid-current meters.



(B)

Fig 15—Details of the 2- and 6-meter low-pass filters. See Table 3 for part-supplier contact information.

50 MHz
C1, C2—50 pF doorknob capacitor
L1, L3—4 turns #14 AWG copper, 5/16-inch ID×1/2-inch long
L2—5 turns #14 AWG copper, 5/16-inch ID×5/8-inch long

144 MHz
C3, C6—25 pF doorknob capacitor
C4, C5—40 pF doorknob capacitor
L4-L6—3.5 turns #16 AWG silver-plated copper wire, 1/4-inch ID×5/16-inch long

Bias and Keying

Operating bias is developed across D1 of Fig 7, an 8.2-V, 50-W stud-mounted Zener diode. The Zener, along with K6, R2 and R3, is mounted on a 2×3-inch piece of 0.100-inch aluminum that serves as a heat sink (Fig 13). The entire assembly is mounted on two 1/2-inch Teflon stand-offs on the back panel of the cabinet. The 2-A cathode fuse is mounted on the back panel alongside the ac-line fuses.

The keying PC board is mounted on the rear floor of the main cabinet (Fig 8). The PC board contains keying circuits for both 6 and 2 meters (Fig 14). Because they are identical, I'll only describe the operation of one.

At rest, Q2 is biased on, holding the base of Q1 low. Grounding J2 will take the base of Q2 low, turning it off. This allows voltage to appear at the base Q1, turning it on and energizing K5. Alternatively, applying +12 V to J1 will also turn Q1 on, keying the amplifier. Both keying inputs are logic-compatible. The bias relay is energized through D1 or D2, which comprise a two-input **OR** gate. Each transfer relay is a DowKey #260B with the "C" option, which is a pair of DPDT signaling switches. Each of the relays' supply voltages is routed through one of the opposite relay's normally closed signaling contacts. This mechanical **EXCLUSIVE-OR** gate keeps the amplifier from being simultaneously keyed on both bands. The normally open contacts are used to light transmit LEDs on the front panel; these are mounted between their respective tune and load controls to help eliminate confusion while tuning.

With this type of transfer relay, input and output relay sequencing is obviously not an option, and hot switching of the relays will result unless preventive measures are taken. One option is to key the amplifier and let it key the exciter. Each relay's spare set of signaling contacts was brought to the back panel for this possible use. Another option, which I employ, is to use an outboard keying sequencer. The sequencer takes care of my mast-mounted preamplifiers as well as the exciter-amplifier timing. A suitable sequencer can be found in the references.[9] When using separate exciters for each band, the spare set of signaling contacts can be used for audio muting of the unkeyed exciter while transmitting.

(A)

(B)

Fig 16—The completed amplifier ready for its ride to the mountain top. (A) Front view, (B) back panel.

General Construction Notes

The entire amplifier cabinet was built using common hand and power tools. All aluminum was cut with carbide saw blades in a radial arm saw and table saw. Blades with a 5° negative hook angle seem to work best for aluminum. The main amplifier cabinet measures $18\times16\times7^1/_2$-inches (WDH). The plate transformer determined the cabinet height. The entire cabinet is built from 0.100-inch aluminum sheet and $^3/_4\times{}^3/_4\times{}^1/_8$-inch aluminum angle stock. All outside corners are secured with #8-32 pan-head screws tapped into the angle stock. The area under the plate transformer and high-voltage supply assembly was beefed up with a second layer of 0.100-inch aluminum on the cabinet floor. A square hole cut in the cabinet floor provides access the cathode compartment; it is normally covered with a piece of $^1/_{16}$-inch aluminum. All round holes were cut with chassis punches. The cabinet was painted with Dupont Chromaclear, which is a two-step (color coat/clear coat) automotive paint. After the base coat was applied over the primer, dry transfer lettering was applied to the front panel controls and Brothers P-touch labels were applied to the rear panel. Two coats of clear were applied over the base coat to finish the job.

All interior wiring was done with scrap Teflon-coated wire. The wiring harnesses were arranged so that all side panels can be removed and laid flat next to the chassis for ease of alignment and debugging. The internal coax runs were made with Teflon-dielectric coax to handle these high power levels, most other coax falls short, particularly at 144 MHz.

Adjustment and Operation

Use an ohmmeter to check the ac paths and RF deck for possible shorts and wiring mistakes. Be sure to blow out the entire cabinet with compressed air to remove any hidden debris. Set the filament-voltage rheostat to maximum resistance. If you haven't already done so, apply 24 V from an external supply to the 24-V bus and check the timing, control and keying circuitry for proper operation. Connect the amplifier to a suitable ac mains supply and connect an accurate DMM to the filament-voltage test points. Connect a 2-meter exciter and dummy load. Turn on the main power switch and quickly set the filament voltage to 13.5 V, or slightly less. *Never* operate the filaments below the 12.9-V recommended minimum. Be sure to check the filament voltage again after 5 to 10 minutes to detect any thermal drift in the rheostat. Verify that the plate is at approximately 2375 V.

Both tank circuits should be set to the tuning values found during initial setup. After the three-minute delay has elapsed, key the amplifier with no drive and check for a cathode idling current of approximately 30-50 mA. Apply a little 144-MHz drive and tune C5 in the input circuit for minimum reflected power. The 2-meter output link should be preset approximately 1/2 inch above the top plane of the balanced line. Still with little drive, adjust the 2-meter tuning capacitor for maximum power output. Adjust the 2-meter load capacitor for peak output power and then leave it there until you are finished. Next, adjust link spacing as you would a load capacitor. Keep increasing the drive while increasing the link coupling for maximum power output, without exceeding grid-current limits and while also touching up the main tune capacitor, C3. Once the desired output power level is reached, check the link coupling by slowly decreasing the coupling until a slight decrease in output power is observed. Then increase the coupling slightly past the point where output power peaks and grid current is reduced to within operating limits. This should be very close to critical coupling, which will result in maximum efficiency. Go back and touch up the input network for minimum reflected power. The link is now set for operation and any further tuning adjustments can be made with C2 and C3.

Tune the 6-meter tank circuit the same way you would any other π network: by slowly increasing drive and tuning for maximum power output and best efficiency while touching up the input network for lowest reflected power. Then increase loading until a 2% decrease in power output is observed.

With values of loaded Q below 30 in the output tanks, there is no way to keep harmonics below the 60-dB-down figure required by the FCC without outboard filtering. Harmonic filters are a very small price to pay for the higher efficiency, lower drive requirement and reduced thermal drift that the lower loaded Q provides. I've included descriptions of two suitable filters from the references;[10, 11] they are very easy to build. The silver-mica capacitors originally in the filters have been replaced with surplus doorknob capacitors for better power-handling capability. With both filters installed all harmonics and spurious signals are more than 70 dB down from the fundamental.

Final Thoughts

This amplifier was completed just before the summer E-skip season and was immediately put on the air. On-the-air signal reports were all good, and they compared well to the 8877 amplifier I normally use. Close in IMD testing was done with the help of NN7DX who is located nine miles away, over flat ground. Results showed no excessive signal bandwidth while our antennas were positioned to bring signals down to S-9 levels during 1500-W output testing. The amplifier runs cool to the touch even after several hours.

The biggest problem encountered with the operation of the amplifier had to do with interfacing with certain exciters. Some of the new multiband exciters provide multiple antenna ports that can be configured for HF, 6- and 2-meter outputs, but provide only one buffered keying line. The ICOM-746 is built this way, but it does provide two unbuffered keying lines: one for HF and 6 meters and one for VHF. The ICOM-706-series radios also provide two unbuffered lines but only two antenna ports. Both of these radios require some sort of external switching for the keying line—and the antenna line in the case of the '706—if this amplifier and an HF amplifier are both used. ICOM does provide band logic as a variable voltage at the radio's ACC plug. I've designed a simple decoder/buffer that uses this logic and the radio's internal power supply to automatically switch the keying line between HF, 6- and 2-meter amplifiers. The buffer will sink up to 3 A of relay current. It also selects from three separate ALC input lines for those of you that employ sequencers using the ALC line for transmit inhibit. The decoder is described in "Automatic Amplifier Selection for the ICOM IC-746, -736 and -706MKII Transceivers," *QST*, May 2000, pp 33-36. With the decoder in place, amplifier selection with an IC-746 becomes totally automatic when changing bands between HF, 6 and 2 meters. For the IC-706 series, an additional coaxial relay can switch one of the antenna ports between HF and 6 meters to allow automatic selection. The decoder also works with the ICOM IC-736, choosing between HF and 6 meters. Although not verified, I've been told that the Yaesu FT-847 does have separate buffered keying lines for each of its four antenna ports, making interface to this amplifier easy. Performance figures for the amplifier are listed in Table 2 and the completed amplifier is shown in Fig 16. Enjoy!

Acknowledgements

Thanks to Marv Gonsier, W6FR, and Roy Scanlon, NN7DX, for help in reviewing the text.

Notes

1 C. M. Maer Jr, W0IC, "The Perseids Powerhouse," *QST,* Oct 1959, p 32.

2 R. M. Richardson, W4UCH, "A Kilowatt Amplifier for 6 and 2 Meters," *QST,* June 1973, p 16.

3 The surplus sockets have a built-in 0.005 µf grid/screen bypass capacitor. RF Parts has a limited supply of these sockets. Substitutes include Johnson #124-0311-110 (with a grid collet) and Eimac SK1900.

4 G. R. Jessop, G6JP, *VHF/UHF Manual*, 4th edition, (Hertfordshire, England: Radio Society of Great Britain, 1983) Chapter 3.

5 E. P. Tilton, W1HDQ, *The Radio Amateur's VHF Manual*, (Newington: ARRL, 1972) p 77.

6 R. D. Straw, Ed., *The 1999 ARRL Handbook* (Newington: ARRL, 1998), pp 6-46 and 6-47.

7 R. Myers, W1FBY, *The 1975 ARRL Handbook* (Newington: ARRL, 1974), p 46.

8 F. M. Mims III, *The Forrest Mims Engineer's Notebook*, (San Diego, California: Hightext Publications Inc, 1992) p 107.

9 *The 1999 ARRL Handbook,* pp 22-53 through 22-56.

10 I. White, G3SEK, Editor, *The VHF/UHF DX Book,* Vol 1 First Edition (Buckingham, England: DIR Publishing, 1992) p 12-35.

11 R. Schetgen, KU7G, Ed. *The 1993 ARRL Handbook,* (Newington: ARRL, 1992) p 31-37, Fig 95.

Paul has been an electronics experimenter since age 8 and has always enjoyed it. He was licensed 12 years ago when time finally allowed it. He currently owns and operates a construction company that specializes in custom beachfront homes.

Notes on Standard Design LPDAs for 3-30 MHz Pt 2: 164-Foot Boom Designs

Let's analyze two longer designs in search of maximum HF gain, summarize what we've learned and consider avenues for further exploration.

By L. B. Cebik, W4RNL

Part 2 of these preliminary design notes adds the criterion of gain to those used in Part 1 (*QEX/Communications Quarterly*, May/June 2000): a usable SWR curve and good pattern control. The present exercise looks at a pair of designs based on increasing the array length again. One design expands the 26-element design to fill the 164-foot boom length, which results in a Tau of 0.9 and a Sigma of 0.05. The other design maintains the segment density of the 100-foot model by selecting a Tau of 0.94, with a resulting Sigma of 0.032. The number of elements in the larger array is 42. Both arrays were designed using "Tau-tapered" element diameters between 0.5 and 6.5 inches.

Both designs are capable of a free-space gain of about 7 dBi over much of the 3-30 MHz design passband, but not over all of it. The 26-element design falls short at the lower end of the spectrum, while the 42-element array shows reduced gain above 24 MHz. The front-to-back ratios of both designs are quite satisfactory for most operating circumstances, and obtaining usable SWR curves across the entire passband is no longer a problem.

The notes will also briefly explore the temptation to spot-modify LPDAs to enhance performance at one or more frequencies. The results often produce unpleasant surprises at other frequen-cies with no obvious harmonic relationship to the optimized frequency. Finally, we shall see the utility of making more detailed frequency sweeps across the intended passband of a given design.

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Preliminary Design and Modeling Considerations

Because of time constraints, the sampling of different models had necessary limits. The models all require at least 1000 segments to meet minimal segmentation density requirements across the range from 3 to 30 MHz. The larger model approaches 1700 segments in 42 wires. Even on a 400-MHz computer, 1-MHz-increment frequency sweeps across the passband required from 30 to 60 minutes.

Consequently, model construction

required considerable selectivity. A boom length of 164 feet resulted from considering various preliminary designs. The length is nearly the same percentage increment above the 100-foot designs as those were above the initial 60-foot design.

The models presented in this part of the notes represent two different design philosophies, despite the use of a constant 164-foot boom length. The 26-element model sought to increase Sigma to a better value by expanding the spacing between elements relative to the 100-foot design of Part 1. The design goal was to achieve a higher gain throughout the passband, whatever might be the results for source impedance, front-to-back ratio and pattern shape.

The 42-element design resulted from trying to sustain an element density similar to that used in the 100-foot 26-element design. The goal was to maintain the high front-to-back ratio and smooth SWR curve across the passband—with special emphasis on the lower frequencies—and to let the gain and pattern shape be whatever might emerge.

Although the 100-foot designs required terminating stubs for the inter-element transmission lines, neither of the two 164-foot designs seemed to benefit significantly from the terminations. Therefore, as a further move toward simplification, the stubs were omitted from the present design models.

164-foot, 26-Element LPDA

By increasing the boom length of the 26-element LPDA to 164 feet, the value of Sigma increases to about 0.05. This value is closer to optimal for highest gain from the array, since the spacing between elements is increased significantly (as the outline in Fig 1 shows). Selection of the precise length was made, in fact, by the second of our two models, which was designed by selecting both Tau and Sigma and allowing the length to be what it would. However, the 164-foot length proved useful, since it increased the array length over the 100-foot models by nearly two-thirds, the same ratio as between the 100 and 60-foot models. Therefore, to equalize lengths between the two designs used here, the 164-foot length was retained for both.

Table 1 provides a listing of element half-lengths and cumulative element spacing for the 164-foot, 26-element model. To maintain a relatively constant length-to-diameter ratio, "Tau-

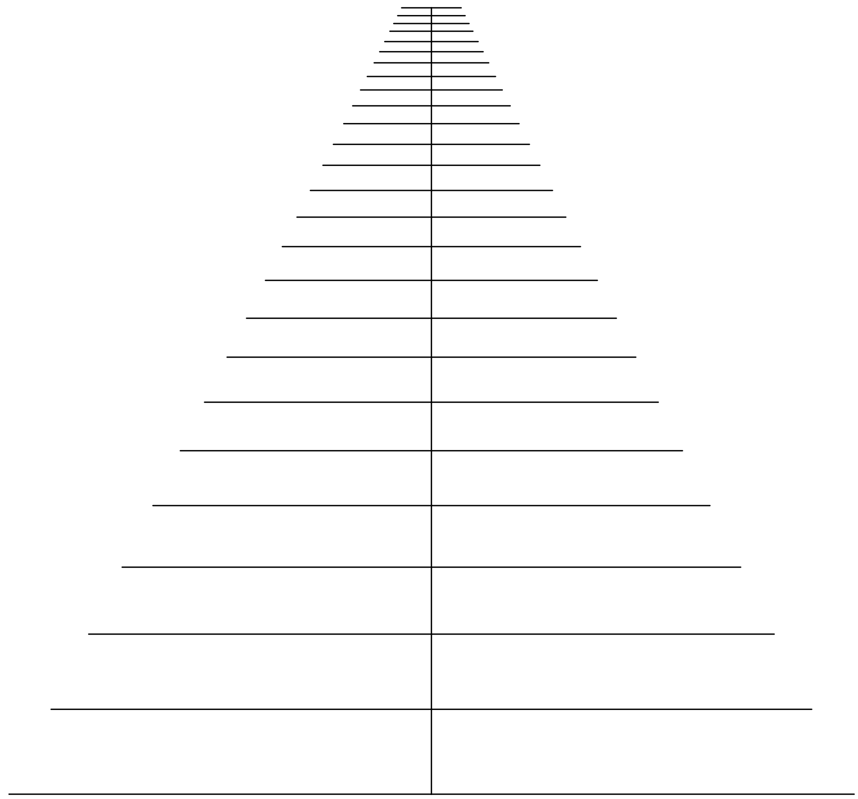

Fig 1—Outline of the 164-foot, 26-element 3-30 MHz LPDA; Tau = 0.90, Sigma = 0.05.

Table 1—Element half-lengths and cumulative spacing of the 164-foot, 26-element 3-30 MHz LPDA model.

Element No.	*Half Length (inches)*	*Cumulative Spacing (inches)*
1	1003.68	0.00
2	905.81	208.31
3	817.49	396.32
4	737.77	565.99
5	655.83	719.11
6	600.91	857.30
7	542.31	982.02
8	489.43	1094.58
9	441.71	1196.16
10	398.64	1287.84
11	359.76	1370.57
12	324.68	1445.24
13	293.02	1512.63
14	264.45	1573.45
15	238.66	1628.34
16	215.39	1677.87
17	194.39	1722.58
18	175.43	1762.92
19	158.33	1799.33
20	142.89	1832.19
21	128.96	1861.85
22	116.38	1888.62
23	105.03	1912.77
24	94.79	1934.57
25	85.55	1954.24
26	77.21	1972.00

tapering" was used for element diameters as well as lengths. Although noted in Part 1, the list of diameters is repeated in Table 2 for reference. As in past diameter tables, the progression is from the smallest to the largest element.

The omission of the terminating stub left open a question: What characteristic impedance would be best for the interelement transmission line? Therefore, the model was examined at 3-MHz intervals for its primary characteristics (free-space gain, front-to-back ratio and source impedance) to see if there was an advantage to one value over another. Characteristic impedances of 100, 150, 200 and 250 Ω were used at each check frequency. The results are recorded in Table 3. "Gain" is free-space gain in dBi; "Front-to-Back" is the 180° front-to-back ratio in decibels; and "Impedance" is the feedpoint or source impedance recorded as resistance ± reactance in ohms. The highest gain and front-to-back values for each frequency are marked with an asterisk.

Certain trends are immediately apparent. First, the highest gain figures occur at the lowest interelement transmission-line characteristic impedance. Second, as the frequency increases, the gain values tend to fall off more rapidly with increasing values of transmission-line impedance. Consequently, it would appear that the use of 100-Ω line would be automatic. At that value, all frequencies except 3 MHz would show a free-space gain of at least 7.0 dBi.

Before we select a line value, let's examine some of the free-space azimuth patterns yielded by the model. The 150-Ω, 3-MHz pattern shown in Fig 2, for example, is perfectly normal relative to expectations of pattern shape that we developed from looking at smaller models. To this point, we have come to expect the gain and front-to-back ratio of the array at 3 MHz to be lower than at every other frequency.

At 30 MHz, the 150-Ω pattern exhibits irregularities, as shown in Fig 2, indicating incipient side lobes. The spade-shaped forward lobe is also unusual. The irregularities grow more prominent with further reductions in transmission-line impedance. At 21 and 24 MHz, detectable side lobes appear in both the forward and rearward quadrants with a line impedance of 100 Ω, although they shrink to small irregularities with line values of 150 Ω.

Perhaps the worst case occurs at 18 MHz. The overlaid patterns in Fig 3 are for line values of 100 and 150 Ω. The triple rear lobes and side lobes on the forward lobe are clear for the lower line value. To a large measure, they diminish by increasing the line value to 150 Ω, although a smooth pattern is not obtained until the line value reaches 200 Ω. There are stub techniques for overcoming some pattern disturbances when using lower line-impedance values, but for the present exercise, they were not used.

In addition to pattern irregularities with low values of line impedance, the use of a 100-Ω line also produces a high source-impedance value (>100 Ω) at 3 MHz. This high impedance value shows up clearly when the SWR for the

Table 2—"Tau-tapered" element diameters for the 164-foot, 26-element 3-30 MHz LPDA.

Element No.	*Diameter (Inches)*	*Element No.*	*Diameter (Inches)*
26	0.50	13	1.91
25	0.56	12	2.12
24	0.62	11	2.34
23	0.69	10	2.59
22	0.76	9	2.87
21	0.85	8	3.18
20	0.94	7	3.52
19	1.04	6	3.90
18	1.15	5	4.32
17	1.27	4	4.79
16	1.41	3	5.30
15	1.56	2	5.87
14	1.73	1	6.50

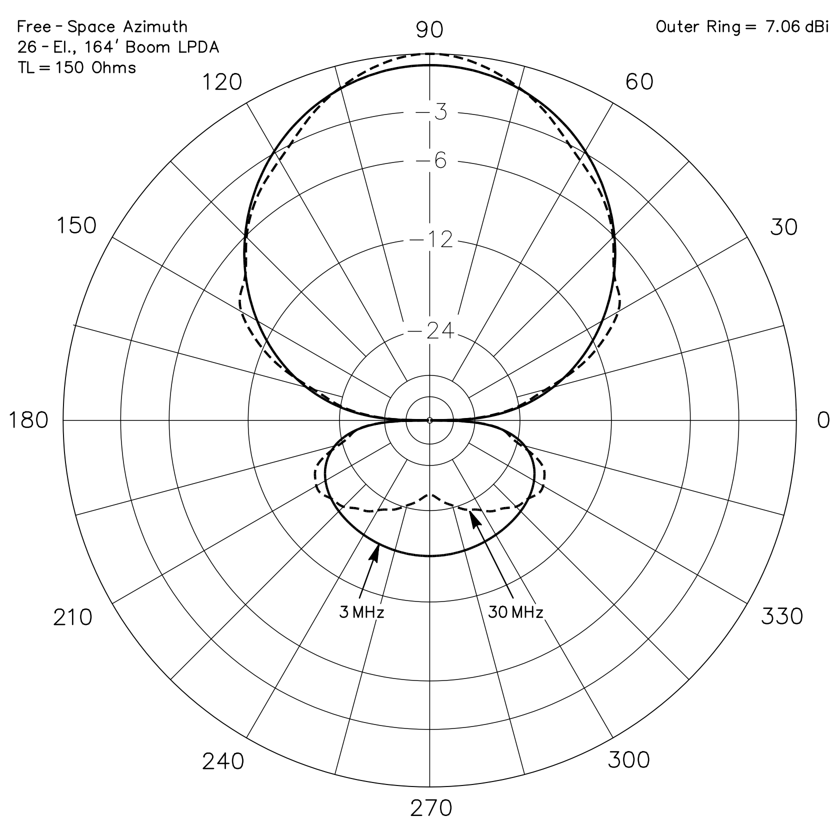

Fig 2—Free-space azimuth patterns of the 164-foot, 26-element LPDA model at 3 and 30 MHz.

Table 3—Performance of the 164-foot, 26-element model LPDA at 3-MHz increments from 3 to 30 MHz with different phase-line characteristic impedances.

Frequency	*Inter-Element Transmission Line Impedance (Ω)*			
	100	*150*	*200*	*250*
3 MHz				
Gain	6.57*	6.52	6.47	6.42
Front-to-Back	17.1*	16.5	16.1	15.7
Impedance	104. –*j*1	84. –*j*32	82. +*j*0	120. +*j*36
6 MHz				
Gain	7.00*	6.99	6.97	6.95
Front-to-Back	22.4	22.8	23.0	23.1*
Impedance	59. +*j*6	104. –*j*14	134. –*j*14	126. –*j*30
9 MHz				
Gain	7.81*	7.51	7.33	7.22
Front-to-Back	27.3*	26.2	25.6	25.3
Impedance	57. –*j*4	83. +*j*12	125. +*j*14	153. –*j*13
12 MHz				
Gain	7.85*	7.49	7.26	7.12
Front-to-Back	27.0	27.3	27.5	27.6*
Impedance	75. –*j*12	80. –*j*14	90. +*j*1	114. +*j*18
15 MHz				
Gain	7.99*	7.59	7.34	7.19
Front-to-Back	31.9*	29.8	29.2	29.0
Impedance	70. –*j*12	78. –*j*11	92. +*j*2	118. +*j*16
18 MHz				
Gain	7.54*	7.38	7.22	7.11
Front-to-Back	21.3	26.1	28.6	29.8*
Impedance	80. +*j*4	104. –*j*16	106. –*j*27	106. –*j*23
21 MHz				
Gain	7.26*	7.06	6.90	6.82
Front-to-Back	35.3*	29.2	28.0	27.5
Impedance	58. –*j*10	70. –*j*3	87. +*j*9	114. +*j*21
24 MHz				
Gain	7.54*	7.12	6.83	6.63
Front-to-Back	31.4*	30.0	29.7	29.3
Impedance	53. –*j*5	73. +*j*2	99. +*j*5	125. –*j*3
27 MHz				
Gain	7.49*	7.01	6.72	6.58
Front-to-Back	24.9	26.1	26.4	26.6*
Impedance	41. +*j*1	55. +*j*14	77. +*j*29	110. +*j*40
30 MHz				
Gain	7.49*	7.06	6.81	6.67
Front-to-Back	27.4	27.4	27.6	27.6*
Impedance	47. –*j*17	56. –*j*12	69. –*j*3	89. +*j*3

100-Ω-line model is plotted from 3 to 30 MHz in 1-MHz steps, as the graph in Fig 4 demonstrates. The highest value of 50-Ω SWR other than at 3 MHz occurs at 29 MHz: only 1.84:1. The use of a shorted terminating stub (90 inches of 100-Ω line) reduces the 3-MHz impedance to 62 –*j*19 Ω, well within the 2:1 SWR range desired; however, there are alternatives to the use of a terminating stub.

If the SWR curve is referenced to 65 Ω, as it is in Fig 5, no value of SWR rises above 1.6:1. As noted in Part 1, none of the impedance values have changed, but the reference impedance for the smoothest curve may help determine the best way to match the antenna to the main feed line for the array.

The SWR curve for a 150-Ω line is interesting when the reference value is 75 Ω. See Fig 6. The highest value of SWR is 1.6:1, and the SWR exceeds 1.5:1 at only three of the frequencies checked in the 1-MHz-increment sweep. Consequently, direct feed of the system with a 75-Ω main feed line is feasible.

The SWR curve for a 200-Ω line when referenced to 95 Ω is even smoother, as shown in Fig 7. This curve suggests the use of a 2:1 balun at the feed point with a 50-Ω main feed line. A reminder is due here. Although checks at 1-MHz intervals indicate a smooth curve, they do not guarantee that every intermediate frequency will be as well behaved. Were one to seriously consider implementing one or more of these study designs, a more thorough sweep would be in order.

Perhaps the best compromise among the criteria of gain, pattern smoothness and SWR is achieved by the version with a 150-Ω interelement transmission line. It provides about 7 dBi of free-space gain from 6 MHz upward, with only the low end of the band exhibiting lesser performance. However, should such a design be used with "Tau-tapered" wire elements according to the suggestion made in Part 1, one might expect at least a 0.1 to 0.2 dB reduction in gain, since the effects of the small gain deficit for the wire equivalents are cumulative for all active elements at any particular frequency.

164-foot, 42-Element LPDA

Although the 26-element LPDA might fulfill about 90% of the demands for such an array, it still suffers reduced gain and front-to-back ratio below 6 MHz. Therefore, it seemed appropriate to see whether an alternative design might show improvements in this regard. It would be obvious to try a design having about the same element density as the 100-foot 26-element design of Part 1. The result is a 42-element model with a Tau of 0.94 and a Sigma of 0.032. The outline of the design appears in Fig 8. Table 4 lists the element half-lengths and cumulative spacing for this model.

Standard LPDA design theory, as implemented in *LPCAD*, predicts a free-space gain of about 7.5 dBi and a front-to-back range of 22 to 28 dB. For the most part, the actual antenna as modeled in *NEC-4* will show less gain and superior front-to-back ratios.

In accord with the 26-element model, the 42-element model used "Tau-tapered" element diameters. Because of the greater number of elements, the tapering schedule differs, as Table 5 shows.

When checked across a set of line-impedance values and at 3-MHz intervals across the passband, the element-dense LPDA exhibits some interesting properties, some of which are at odds with the 26-element model. For example, the most favored interelement transmission-line characteristic impedance will be higher, rather than lower. Likewise, the best performance occurs at the lower end of the 3-30 MHz passband. The results are summarized in Table 6, which is a corresponds to Table 3 for the preceding model.

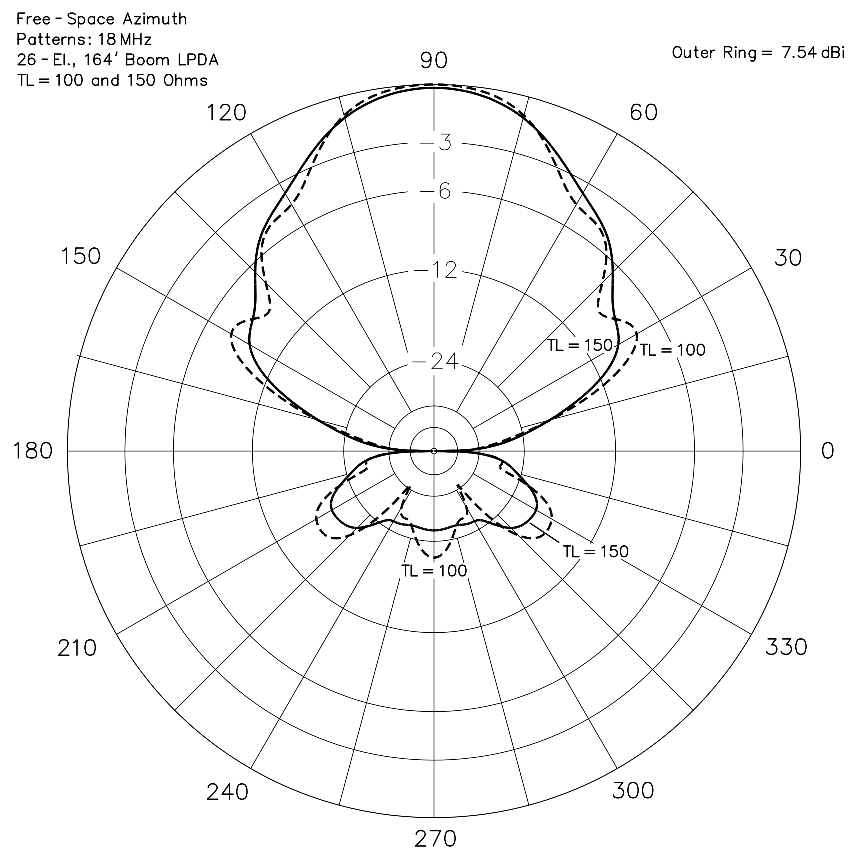

Fig 3—Free-space azimuth pattern of the 164-foot, 26-element LPDA model at 18 MHz with 100 Ω and 150 Ω transmission lines.

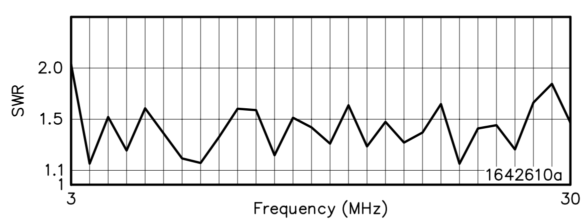

Fig 4—3-30 MHz SWR sweep of the 164-foot, 26-element LPDA model referenced to 50 Ω with a 100-Ω phase line.

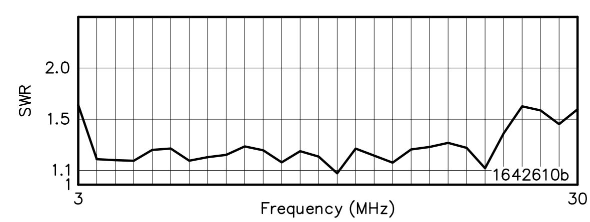

Fig 5—3-30 MHz SWR sweep of the 164-foot 26-element LPDA model referenced to 65 Ω with a 100-Ω phase line.

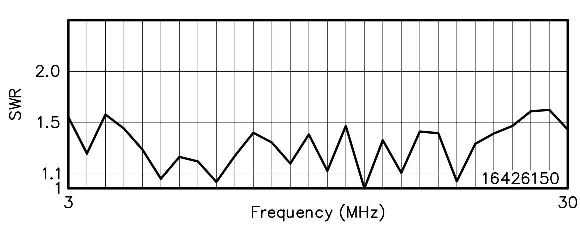

Fig 6—3-30 MHz SWR sweep of the 164-foot 26-element LPDA model referenced to 75 Ω with a 150-Ω phase line.

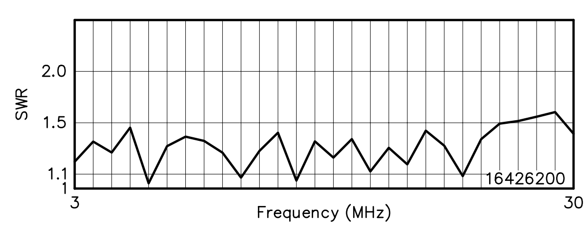

Fig 7—3-30 MHz SWR sweep of the 164-foot 26-element LPDA model referenced to 95 Ω with a 200-Ω phase line.

Perhaps the most notable trend is that the greatest number of peak values of gain and front-to-back ratio occur when using a 250-Ω transmission line (9 of 20 possible values). Indeed, 4 of the 10 gain peaks occur with the highest transmission-line impedance. Although the highest gain value for 3 MHz occurs with a 100-Ω line, its free-space gain is still well above 7.0 dBi with a very satisfactory front-to-back ratio when the array uses a 250-Ω line. Therefore, we may focus our attention on this version of the array in further comments on performance.

Unlike all other models of 3-30 MHz LPDAs that we have examined, the free-space azimuth pattern for 3 MHz (using the 250-Ω line) is a model of Yagi-like behavior. (See Fig 9.) The worst-case front-to-back ratio is above 22 dB at the peak of the rear lobes. Up to about 9 to 10 MHz, this particular design shows performance superior to the 26-element design.

Nonetheless, it is perhaps unreasonable to expect a 3.5-octave LPDA of standard design to show such characteristics throughout its entire passband. We may sample another pattern or two to see the array return to behavior more normal for LPDAs.

The free-space azimuth pattern for 15 MHz in Fig 9 is quite well behaved. However, the rear lobes show the beginning of irregularity relative to standard Yagi-based expectations of smoothly curved lobes. The rear lobes in this case are a bit blocky, but perhaps less so than some of the models we explored in Part 1. Incidentally, 15 MHz is about the frequency beyond which the large array ceases to show superior performance relative to the 26-element model we examined earlier. As Table 6 reveals, free-space gain from 15 through 30 MHz rarely equals that obtainable from the 26-element model with its more optimal Sigma value.

The free-space azimuth pattern for 27 MHz in Fig 9 shows evidence of the “blockiness” and irregularities that we tend to expect from standard-design LPDAs that do not use compensating stubs or other corrections. Although the pattern might be perfectly acceptable for virtually every application, a certain “squaring” of both the forward and rear lobes is evident. The blocking of the pattern would be even more evident if we had a comparable monoband Yagi pattern to place over this LPDA pattern.

Nonetheless, the 42-element, 164-foot-long LPDA design shows greater pattern control than the 26-element model of the same length. There are no instances of spade-shaped forward lobes and no tendencies toward detectable side lobes. Irregularities are minor, compared to the pattern outlines of the 26-element model at upper frequencies in the passband at the most favored line impedances. Despite the array’s lower gain at upper HF frequencies, the 42-element array provides very good pattern control throughout its range.

The consequences of lower performance at upper HF frequencies extend to the SWR curve, as we can see in Fig 10. When using a 75-Ω standard, we can reduce the peak SWR value at 29 MHz to 2.10:1; however, the remainder of the curve shows irregularities that are absent if we select 100 Ω as the reference impedance.

The 100-Ω curve in Fig 10 shows exceptionally good values up to about 24 MHz. Only at 23 MHz does the SWR exceed 1.4:1. Most of the values above 24 MHz are greater than 1.4:1, with

Table 4—Element half-lengths and cumulative spacing of the 164-foot, 42–element 3–30 MHz LPDA model.

Element No.	*Half Length (inches)*	*Cumulative Spacing (inches)*
1	1003.68	0.00
2	943.46	128.47
3	888.35	249.23
4	833.64	362.75
5	783.62	469.46
6	736.60	569.76
7	692.41	664.05
8	650.86	752.67
9	611.81	835.99
10	575.10	914.30
11	540.60	987.91
12	508.16	1057.11
13	477.67	1122.15
14	449.01	1183.29
15	422.07	1240.77
16	396.75	1294.79
17	372.94	1345.58
18	350.57	1393.31
19	329.53	1438.18
20	309.76	1480.36
21	291.17	1520.01
22	273.70	1557.28
23	257.20	1592.32
24	241.84	1625.25
25	227.33	1656.21
26	213.69	1685.30
27	200.87	1712.66
28	188.82	1738.37
29	177.49	1762.54
30	166.84	1785.26
31	156.83	1806.61
32	147.42	1826.69
33	138.58	1845.56
34	130.26	1863.29
35	122.45	1879.97
36	115.10	1895.64
37	108.19	1910.37
38	101.70	1924.22
39	95.60	1937.24
40	89.86	1949.48
41	84.47	1960.98
42	79.40	1971.79

29 MHz showing a value of 2.8:1 relative to the 100-Ω standard—a function of the low resistive value of the impedance at that frequency. One must begin to suspect that the element density at the higher end of the HF range for this design is not optimal for either gain or source impedance.

We may graph the gain values for the two models and derive a bit more data, as evidenced by Fig 11. The graph uses the 250-Ω version of the 42-element array and the 150-Ω version of the 26-element array as perhaps the best of each design. Except for frequencies below about 7 MHz, the 26-element design shows considerably higher gain at most frequencies. The gain of the 42-element model remains consistent until just past 21 MHz, after which it decreases rapidly. The 26-element design shows peak gain between 9 and 19 MHz, but drops below 7 dBi only at 3 and 6 MHz.

With respect to 180° front-to-back ratio (shown in Fig 12), the 42-element array shows the more consistent curve. The advantage of the 42-element model is especially clear below 6 MHz. Above that frequency, both antennas show values of front-to-back ratio that would satisfy the most stringent operating specifications: values in excess of 25 dB across most of the passband.

Because the graphs employ the most favored version of each array, they cannot display certain design weaknesses. For example, the 42-element LPDA design shows periodic depressions in its gain. These depressions show themselves most clearly through the tabular performance listings for a 100-Ω interelement transmission line. Gain values drop well below 7 dBi at 9, 18 and 27 MHz and at 15 and 30 MHz.

Table 5—"Tau-tapered" element diameters for the 164-foot, 42-element 3-30 MHz LPDA.

Element	*Diameter (Inches)*	*Element*	*Diameter (Inches)*	*Element*	*Diameter (Inches)*
42	0.50	28	1.22	14	2.91
41	0.55	27	1.30	13	3.09
40	0.58	26	1.38	12	3.29
39	0.62	25	1.47	11	3.50
38	0.66	24	1.57	10	3.72
37	0.70	23	1.67	9	3.96
36	0.75	22	1.77	8	4.22
35	0.79	21	1.89	7	4.48
34	0.84	20	2.00	6	4.77
33	0.90	19	2.13	5	5.07
32	0.95	18	2.27	4	5.40
31	1.02	17	2.42	3	5.74
30	1.08	16	2.57	2	6.11
29	1.15	15	2.73	1	6.50

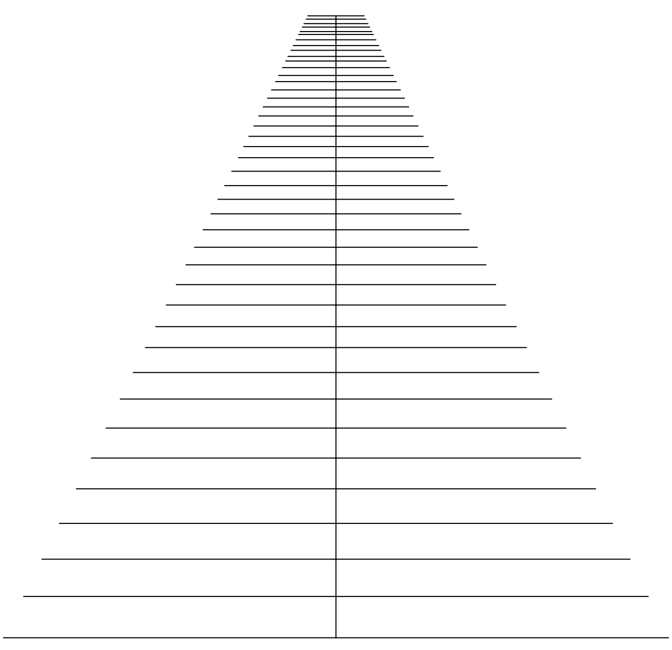

Fig 8—Outline of the 164-foot, 42-element 3-30 MHz LPDA; Tau = 0.94, Sigma = 0.032.

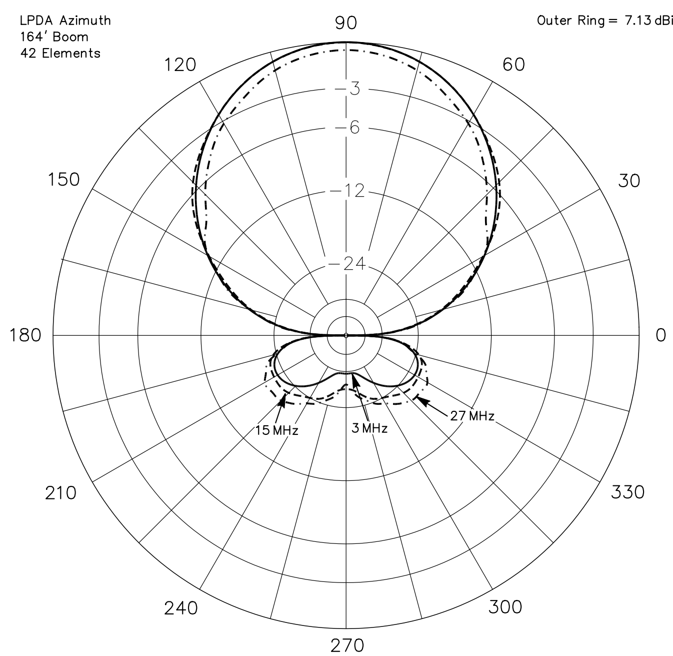

Fig 9—Free-space azimuth pattern of the 164-foot, 42-element LPDA model at 3, 15 and 27 MHz.

Both sets of lower gain values are harmonically related. To what degree harmonically related phenomena are endemic to long LPDA designs with higher element densities would be found through the examination of many more very large models.

Given our limited choice of two designs, the wider-spaced 26-element model shows fewer weaknesses than the larger design, despite the superior performance of the 42-element model at frequencies below 9 MHz. With a 150-Ω interelement transmission line, the 26-element model with "Tau-tapered" element diameters might well meet many sets of operational specifications.

Optimizing an LPDA

Before closing the book on the 164-foot, 26-element design, it may be interesting to flirt with the temptation to manually optimize the lower-frequency performance of the array. With the calculated lengths, the two longest elements are resonant at roughly 2.8 and 3.1 MHz. The high length-to-diameter ratio might suggest that we lengthen the rear element, but, in fact, precisely the opposite tack is required to improve performance at 3 MHz.

How far one should carry an optimizing process—especially when performed manually—is a matter of judgment. At 3 MHz, the original design showed a free-space gain of 6.52 dBi when using the 150-Ω transmission line. The front-to-back ratio was 16.5 dB. For the present exercise, the goal was a front-to-back ratio of at least 20 dB, with whatever gain increase the process might yield. The goal was achieved with a resultant gain of 6.75 dBi.

The final changes involved only the rear two elements. The rearmost element began at 2007 inches. It was shortened to 1960 inches and moved rearward 24 inches (thus increasing the antenna length by 2 feet). The second-longest element was increased from 1811.6 to 1814 inches. No other changes were necessary to produce the reported gain and front-to-back ratio. Source impedance was 75 $-j39$ Ω, an easily manageable value when referenced to 75 Ω in accord with the original model. See Table 7 for a listing of element half-lengths and cumulative spacing.

Spot optimization of LPDAs is a somewhat dangerous process, though, unless a thorough frequency sweep is performed for both the original and final products. Table 8 lists perform-

Table 6—Performance of the 164-foot, 42-element model LPDA at 3-MHz increments from 3 to 30 MHz with different phase-line characteristic impedances.

Frequency	*Inter-Element Transmission Line Impedance (ohms)*			
	100	*150*	*200*	*250*
3 MHz				
Gain	7.26*	7.21	7.17	7.13
Front-to-Back	32.6	36.0	36.9*	35.0
Impedance	61. $+j8$	79. $-j15$	72. $+j6$	119. $+j27$
6 MHz				
Gain	7.22	7.22	7.22	7.22
Front-to-Back	24.5*	24.5	24.4	24.4*
Impedance	67. $+j8$	75. $-j14$	77. $+j5$	115. $+j16$
9 MHz				
Gain	6.77	6.99	7.06	7.10*
Front-to-Back	25.2	25.4	25.5	25.6*
Impedance	55. $-j10$	67. $+j8$	107. $+j8$	110. $-j22$
12 MHz				
Gain	7.03*	7.25	7.29*	7.25
Front-to-Back	27.8	27.9	28.1	28.2*
Impedance	59. $+j10$	87. $-j11$	78. $-j14$	87. $+j5$
15 MHz				
Gain	6.71	6.90	7.02	7.12*
Front-to-Back	29.2*	29.1	29.0	29.1
Impedance	56. $-j10$	61. $+j2$	90. $+j14$	121. $-j6$
18 MHz				
Gain	6.56	6.80	6.90	6.96*
Front-to-Back	30.5	30.5	30.8	30.8*
Impedance	68. $-j3$	67. $-j15$	71. $-j3$	93. $+j7$
21 MHz				
Gain	7.02	7.11*	7.08	7.00
Front-to-Back	31.2*	31.2	30.9	30.4
Impedance	46. $-j9$	70. $+j4$	87. $+j5$	105. $-j14$
24 MHz				
Gain	6.72	6.81*	6.76	6.64
Front-to-Back	30.5*	30.3	30.1	29.7
Impedance	43. $-j11$	54. $+j2$	80. $+j8$	104. $-j9$
27 MHz				
Gain	6.56	6.69	6.71*	6.66
Front-to-Back	29.6	29.8	30.1	30.1*
Impedance	74. $-j11$	59. $-j27$	54. $-j16$	63. $-j2$
30 MHz				
Gain	5.91	6.02	6.11	6.20*
Front-to-Back	26.5	26.9	27.0	27.3*
Impedance	20. $+j4$	59. $-j27$	53. $+j41$	101. $+j53$

ance figures for the original and the modified 26-element designs using a 150-Ω line. As an additional reference, performance values are also listed for the 42-element design. The checkpoints for this table are in 1-MHz increments between 3 and 9 MHz. The remaining values from 10 to 30 MHz for the 26-element design change by less than 0.03 dB in gain and 0.2 dB in front-to-back ratio. These changes are operationally insignificant.

The third column in Table 8 is simply a convenient way to confirm the consistent performance of the 42-element array in the lower third of the spectrum for which the antenna is designed. The chief purpose of the table, however, is to demonstrate a performance anomaly created by manual-optimizing activity. Although 3-MHz performance has been improved by about 0.25 dB in gain and by 3.5 dB in front-to-back ratio, the modification of the two rearmost elements has seriously disturbed 5 MHz performance. Despite the fact that neither element seems relevant to 5-MHz performance (since their resonances are at least 2 MHz lower), the 5-MHz front-to-back ratio has dropped below 10 dB: a 5.5 dB decrease.

One of the enduring myths about LPDA designs is that only the two elements closest to the one nearest resonance are truly active. An intermediate frequency might bring four elements into play. Actually, all elements are active all the time at all frequencies. At least two to three elements on each side of a nearly resonant element carry significant current and are very influential on performance. Resonance and pure harmonic relationships do not appear to be requisites for a distant element to affect the performance of the array on various frequencies.

In the present case, the modifications have adverse consequences on 5-MHz performance. Had the frequency region of reduced front-to-back ratio not appeared on one of the spot-check frequencies, it might well have been missed. As a result, spot modifications must be undertaken with great care, and a subsequent frequency sweep of the final design at all frequencies of interest is recommended.

The 164-foot, 26 Element LPDA: More thorough Sweeping

Because of the suggestion above, I undertook a performance sweep of the 164-foot, 26-element model. Among the models evaluated, this model appeared to be the most promising with respect

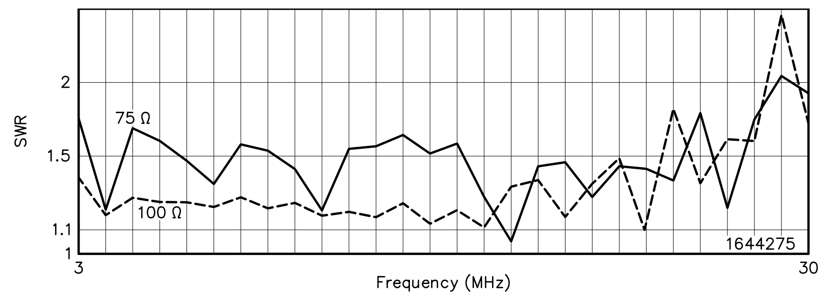

Fig 10—3-30 MHz SWR sweep of the 164-foot, 42-element LPDA model referenced to 75 Ω (solid line) and 100 Ω (dashed) with a 250-Ω phase line.

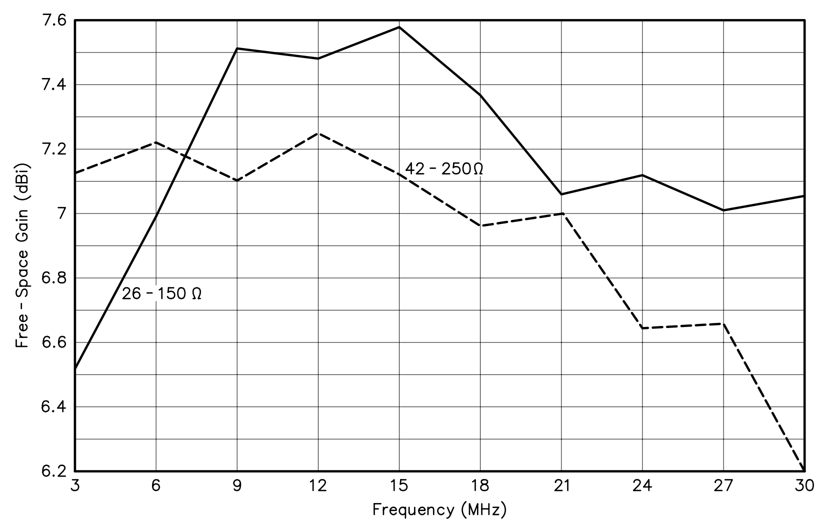

Fig 11—Comparison of the free-space gain of the 26-element (solid) and the 42-element (dashed), 164-foot LPDA models at 3 MHz intervals from 3 to 30 MHz.

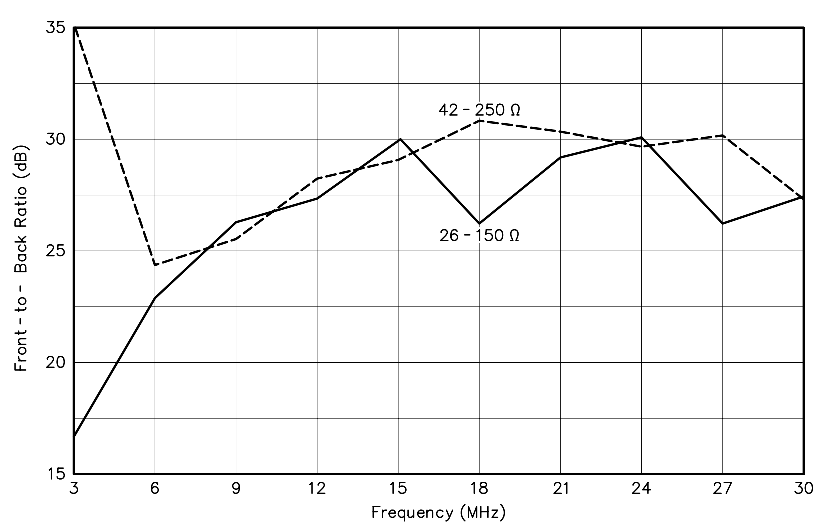

Fig 12—Comparison of the front-to-back ratio of the 26-element (solid) and the 42-element (dashed), 164-foot LPDA models at 3 MHz intervals from 3 to 30 MHz.

to gain, front-to-back ratio, SWR curve and pattern control, especially if used with a 150-Ω interelement transmission line.

For the frequency sweeps, I used the basic model with no terminating stub. I also modified the model by adding a 40-inch shorted stub of 150-Ω transmission line to the center of the longest element. The stub length was experimentally determined in the model to yield slightly improved performance at 3 MHz. The goal was to see whether the stub had any significant effects on antenna performance at frequencies distant from 3 MHz.

For the sweep, I chose frequency intervals of 0.5 MHz from 3 to 30 MHz. Although these frequencies are farther apart than one might wish for truly detailed analysis within a band of interest, smaller increments would not have produced readable graphs.

The graph of free-space gain across the antenna's passband is a case in point and appears in Fig 13. Even with 0.5-MHz increments, one must use great care in tracing the curves. Most notably, the graph reveals some significant gain deviations compared with the 3-MHz increments of earlier performance checks. The version with no stub shows low gain (well below 7 dBi) at 7 MHz, and both versions show lower gain in the 26 to 26.5 MHz region and again in the 28.5 to 29.5 MHz region. These deficits, of course, did not appear in the earlier checks. Also notable is the unusually high gain (7.96 dBi) at 8.5 MHz in the version of the antenna with the stub attached.

Otherwise, the two plots overlay each other quite closely. Above 10 MHz, a few spot values differ by as much as 0.15 dB, but most differences are below 0.05 dB. Neither level of difference is operationally significant.

The graph of 180° front-to-back ratio in Fig 14 shows the consistently high front-to-back ratio of the design. The stub-less model shows noticeably lower values at 5, 7.5 and 8 MHz, with a surprisingly higher value at 4.5 MHz. A shorted stub can smooth the front-to-back performance of the antenna below 10 MHz, suggesting that adding the stub to the system has merit.

Each model shows a frequency region where the front-to-back ratio reaches a peak at about 35 dB. Interestingly, the model with the stub raises the frequency of this peak by about 4 MHz without creating significant changes in the antenna gain. Since the typical rear-quadrant pattern shows lobes to each side of the maximum front-to-back ratio, the peaks are of numeric rather than operational significance. Even with the exceptions noted for the stub-less model and the decreased values at the lowest frequency of use, the model shows a front-to-rear ratio that is consistently 20 dB or better.

The graphs of resistance (Fig 15) and reactance (Fig 16) at the feed point of the antenna require less examination. From 9 MHz upward, they so closely overlap that a single line suffices for both models. Below 9 MHz, the feed point resistance plots very closely

Table 7—Element half-lengths and cumulative spacing of the modified 164-foot, 26-element 3-30 MHz LPDA model.

Element	*Half Length (inches)*	*Cumulative Spacing (inches)*
1	980.00	–24.00
2	907.00	208.31
3	817.49	396.32
4	737.77	565.99
5	655.83	719.11
6	600.91	857.30
7	542.31	982.02
8	489.43	1094.58
9	441.71	1196.16
10	398.64	1287.84
11	359.76	1370.57
12	324.68	1445.24
13	293.02	1512.63
14	264.45	1573.45
15	238.66	1628.34
16	215.39	1677.87
17	194.39	1722.58
18	175.43	1762.92
19	158.33	1799.33
20	142.89	1832.19
21	128.96	1861.85
22	116.38	1888.62
23	105.03	1912.77
24	94.79	1934.57
25	85.55	1954.24
26	77.21	1972.00

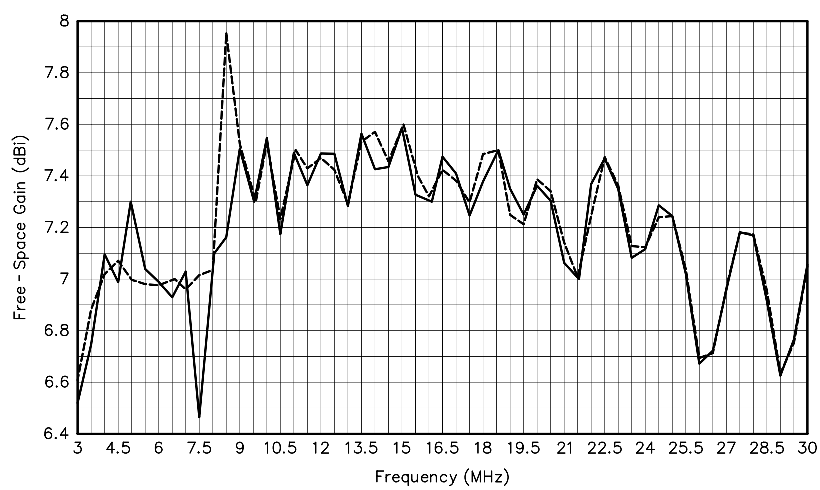

Fig 13—Free-space gain of the 26-element, 164-foot LPDA model without (solid) and with (dashed) a shorted stub at 0.5 MHz intervals. See text for stub dimensions.

coincide, with the stub-model tending to reduce the resistance slightly from the values for the stub-less model.

The stub makes a greater difference in the value of reactance at the feed point below 9 MHz. The reactance yielded by each model shows the greatest divergence at 3, 5 and 7.5 MHz; however, in no case does the reactance value exceed the maximum values found on the main plot—either inductively or capacitively. Consequently, we should expect that SWR curves for either version will overlap considerably, regardless of the reference impedance value we choose for the plot.

The rough coincidence of resistance and reactance values between models with and without a stub would show no differences worth noting in a pair of SWR curves set to the same reference impedance value. Therefore, I set the reference impedance value for the stub-less model to 100 Ω and the value for the model with the stub to 75 Ω. A 100-Ω value is useful to designers because one may introduce a wide-band 2:1 matching device at the feed point and feed the array with 50-Ω coaxial cable. The 75-Ω standard tends to imply direct feed with 75-Ω cable or a 1.4:1 wide-band matching device.

Fig 17 shows the resulting SWR plots. The 100-Ω curve shows peak values of SWR above 1.8:1 at 26.5, 27 and 30 MHz, with all other values below 1.5:1. In contrast, the 75-Ω curve shows more variance among values at all frequencies, but it displays peaks above 1.6:1 only at 4.5 and 28.5 MHz. For most purposes, either approach—and the matching techniques implied by it—would prove satisfactory for the antenna.

Conclusion

Bringing "preliminary notes" to a conclusion is nearly a contradiction in terms. The purpose of this exercise has been to see what light is shed by method-of-moments modeling on standard, 3.5-octave HF LPDA design. The notes have surveyed array lengths from 60 to 164 feet with 20 to 42 elements. However, limitations imposed by the size of the models preclude anything close to sufficient coverage of array sizes between those selected for modeling. Likewise, run time for the large models limited the frequencies at which performance checks were made.

Nonetheless, the collection of models has demonstrated both the potential and some weaknesses of conventional LPDA designs that are limited in array length. Strong low-end performers

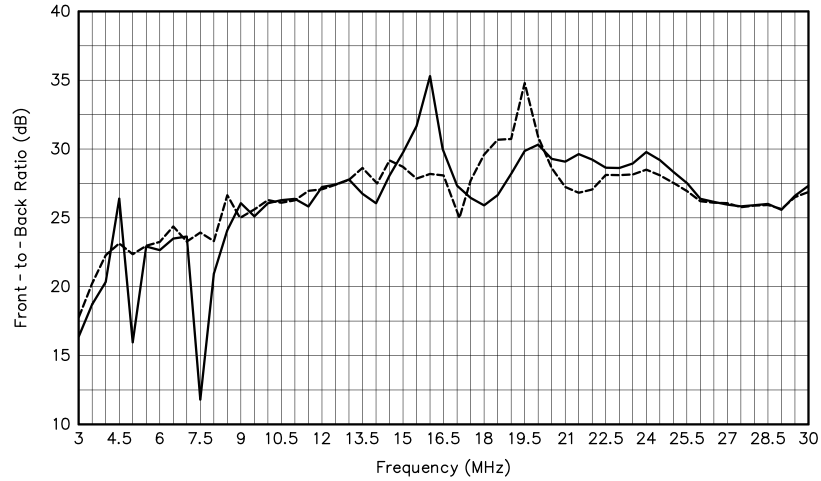

Fig 14—Front-to-back (180°) ratio of the 26-element, 164-foot LPDA model without (solid) and with (dashed) a shorted stub at 0.5 MHz intervals. See text for stub dimensions.

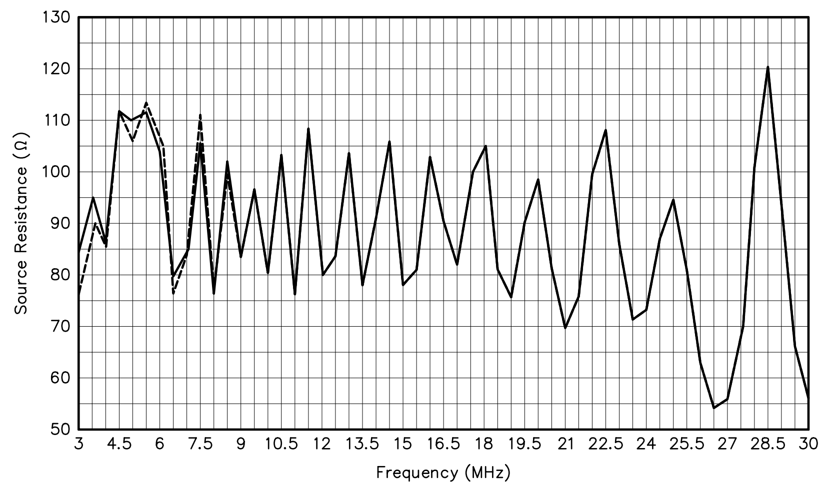

Fig 15—Source resistance of the 26-element, 164-foot LPDA model without (solid) and with (dashed) a shorted stub at 0.5 MHz intervals. See text for stub dimensions.

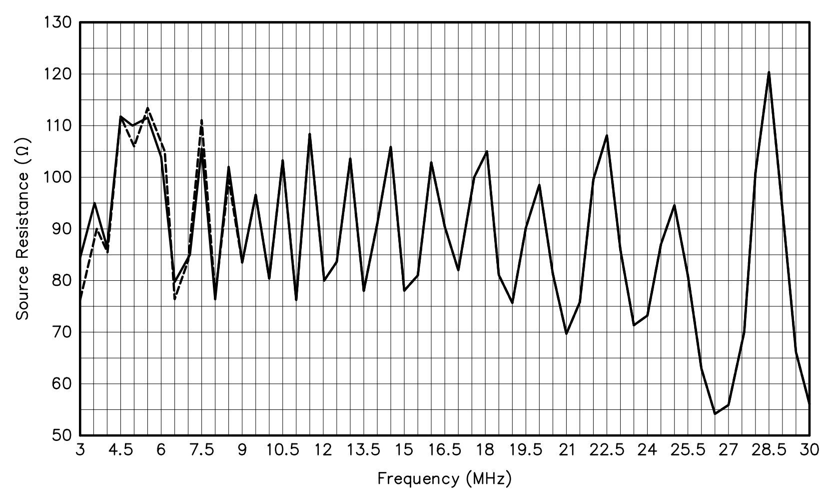

Fig 16—Source reactance of the 26-element, 164-foot LPDA model without (solid) and with (dashed) a shorted stub at 0.5 MHz intervals. See text for stub dimensions.

showed weaknesses at higher frequencies, while designs that performed strongly at the upper end of the passband were weaker performers at the lowest frequencies. Of the models surveyed, perhaps the original 164-foot 26-element design with "Tau-tapered" elements deserves the most attention. Undoubtedly, careful redesign can tweak its performance even further.

In addition, numerous alternative design techniques exist that have not been covered in these preliminary notes. Hybrid designs and designs using tapered Tau and Sigma values have yet to be explored. The results of these explorations may well be alternative design algorithms that may yield smoother gain performance across the 3-30 MHz spectrum. Other directions still to be examined involve setting the value of Tau to be referenced to frequencies of interest and altering the linear nature of the elements themselves. A short bibliography of both basic and innovative design ideas appears at the end of this article.

One question has been intentionally bypassed: Are any of the better designs mechanically workable? Although a 164-foot rotatable boom is not easily made feasible, its construction may be possible. At a lesser gain, the 100-foot, 26-element model may also be practical under certain circumstances. With wire elements, even the longest design might serve as a fixed-position beam.

This has been a design study attempting to bring *NEC-4* to bear on standard LPDA designs for antennas having a wide frequency range. It cannot be complete, but perhaps it may serve as a beginning for better understanding of standard LPDA performance throughout the HF range.

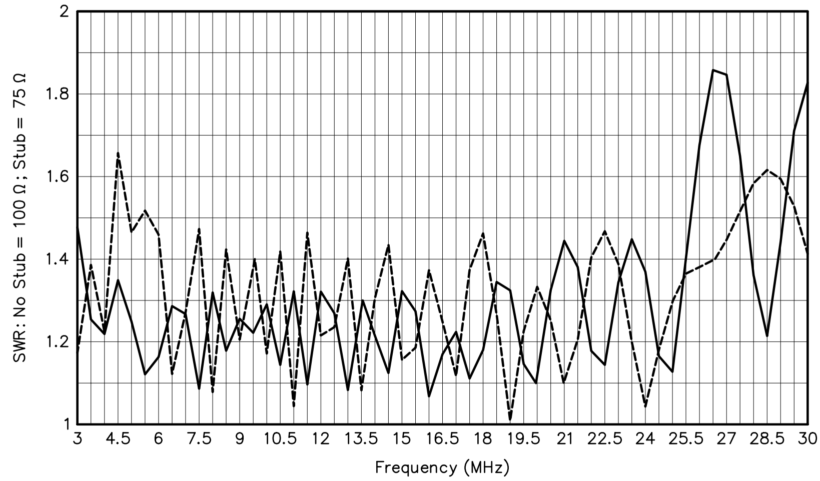

Fig 17—SWR curves of the 26-element, 164-foot LPDA model without (solid) and with (dashed) a shorted stub at 0.5 MHz intervals. See text for stub dimensions. The "no-stub" model is referenced to 100 Ω; the "with-stub" model is referenced to 75 Ω.

Table 8—Spot performance checks at 1-MHz intervals from 3 to 9 MHz for the initial and modified 26-element LPDA models and for the 42-element LPDA model.

	Antenna Design		
	Original 26	*Modified 26*	*42*
3 (MHz)			
Gain	6.52	6.75	7.13
Front-to-Back	16.5	20.0	35.0
Impedance	83. –*j*32	75. –*j*39	119. –*j*27
4 (MHz)			
Gain	7.10	7.11	7.24
Front-to-Back	20.5	20.2	23.6
Impedance	86. +*j*12	86. +*j*12	88. –*j*6
5 (MHz)			
Gain	7.30	7.19	7.22
Front-to-Back	15.7	9.4	23.8
Impedance	110. –*j*21	104. –*j*31	125. +*j*3
6 (MHz)			
Gain	6.99	6.99	7.22
Front-to-Back	22.8	22.7	24.4
Impedance	104. –*j*14	104. –*j*14	115. +*j*16
7 (MHz)			
Gain	7.03	7.02	7.23
Front-to-Back	23.8	23.5	24.5
Impedance	85. –*j*16	85. –*j*16	101. –*j*21
8 (MHz)			
Gain	7.11	7.08	7.23
Front-to-Back	21.0	21.0	24.8
Impedance	76. +*j*4	76. +*j*4	92. –*j*16
9 (MHz)			
Gain	7.51	7.57	7.10
Front-to-Back	26.2	28.3	25.6
Impedance	83. +*j*12	83. +*j*12	110. –*j*22

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15. J. D. Kraus, *Antennas*, 2nd Ed. (New York: McGraw-Hill, 1988), Chapter 15.
16. V. H. Rumsey, *Frequency Independent Antennas* (New York: Academic Press, 1966).
17. R. Dean Straw, Ed., *The ARRL Antenna Book*, 18th Ed. (Newington: ARRL, 1997), Chapter 10. ARRL Order #6133, $30.

Most standard college texts on basic antenna theory and practice have a chapter devoted to the fundamentals of LPDA design.

Appendix

You can download this package from the ARRL Web site: **http://www.arrl.org/files/*qex*/**. Look for LPDASPT2.ZIP.

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Antenna Model Descriptions

164' 26-Element 3-30 MHz LPDA                          Frequency = 3  MHz.

Wire Loss: Aluminum  Resistivity = 4E08 ohmm, Rel. Perm. = 1

               WIRES
Wire Conn.  End 1 (x,y,z : in)  Conn.  End 2 (x,y,z : in)   Dia(in) Segs
1          1003.7,  0.000,  0.000       1003.68,  0.000,  0.000 6.50E+00 107
2          905.81,208.310,  0.000       905.810,208.310,  0.000 5.87E+00  97
3          817.49,396.320,  0.000       817.490,396.320,  0.000 5.30E+00  87
4          737.77,565.990,  0.000       737.770,565.990,  0.000 4.79E+00  79
5          655.83,719.110,  0.000       655.830,719.110,  0.000 4.32E+00  71
6          600.91,857.300,  0.000       600.910,857.300,  0.000 3.90E+00  65
7          542.31,982.020,  0.000       542.310,982.020,  0.000 3.52E+00  57
8          489.43,1094.58,  0.000       489.430,1094.58,  0.000 3.18E+00  53
9          441.71,1196.16,  0.000       441.710,1196.16,  0.000 2.87E+00  47
10         398.64,1287.84,  0.000       398.640,1287.84,  0.000 2.59E+00  43
11         359.76,1370.57,  0.000       359.760,1370.57,  0.000 2.34E+00  39
12         324.68,1445.24,  0.000       324.680,1445.24,  0.000 2.12E+00  35
13         293.02,1512.63,  0.000       293.020,1512.63,  0.000 1.91E+00  31
14         264.45,1573.45,  0.000       264.450,1573.45,  0.000 1.73E+00  29
15         238.66,1628.34,  0.000       238.660,1628.34,  0.000 1.56E+00  25
16         215.39,1677.87,  0.000       215.390,1677.87,  0.000 1.41E+00  23
17         194.39,1722.58,  0.000       194.390,1722.58,  0.000 1.27E+00  21
18         175.43,1762.92,  0.000       175.430,1762.92,  0.000 1.15E+00  19
19         158.33,1799.33,  0.000       158.330,1799.33,  0.000 1.04E+00  17
20         142.89,1832.19,  0.000       142.890,1832.19,  0.000 1.15E+00  15
21         128.96,1861.85,  0.000       128.960,1861.85,  0.000 1.04E+00  15
22         116.38,1888.62,  0.000       116.380,1888.62,  0.000 9.40E01  13
23         105.03,1912.77,  0.000       105.030,1912.77,  0.000 8.40E01  11
24         94.790,1934.57,  0.000        94.790,1934.57,  0.000 7.60E01  11
25         85.550,1954.24,  0.000        85.550,1954.24,  0.000 6.90E01   9
26         77.210,1972.00,  0.000        77.210,1972.00,  0.000 5.00E01   9

               SOURCES
Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.      Actual      (Specified)
1           5     26 / 50.00    ( 26 / 50.00)      1.000        0.000       V

                 TRANSMISSION LINES
Line  Wire #/% From End 1   Wire #/% From End 1    Length       Z0   Vel Rev/
      Actual  (Specified)   Actual  (Specified)              ohms Fact Norm
1       1/50.0  (  1/50.0)    2/50.0  (  2/50.0)  Actual dist  150.0  1.00  R
2       2/50.0  (  2/50.0)    3/50.0  (  3/50.0)  Actual dist  150.0  1.00  R
3       3/50.0  (  3/50.0)    4/50.0  (  4/50.0)  Actual dist  150.0  1.00  R
4       4/50.0  (  4/50.0)    5/50.0  (  5/50.0)  Actual dist  150.0  1.00  R
5       5/50.0  (  5/50.0)    6/50.0  (  6/50.0)  Actual dist  150.0  1.00  R
6       6/50.0  (  6/50.0)    7/50.0  (  7/50.0)  Actual dist  150.0  1.00  R
7       7/50.0  (  7/50.0)    8/50.0  (  8/50.0)  Actual dist  150.0  1.00  R
8       8/50.0  (  8/50.0)    9/50.0  (  9/50.0)  Actual dist  150.0  1.00  R
9       9/50.0  (  9/50.0)   10/50.0  ( 10/50.0)  Actual dist  150.0  1.00  R
10     10/50.0  ( 10/50.0)   11/50.0  ( 11/50.0)  Actual dist  150.0  1.00  R
11     11/50.0  ( 11/50.0)   12/50.0  ( 12/50.0)  Actual dist  150.0  1.00  R
12     12/50.0  ( 12/50.0)   13/50.0  ( 13/50.0)  Actual dist  150.0  1.00  R
13     13/50.0  ( 13/50.0)   14/50.0  ( 14/50.0)  Actual dist  150.0  1.00  R
14     14/50.0  ( 14/50.0)   15/50.0  ( 15/50.0)  Actual dist  150.0  1.00  R
15     15/50.0  ( 15/50.0)   16/50.0  ( 16/50.0)  Actual dist  150.0  1.00  R
16     16/50.0  ( 16/50.0)   17/50.0  ( 17/50.0)  Actual dist  150.0  1.00  R
17     17/50.0  ( 17/50.0)   18/50.0  ( 18/50.0)  Actual dist  150.0  1.00  R
18     18/50.0  ( 18/50.0)   19/50.0  ( 19/50.0)  Actual dist  150.0  1.00  R
19     19/50.0  ( 19/50.0)   20/50.0  ( 20/50.0)  Actual dist  150.0  1.00  R
20     20/50.0  ( 20/50.0)   21/50.0  ( 21/50.0)  Actual dist  150.0  1.00  R
21     21/50.0  ( 21/50.0)   22/50.0  ( 22/50.0)  Actual dist  150.0  1.00  R
22     22/50.0  ( 22/50.0)   23/50.0  ( 23/50.0)  Actual dist  150.0  1.00  R
23     23/50.0  ( 23/50.0)   24/50.0  ( 24/50.0)  Actual dist  150.0  1.00  R
24     24/50.0  ( 24/50.0)   25/50.0  ( 25/50.0)  Actual dist  150.0  1.00  R
25     25/50.0  ( 25/50.0)   26/50.0  ( 26/50.0)  Actual dist  150.0  1.00  R
Ground type is Free Space
```

```
164' 42-Element 3-30 MHz LPDA                              Frequency = 3  MHz.
Wire Loss: Aluminum  Resistivity = 4E08 ohmm, Rel. Perm. = 1

               WIRES
Wire Conn.  End 1 (x,y,z : in)  Conn.  End 2 (x,y,z : in)   Dia(in) Segs
1          1003.7,  0.000,  0.000       1003.68,  0.000,  0.000 6.50E+00 107
2          943.46,128.470,  0.000       943.460,128.470,  0.000 6.11E+00 101
3          888.35,249.230,  0.000       888.350,249.230,  0.000 5.74E+00  95
4          833.64,362.750,  0.000       833.640,362.750,  0.000 5.40E+00  89
5          783.62,469.460,  0.000       783.620,469.460,  0.000 5.07E+00  83
6          736.60,569.760,  0.000       736.600,569.760,  0.000 4.77E+00  79
7          692.41,664.050,  0.000       692.410,664.050,  0.000 4.48E+00  73
8          650.86,752.670,  0.000       650.860,752.670,  0.000 4.22E+00  69
9          611.81,835.990,  0.000       611.810,835.990,  0.000 3.96E+00  65
10         575.10,914.300,  0.000       575.100,914.300,  0.000 3.72E+00  61
11         540.60,987.910,  0.000       540.600,987.910,  0.000 3.50E+00  57
12         508.16,1057.11,  0.000       508.160,1057.11,  0.000 3.29E+00  55
13         477.67,1122.15,  0.000       477.670,1122.15,  0.000 3.09E+00  51
14         449.01,1183.29,  0.000       449.010,1183.29,  0.000 2.91E+00  47
15         422.07,1240.77,  0.000       422.070,1240.77,  0.000 2.73E+00  45
16         396.75,1294.79,  0.000       396.750,1294.79,  0.000 2.57E+00  43
17         372.94,1345.58,  0.000       372.940,1345.58,  0.000 2.42E+00  39
18         350.57,1393.31,  0.000       350.570,1393.31,  0.000 2.27E+00  37
19         329.53,1438.18,  0.000       329.530,1438.18,  0.000 2.13E+00  35
20         309.76,1480.36,  0.000       309.760,1480.36,  0.000 2.00E+00  33
21         291.17,1520.01,  0.000       291.170,1520.01,  0.000 1.89E+00  31
22         273.70,1557.28,  0.000       273.700,1557.28,  0.000 1.77E+00  29
23         257.20,1592.32,  0.000       257.200,1592.32,  0.000 1.67E+00  27
24         241.84,1625.25,  0.000       241.840,1625.25,  0.000 1.57E+00  25
25         227.33,1656.21,  0.000       227.330,1656.21,  0.000 1.47E+00  25
26         213.69,1685.30,  0.000       213.690,1685.30,  0.000 1.38E+00  23
27         200.87,1712.66,  0.000       200.870,1712.66,  0.000 1.30E+00  21
28         188.82,1738.37,  0.000       188.820,1738.37,  0.000 1.22E+00  21
29         177.49,1762.54,  0.000       177.490,1762.54,  0.000 1.15E+00  19
30         166.84,1785.26,  0.000       166.840,1785.26,  0.000 1.08E+00  17
31         156.83,1806.61,  0.000       156.830,1806.61,  0.000 1.02E+00  17
32         147.42,1826.69,  0.000       147.420,1826.69,  0.000 9.50E01  15
33         138.58,1845.56,  0.000       138.580,1845.56,  0.000 9.00E01  15
34         130.26,1863.29,  0.000       130.260,1863.29,  0.000 8.40E01  15
35         122.45,1879.97,  0.000       122.450,1879.97,  0.000 7.90E01  13
36         115.10,1895.64,  0.000       115.100,1895.64,  0.000 7.50E01  13
37         108.19,1910.37,  0.000       108.190,1910.37,  0.000 7.00E01  11
38         101.70,1924.22,  0.000       101.700,1924.22,  0.000 6.60E01  11
39         95.600,1937.24,  0.000        95.600,1937.24,  0.000 6.20E01  11
40         89.860,1949.48,  0.000        89.860,1949.48,  0.000 5.80E01  11
41         84.470,1960.98,  0.000        84.470,1960.98,  0.000 5.50E01  11
42         79.400,1971.79,  0.000        79.400,1971.79,  0.000 5.00E01  11

               SOURCES
Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.     Actual      (Specified)
1           6    42 / 50.00   ( 42 / 50.00)      1.000       0.000       V

                 TRANSMISSION LINES
Line  Wire #/% From End 1   Wire #/% From End 1    Length       Z0   Vel Rev/
      Actual  (Specified)   Actual  (Specified)               ohms Fact Norm
1      1/50.0  (  1/50.0)    2/50.0  (  2/50.0)  Actual dist  250.0  1.00  R
2      2/50.0  (  2/50.0)    3/50.0  (  3/50.0)  Actual dist  250.0  1.00  R
3      3/50.0  (  3/50.0)    4/50.0  (  4/50.0)  Actual dist  250.0  1.00  R
4      4/50.0  (  4/50.0)    5/50.0  (  5/50.0)  Actual dist  250.0  1.00  R
5      5/50.0  (  5/50.0)    6/50.0  (  6/50.0)  Actual dist  250.0  1.00  R
6      6/50.0  (  6/50.0)    7/50.0  (  7/50.0)  Actual dist  250.0  1.00  R
7      7/50.0  (  7/50.0)    8/50.0  (  8/50.0)  Actual dist  250.0  1.00  R
8      8/50.0  (  8/50.0)    9/50.0  (  9/50.0)  Actual dist  250.0  1.00  R
9      9/50.0  (  9/50.0)   10/50.0  ( 10/50.0)  Actual dist  250.0  1.00  R
10    10/50.0  ( 10/50.0)   11/50.0  ( 11/50.0)  Actual dist  250.0  1.00  R
11    11/50.0  ( 11/50.0)   12/50.0  ( 12/50.0)  Actual dist  250.0  1.00  R
12    12/50.0  ( 12/50.0)   13/50.0  ( 13/50.0)  Actual dist  250.0  1.00  R
13    13/50.0  ( 13/50.0)   14/50.0  ( 14/50.0)  Actual dist  250.0  1.00  R
14    14/50.0  ( 14/50.0)   15/50.0  ( 15/50.0)  Actual dist  250.0  1.00  R
15    15/50.0  ( 15/50.0)   16/50.0  ( 16/50.0)  Actual dist  250.0  1.00  R
16    16/50.0  ( 16/50.0)   17/50.0  ( 17/50.0)  Actual dist  250.0  1.00  R
17    17/50.0  ( 17/50.0)   18/50.0  ( 18/50.0)  Actual dist  250.0  1.00  R
18    18/50.0  ( 18/50.0)   19/50.0  ( 19/50.0)  Actual dist  250.0  1.00  R
19    19/50.0  ( 19/50.0)   20/50.0  ( 20/50.0)  Actual dist  250.0  1.00  R
20    20/50.0  ( 20/50.0)   21/50.0  ( 21/50.0)  Actual dist  250.0  1.00  R
21    21/50.0  ( 21/50.0)   22/50.0  ( 22/50.0)  Actual dist  250.0  1.00  R
22    22/50.0  ( 22/50.0)   23/50.0  ( 23/50.0)  Actual dist  250.0  1.00  R
23    23/50.0  ( 23/50.0)   24/50.0  ( 24/50.0)  Actual dist  250.0  1.00  R
24    24/50.0  ( 24/50.0)   25/50.0  ( 25/50.0)  Actual dist  250.0  1.00  R
25    25/50.0  ( 25/50.0)   26/50.0  ( 26/50.0)  Actual dist  250.0  1.00  R
26    26/50.0  ( 26/50.0)   27/50.0  ( 27/50.0)  Actual dist  250.0  1.00  R
27    27/50.0  ( 27/50.0)   28/50.0  ( 28/50.0)  Actual dist  250.0  1.00  R
28    28/50.0  ( 28/50.0)   29/50.0  ( 29/50.0)  Actual dist  250.0  1.00  R
29    29/50.0  ( 29/50.0)   30/50.0  ( 30/50.0)  Actual dist  250.0  1.00  R
30    30/50.0  ( 30/50.0)   31/50.0  ( 31/50.0)  Actual dist  250.0  1.00  R
31    31/50.0  ( 31/50.0)   32/50.0  ( 32/50.0)  Actual dist  250.0  1.00  R
32    32/50.0  ( 32/50.0)   33/50.0  ( 33/50.0)  Actual dist  250.0  1.00  R
33    33/50.0  ( 33/50.0)   34/50.0  ( 34/50.0)  Actual dist  250.0  1.00  R
34    34/50.0  ( 34/50.0)   35/50.0  ( 35/50.0)  Actual dist  250.0  1.00  R
35    35/50.0  ( 35/50.0)   36/50.0  ( 36/50.0)  Actual dist  250.0  1.00  R
36    36/50.0  ( 36/50.0)   37/50.0  ( 37/50.0)  Actual dist  250.0  1.00  R
37    37/50.0  ( 37/50.0)   38/50.0  ( 38/50.0)  Actual dist  250.0  1.00  R
38    38/50.0  ( 38/50.0)   39/50.0  ( 39/50.0)  Actual dist  250.0  1.00  R
```

```
Line  Wire #/% From End 1   Wire #/% From End 1    Length       Z0   Vel Rev/
      Actual  (Specified)   Actual  (Specified)                ohms Fact Norm
39    39/50.0  ( 39/50.0)   40/50.0  ( 40/50.0)  Actual dist  250.0  1.00  R
40    40/50.0  ( 40/50.0)   41/50.0  ( 41/50.0)  Actual dist  250.0  1.00  R
41    41/50.0  ( 41/50.0)   42/50.0  ( 42/50.0)  Actual dist  250.0  1.00  R

Ground type is Free Space
```

```
164' 26-Element 3-30 MHz LPDA, modified                           Frequency = 3  MHz.

Wire Loss: Aluminum  Resistivity = 4E08 ohmm, Rel. Perm. = 1

               WIRES
Wire Conn.  End 1 (x,y,z : in)  Conn.  End 2 (x,y,z : in)   Dia(in) Segs
1          980.00,24.000,  0.000       980.000,24.000,  0.000 6.50E+00 107
2          907.00,208.310,  0.000       907.000,208.310,  0.000 5.87E+00  97
3          817.49,396.320,  0.000       817.490,396.320,  0.000 5.30E+00  87
4          737.77,565.990,  0.000       737.770,565.990,  0.000 4.79E+00  79
5          655.83,719.110,  0.000       655.830,719.110,  0.000 4.32E+00  71
6          600.91,857.300,  0.000       600.910,857.300,  0.000 3.90E+00  65
7          542.31,982.020,  0.000       542.310,982.020,  0.000 3.52E+00  57
8          489.43,1094.58,  0.000       489.430,1094.58,  0.000 3.18E+00  53
9          441.71,1196.16,  0.000       441.710,1196.16,  0.000 2.87E+00  47
10         398.64,1287.84,  0.000       398.640,1287.84,  0.000 2.59E+00  43
11         359.76,1370.57,  0.000       359.760,1370.57,  0.000 2.34E+00  39
12         324.68,1445.24,  0.000       324.680,1445.24,  0.000 2.12E+00  35
13         293.02,1512.63,  0.000       293.020,1512.63,  0.000 1.91E+00  31
14         264.45,1573.45,  0.000       264.450,1573.45,  0.000 1.73E+00  29
15         238.66,1628.34,  0.000       238.660,1628.34,  0.000 1.56E+00  25
16         215.39,1677.87,  0.000       215.390,1677.87,  0.000 1.41E+00  23
17         194.39,1722.58,  0.000       194.390,1722.58,  0.000 1.27E+00  21
18         175.43,1762.92,  0.000       175.430,1762.92,  0.000 1.15E+00  19
19         158.33,1799.33,  0.000       158.330,1799.33,  0.000 1.04E+00  17
20         142.89,1832.19,  0.000       142.890,1832.19,  0.000 1.15E+00  15
21         128.96,1861.85,  0.000       128.960,1861.85,  0.000 1.04E+00  15
22         116.38,1888.62,  0.000       116.380,1888.62,  0.000 9.40E01  13
23         105.03,1912.77,  0.000       105.030,1912.77,  0.000 8.40E01  11
24         94.790,1934.57,  0.000        94.790,1934.57,  0.000 7.60E01  11
25         85.550,1954.24,  0.000        85.550,1954.24,  0.000 6.90E01   9
26         77.210,1972.00,  0.000        77.210,1972.00,  0.000 5.00E01   9

               SOURCES
Source    Wire      Wire #/Pct From End 1    Ampl.(V, A)  Phase(Deg.)  Type
          Seg.     Actual      (Specified)
1           5    26 / 50.00   ( 26 / 50.00)      1.000       0.000       V

                 TRANSMISSION LINES
Line  Wire #/% From End 1   Wire #/% From End 1    Length       Z0   Vel Rev/
      Actual  (Specified)   Actual  (Specified)                ohms Fact Norm
1      1/50.0  (  1/50.0)    2/50.0  (  2/50.0)  Actual dist  150.0  1.00  R
2      2/50.0  (  2/50.0)    3/50.0  (  3/50.0)  Actual dist  150.0  1.00  R
3      3/50.0  (  3/50.0)    4/50.0  (  4/50.0)  Actual dist  150.0  1.00  R
4      4/50.0  (  4/50.0)    5/50.0  (  5/50.0)  Actual dist  150.0  1.00  R
5      5/50.0  (  5/50.0)    6/50.0  (  6/50.0)  Actual dist  150.0  1.00  R
6      6/50.0  (  6/50.0)    7/50.0  (  7/50.0)  Actual dist  150.0  1.00  R
7      7/50.0  (  7/50.0)    8/50.0  (  8/50.0)  Actual dist  150.0  1.00  R
8      8/50.0  (  8/50.0)    9/50.0  (  9/50.0)  Actual dist  150.0  1.00  R
9      9/50.0  (  9/50.0)   10/50.0  ( 10/50.0)  Actual dist  150.0  1.00  R
10    10/50.0  ( 10/50.0)   11/50.0  ( 11/50.0)  Actual dist  150.0  1.00  R
11    11/50.0  ( 11/50.0)   12/50.0  ( 12/50.0)  Actual dist  150.0  1.00  R
12    12/50.0  ( 12/50.0)   13/50.0  ( 13/50.0)  Actual dist  150.0  1.00  R
13    13/50.0  ( 13/50.0)   14/50.0  ( 14/50.0)  Actual dist  150.0  1.00  R
14    14/50.0  ( 14/50.0)   15/50.0  ( 15/50.0)  Actual dist  150.0  1.00  R
15    15/50.0  ( 15/50.0)   16/50.0  ( 16/50.0)  Actual dist  150.0  1.00  R
16    16/50.0  ( 16/50.0)   17/50.0  ( 17/50.0)  Actual dist  150.0  1.00  R
17    17/50.0  ( 17/50.0)   18/50.0  ( 18/50.0)  Actual dist  150.0  1.00  R
18    18/50.0  ( 18/50.0)   19/50.0  ( 19/50.0)  Actual dist  150.0  1.00  R
19    19/50.0  ( 19/50.0)   20/50.0  ( 20/50.0)  Actual dist  150.0  1.00  R
20    20/50.0  ( 20/50.0)   21/50.0  ( 21/50.0)  Actual dist  150.0  1.00  R
21    21/50.0  ( 21/50.0)   22/50.0  ( 22/50.0)  Actual dist  150.0  1.00  R
22    22/50.0  ( 22/50.0)   23/50.0  ( 23/50.0)  Actual dist  150.0  1.00  R
23    23/50.0  ( 23/50.0)   24/50.0  ( 24/50.0)  Actual dist  150.0  1.00  R
24    24/50.0  ( 24/50.0)   25/50.0  ( 25/50.0)  Actual dist  150.0  1.00  R
25    25/50.0  ( 25/50.0)   26/50.0  ( 26/50.0)  Actual dist  150.0  1.00  R
Ground type is Free Space
```

A Simple UHF Remote-Control System: Pt 1

Here's a neat trick: wireless keying for your rig. It's possible in very little space, with the help of some SMT ICs designed for home security.

By Sam Ulbing, N4UAU

Through the advances of electronic technology, even a "dummy" like me can build a UHF transceiver. The projects I'll describe in this series employ modern UHF technology. They make HF communications easier and more fun, whether your interests are in digital or SSB communications. I also offer some ideas for further exploration.

I hear many hams saying things like: "Nowadays, technology is so complex that it is much too hard for the amateur to build things. You need to be a radio design genius." This really bothers me, because I know it is easier now than it ever was! I recall sweating through the equations for Chebyshev and Butterworth filters in college many years ago. Yet 30 years later, with only a vestige of my math skills left, I was able to build a excellent audio filter, a switch-mode voltage booster and other sophisticated projects.[1] As recently as 1987, Linear Technologies had commented in an applications note: "… switching regulators are also one of the most difficult linear circuits to design." By the mid 90s, both of these projects had become simple to design because modern technology has made integration of complete circuit functions possible.

I think it is time to demonstrate that radios, too, are easier to design than they used to be. In particular, I wanted to build a UHF or VHF radio because it is obvious, at least to me, that the future holds great promise for those bands. Both homes and businesses are rapidly using wireless communication. By making use of parts already available for such applications, hams can play an active role in this exciting revolution. Moreover, we are extremely fortunate to be free of many of the restrictions that apply to unlicensed users. Let's make better use of this part of our spectrum; if we don't, we may loose it.[2]

I do not have any special radio design skills; but as you will see, I do have several other skills that are needed to make use of new technology. I hope this series will convince you that

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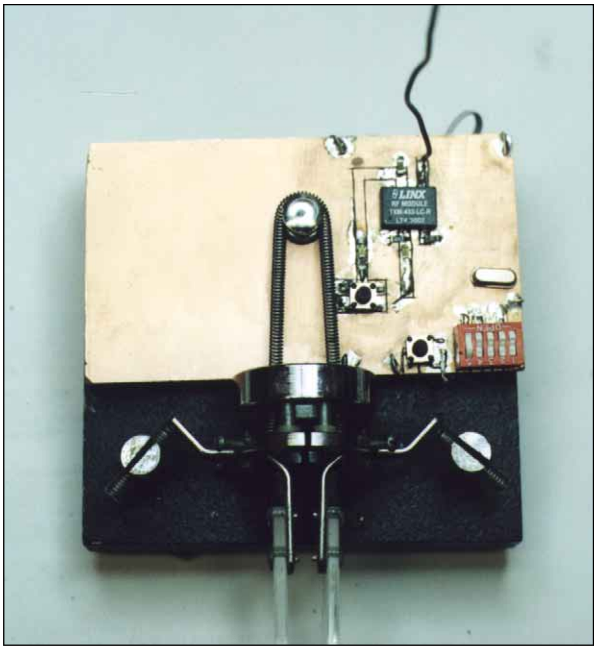
[1]Notes appear on page 39.

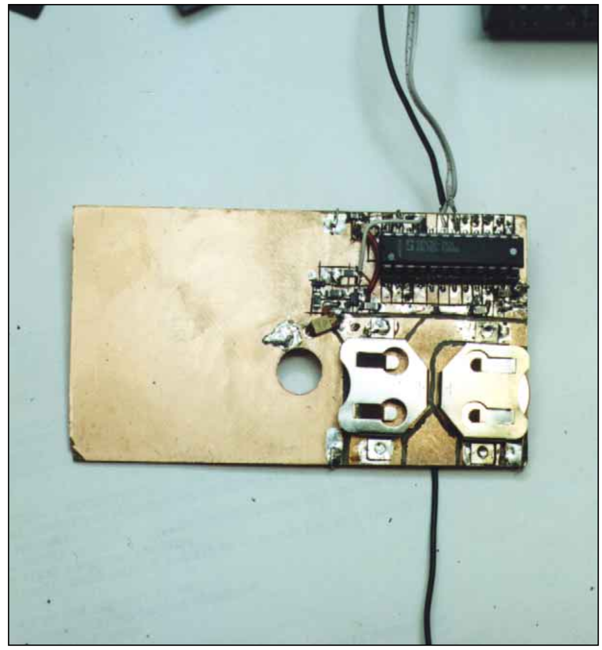
an amateur with only limited design skills can build UHF equipment. I discuss some of the new skills I found necessary. During experimentation with the projects, I uncovered many areas for further experimentation; I hope readers will be encouraged to explore and expand our knowledge base in the rapidly growing UHF area.

What's Out There

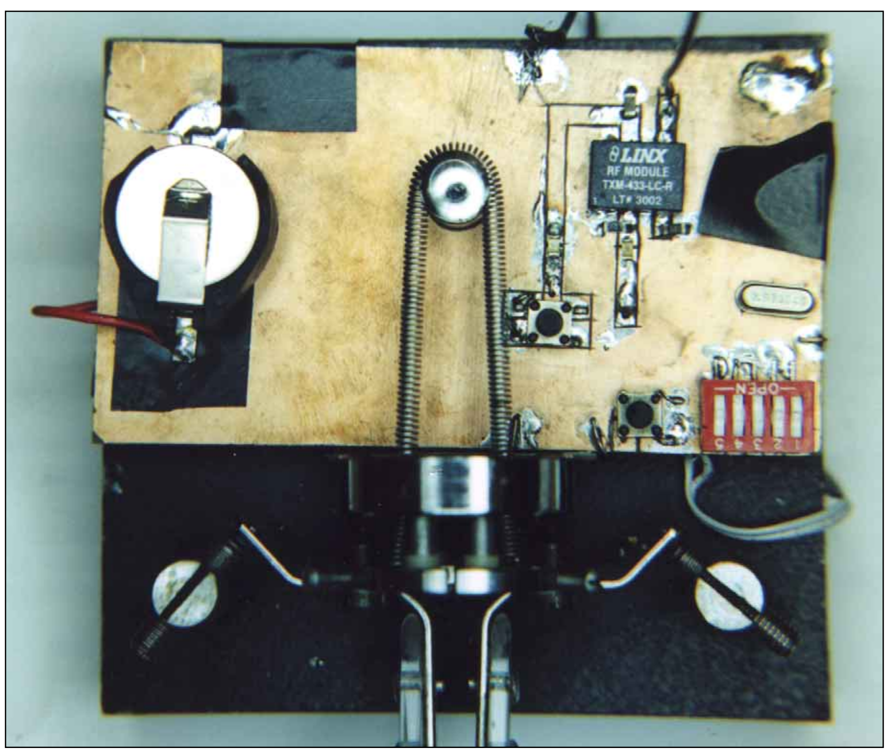
A quick search of the Web reveals many modules that make UHF design relatively simple, such as Maxim's MAX2402 spread-spectrum transmitter IC, MAX2430 high-gain RF amplifier, and MAX242X image-rejecting transceiver. I needed something even easier, though. Within the past year, a number of companies have introduced parts clearly designed for



(A)



(B)



(C)

Fig 1—The Garage Door transmitter circuit on my paddle. (A) shows the transmitter and switches. (B) shows the microcontroller, batteries and power-control circuit. All of the circuitry fits on only half of the board and that might be the best approach for many paddles. My paddle has a metal base, and it is necessary to insulate the parts from it. (C) shows the batteries moved to the top of the board, the method I prefer for my paddle.

folks like me: complete radios on a chip. I built my projects using the Linx Technology products,[3] but a number of other units are available with similar characteristics.[4]

Linx makes UHF transmitter and receiver modules that are intended for unlicensed applications such as garage-door openers, keyless entry, security alarms, lighting control and so forth. Typical circuits shown in their data-sheets use digital encoder and decoder chips for those purposes. As it happens, Linx make modules for three frequencies: 315, 418, and 433.92 MHz. Since 433.92 MHz is one of many frequencies we hams share with other services, I wondered: Why not use 433.92 MHz modules for a remote CW keyer?

I was told by the manufacturer that 433.92 MHz was intended mainly for the European market because of the possibility of interference from other sources in the US (presumably hams). It was a gamble, since the module frequencies are fixed; if I had interference, my project would not work. Looking at the band plans, I saw that 433.92 MHz was not designated for any special purpose. I took the gamble, and I am glad I did. As I started working

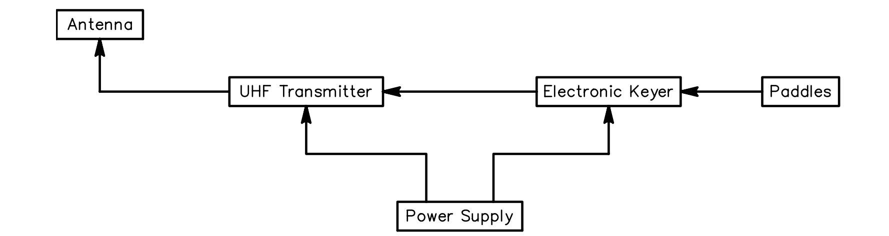


Fig 2—A block diagram of the remote keyer.

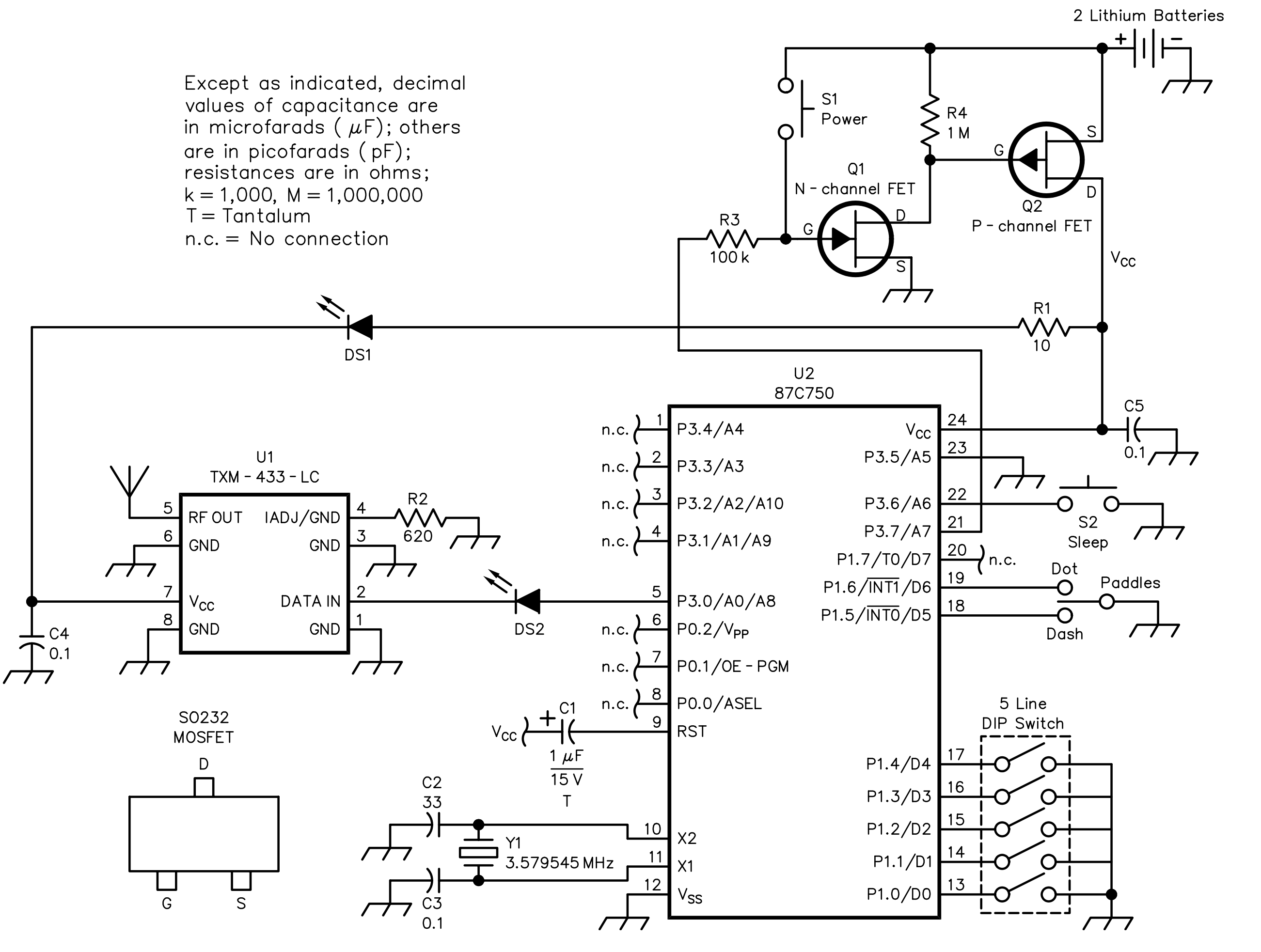


Fig 3—A schematic of the remote keyer. Unless otherwise specified, use 5%-tolerance resistors. Most resistors and capacitors are surface-mount components from the author's "junk" box. Leaded components may be used with dead-bug construction methods. Battery holders—Keystone #3002

C1—1 µF, tantalum
Q1—N-channel MOSFET suitable for logic circuits
Q2—P-channel MOSFET suitable for logic circuits
U1—TXM-433-LC; see Note 3.
U2—87C750 programmed microcontroller in 24-pin DIP. See Note 3 and the sidebar "Programming an 87C750."

with them, I discovered a myriad of other useful and interesting applications. I suspect I may be spending the next several years making variations.[5]

Project 1: How To Surf The Web While Doing CW—A Remote Keyer

After 10 years of CW experience, I have finally gained the ability to copy in my head. Now I sit at my computer and listen to the CW net on the HF rig across the room, but whenever I hear "N4UAU kn" I have to leap up and run across the room to respond. Granted, this is great exercise, but I wanted something better. Fig 1 shows my remote keyer. With it, I sit at the computer and when I hear my call, I can respond without getting up.

Designing a modern wireless system is different than it was in "the good old days." To see what I mean, look at Fig 2, a block diagram of my remote keyer: It consists of five sections. Bencher was kind enough to design and build the "paddles" module and Linx Technologies provided the UHF transmitter solution. That left me only three sections to design: antenna, power supply and keyer, none of which requires any special radio design skills.

Quite a few keyer designs appear in the literature and all of the modern ones, like Uncle Albert's Unique keyer, are based on microprocessors.[6] To build a small and efficient keyer, I needed the facilities to both write code for, and physically program a microprocessor.

There are many microprocessors to choose from, and because I have the skills and equipment to build a keyer using a 87C750, my keyer is based on that microprocessor.[7] The code I wrote is a variant on the code I wrote for the successful Uncle Albert's keyer. (See sidebar "Programming an 87C750.")

Design of the Keyer

Fig 3 is the schematic for my remote keyer. U1 is the Linx transmitter module; U2 is a programmed 87C750 microprocessor. The DIP switch connected to

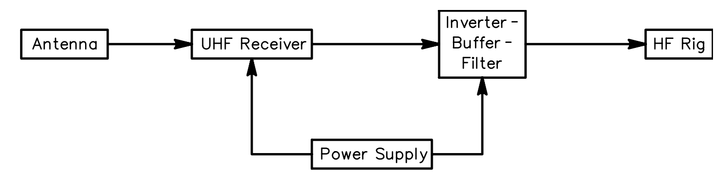


Fig 4—Receiver block diagram.

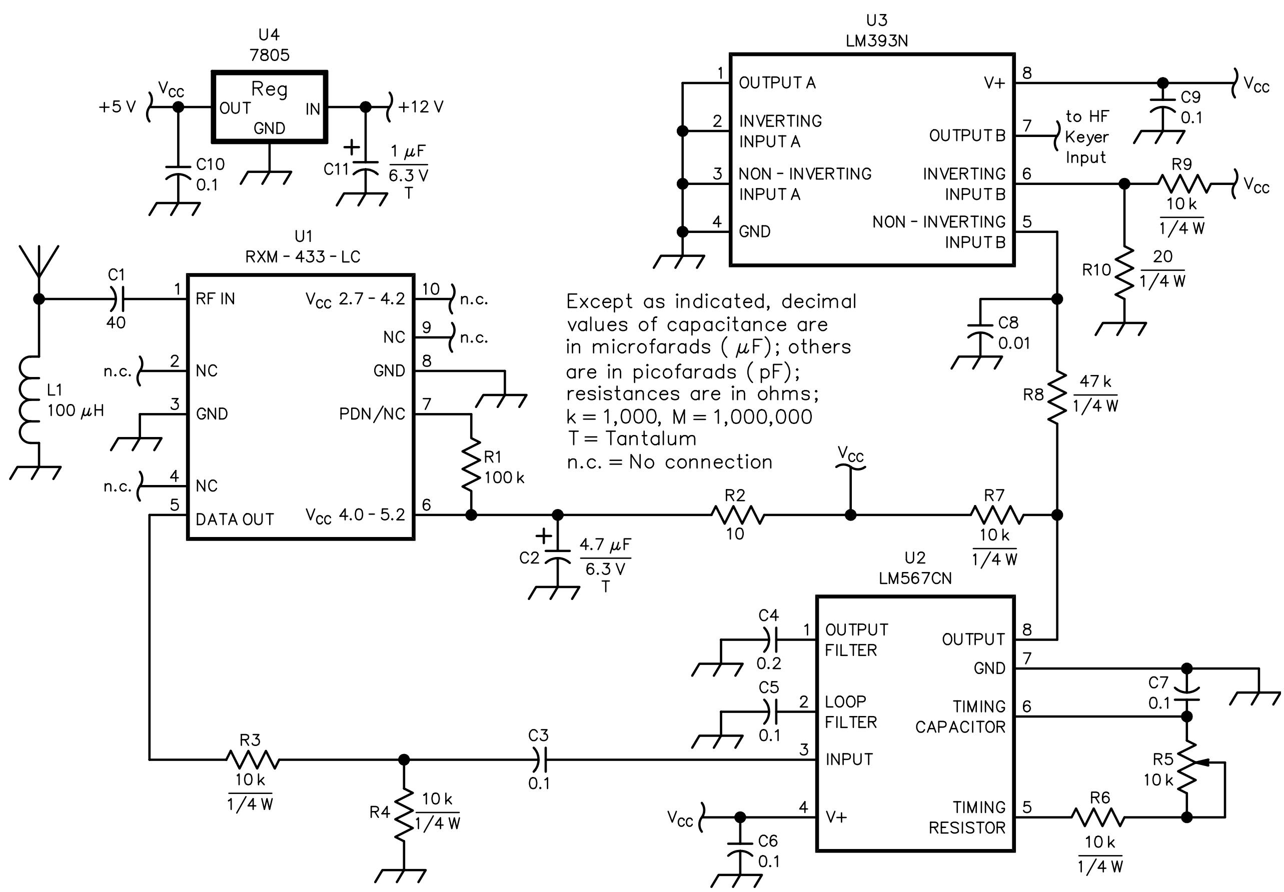


Fig 5—A schematic of the receiver decoder for the remote keyer. Unless otherwise specified, use 5%-tolerance resistors. Most resistors and capacitors are surface-mount components from the author's "junk" box. Leaded components may be used with dead-bug construction methods.

U1—Linx RXM-433-LC; see Note 3
U2—LM567CN
U3—LM393N
U4—LM7805 5 V regulator

pins 13-17 of U2 can set 32 different speeds, from 13 to 44 WPM in 1-WPM steps. The keyer is iambic: If you hold the dot and dash paddles at the same time, you get alternating dots and dashes. Weighting is "standard." I considered implementing different weighting options, but ultimately chose not to, as I find "weighted" code more difficult to copy and it makes the programming more complex.

To minimize power consumption, which is a direct function of the clock speed, I used a color-burst crystal, Y1, running at 3.579 MHz and yielding a cycle rate of about 300 kHz.[8] This microprocessor has a power-reducing idle mode and the program code puts the keyer into this mode whenever it is not actually sending code. An interrupt is needed to "wake up the micro" from this idle mode. That happens when a dot or dash is sent because the paddles are connected to pins 18 and 19, which are interrupt ports.

CW output is taken from pin 5 of U2. The Linx transmitter module uses logic high to turn on the transmitter. Like the micro, it is in an idle mode when not actually transmitting. Turn-on and turn-off times for the transmitter are around 10 µs. Compared with the length of a dot at 45 WPM (about 27 ms), this time is insignificant. Because the transmitter is only on when it is actually sending, it is very power-efficient. The typical current draw of the entire remote keyer is about 5 mA when transmitting and around 1 mA when idle. If you listen half the time and talk half the time, the average current would be only 3 mA.

Since my keyer is battery operated, power consumption is an important consideration; so I included a shut-down feature in the program code. After being idle for about 14 minutes, pin 21 is set low and the micro goes into a "sleep" mode. Sleep mode reduces the current to about 200 µA. To reduce it even further, I designed the power-supply control circuit to reduce it to zero. You can manually force the circuit into sleep mode by pressing S2.

Designing The Power Supply Module

The following factors dictated my choice of power source: The 87C750 needs 5 V with an absolute maximum of 6.5 V, while the transmitter module can operate from 2.7 to 5.2 V with 6 V as its absolute maximum. With the low current draw, I knew I could use a small battery. I happened to have some size-2032 lithium batteries left over from an earlier project, so I wanted to use one of them. While the transmitter would work with just one battery, the microprocessor would not. (There are micros that run at 3 V; but to use one, I would have to learn the language, get an assembler and buy or make the hardware to program it. Doing that would a costly and time-consuming project in itself.[9]) I used two series-connected batteries to give 6 V.

The microprocessor connects directly to the batteries; but since the transmitter has a 6-V maximum, I decided to be on the safe side[10] and used an LED to drop the voltage to the transmitter module.[11] As a side benefit, the LED flashes the code as I send it, which lets me know the circuit is turned on and working. DS2 ensures that the logic high input from U2 does not exceed the power-supply voltage of the transmitter chip. For operation above 3 V, the transmitter module requires that R2 be used between pin 4 and ground.[12]

Q1 and Q2 make a "soft" on/off switch that allows U2 to switch off the power completely when it enters the sleep mode. S1 is the power switch; when it is pressed, Q1 conducts, turning on Q2 and applying power to the circuit. As the microprocessor boots up, it sets pin 21 high, which keeps Q1 on even though S1 is a momentary switch. When pin 21 goes to 0 V—either because the microprocessor has entered the sleep mode or because S2 was pressed—Q1 is off, forcing the base of Q2 high, switching off power to the microprocessor. To minimize current draw, Q1 and Q2 are MOSFETs rather than transistors.

Transmitter Antenna Module

The final design area is the antenna. The transmitter is intended for a 50 Ω load. An obvious choice is a λ/4 vertical, at about 35 Ω.[13] A 6.5-inch piece solid wire comprises the antenna. This monopole needs a good ground plane for best operation. I used two-sided copper-clad PC board for the project, made it as large as I could (while keeping it small enough to fit on the Bencher paddle) and connected the top and bottom ground planes with jumper wires.

Designing a Receiver

Fig 4 is the block diagram for the receiver and Fig 5 is its schematic. I originally thought that the power supply would be the main design focus for the receiver, but I was wrong. According to the datasheet, power may be applied to two places, depending on the supply voltage. Pin 10 is to be used for V_{cc} from 2.7 to 4.2 V (4.2 V absolute maximum), and pin 6 for a voltage between 4.0 and 5.2 V (5.5 V maximum). The current draw, 7 mA active and 370 µA idle, is somewhat greater than for the transmitter. Since size is not important, four AA batteries connected to pin 6 make a good power source with a lot of capacity. The maximum voltage could reach 6.4 V, however, so either a series LED or two diodes are necessary to keep the voltage in a safe range. On the other hand, four NiCd batteries would never exceed 6.0 V and a single

Programming an 87C750

I know of three ways to program a '750. Two are expensive and one requires moderate effort on the part of the builder. If you choose to use the do-it-yourself method of #1 below, you might be pleased to know that the computer I built in 1993 using extremely "ugly construction" is still functioning flawlessly seven years later. That is more than I can say for a commercial 80C51 prototyping board I bought that same year.

1. You can build a plug-in attachment for the programmer, as in my articles "An 8085-Based Computer System," (*QEX*, Nov 1993, pp 3-8) and "A Programmer for 87C51 Microcontrollers," (*QEX*, Dec 1993, pp 10-12). The plug-in would use code very similar to that for the 87C51, but the wiring on the plug-in board would have to be done for a 28-pin rather than a 40-pin IC. If you want to use a larger 87C51 IC, Philips now makes a 3-V version that offers the advantage of complete 3-V operation of the circuit. I have used my programmer with no changes to program one of these, although its timing algorithm is slightly different.
2. I used a DS-750 development tool from Ceibo-Philips to program the '750 for this project. This tool will program '750s, '751s and '752s. It comes with a simulator and assembler software. Contact Ceibo at 800-833-484 or www.ceibo.com. A development board is now $390. When I got it several years ago, it was at $99 because the '750 was a new IC—it has obviously been a success!
3. Several "universal" programmers may program '750s.

diode would suffice. Alkaline batteries have more capacity than NiCds (2000 mAh versus 600 mAh), but the voltage drop of the LED or two diodes would diminish it somewhat.

I realized I might also use a 12-V power supply and 5-V regulator because I was using the receiver in a fixed application. This is the way I chose to go: Pin 7 of U1 is a shutdown pin that might be handy for portable use. I did not need it and—in this RF environment—a pull-up resistor, R1, ensures the receiver will not accidentally switch off.[14]

The output of the Linx receiver, pin 5 of U1, goes from logic low (<0.2 V) to logic high (>3.6 V) when a signal is received. Because the output cannot source a lot of current, it needs to be buffered with a MOSFET transistor or other high-impedance device, such as U2. The turn-on and turn-off times of the receiver are about 40 µs, longer than those of transmitter, but still much less than that of the fastest dot. Further, since turn-on time is nearly equal to turn-off time, the receiver does not change the signal greatly but only delays it.

Gremlins Alter the Challenges Of Design

The circuit worked well as described most of the time; but as I used it more, I uncovered a couple of problems that changed the main design focus of the project. First, when I tested my receiver without keying the HF rig, it worked great; but when I actually transmitted, the useful range of the remote keyer decreased dramatically. A little testing revealed that the problem was desensitization of the UHF receiver front end. A simple high-pass filter, L1 and C1, solved this problem.[15] I used surface-mount (SM) components here to keep the filter quite close to the antenna input pin because of the ultra-high frequencies.

The second challenge arose because I lacked knowledge of how the receiver output works. I found that at odd times the receiver output would go high and stay high for no apparent reason, thus keying the HF rig. The datasheet indicated that the output would go high in the presence of a signal and be low otherwise, but it appeared there was no signal when the output went high. A design engineer at Linx clued me in: The receiver output is a “data slicer,” commonly used in digital reception when the recovered ac signal is quite weak. To improve reception, the signal is averaged over time and sent to one input of a comparator. The other comparator input receives the signal without averaging. Using the comparator in this way helps remove the effects of signal variations due to noise and interference. When I was not sending CW, the average and actual signals were nearly identical, causing the comparator to change state on noise bursts. For garage-door openers, this is not a problem because they incorporate a decoder that will not respond to random signals. I, on the other hand, was connecting the receiver output directly to the HF rig.

I needed to encode and decode the CW signal. Fig 6 shows the difference between “modulated” CW and “real” CW. The receiver/decoder circuit is shown in Fig 5. By sending modulated CW[16] and using the LM567 to decode the modulation, the HF rig is not keyed by random noise. R3 and R4 reduce the input level of the signal to the decoder. R5, R6, and C7 set the response frequency. R7 is a pull-up resistor for the output of the decoder, which otherwise would float when no signal is detected. As the input signal nears the frequency of the decoder, the output of U1 will tend to oscillate between high and low. U3, a comparator, R8 and C8 sharpen the edges of the bandpass area. The trigger level of the comparator is set (by R9 and R10) very close to 0 V, so that the comparator output only goes low when the output of U1 is actually locked onto a signal. Near-frequency oscillations will not trigger it. Because the tone decoder needs several cycles to establish lock, selection of the components is important. Using 100 kΩ for R8 gave a clipped effect to faster dashes, but 47 kΩ reduced clipping significantly. C4 and C5 also have an effect on lock time.

Building the Remote Keyer Project

With smaller, SM parts, building techniques needed today are different from “the old days.” Fortunately, I have developed the skills to work with SM parts.[17] To make this keyer as small as possible, I put parts on both sides of the board, a trick that is impractical for through-hole parts. The keyer shown in Fig 1 is mounted on a Bencher paddle and has a “large” PC board. I left it this large to provide a large ground plane for the antenna, even though all the parts are on only half of it. For use with other paddles or mounting in a stand-alone case, the unused part of the board may be cut off.

I made my own board using the method I described in the article of Note 17. Soldering the Linx modules is bit of a challenge. You can see from Fig 7 that these ICs were not designed with ease of hand assembly as a high priority.[18] The manufacturer recommends the following: extend the pads

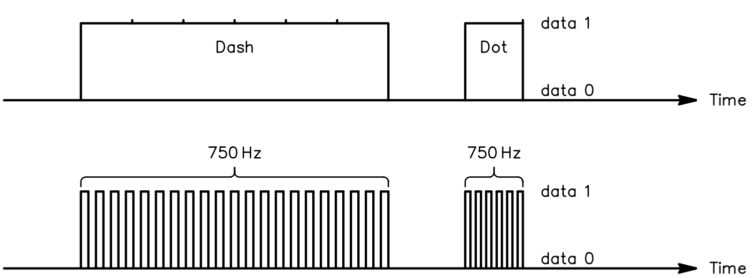


Fig 6—(A) shows the letter “N” as pure CW keying; (B) shows “N” as modulated CW for a VCO transmitter.

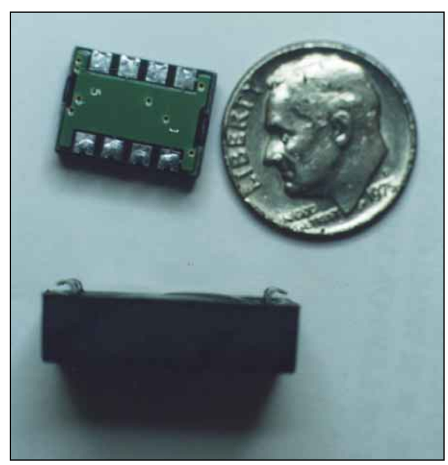


Fig 7—The LC transmitter module (top) and receiver module (bottom) showing the user-*unfriendly* connections. The transmitter pads are flush with the bottom and require “wicking” of the solder. The pins on the receiver are bent to be underneath the package. Although there is some clearance, wicking will be the main way to attach it. Despite these difficulties, it is possible to solder them with a little care and patience.

beyond the IC; use a fine-tipped soldering iron placed on the pad and also in contact with the pin or side indentations on the transmitter IC and introduce the solder at the module's edge where it will wick underneath the module. (I used my flux pen and tinned the pads and pins, as well.) Tack solder one corner and make sure the part is aligned; then work around the remaining attachment points. Be careful not to heat the part too much. Per the datasheet: Maximum solder temperature is 225°C (437°F) for 10 seconds. Recommended solder has a melting temperature is 180°C (356°F) and Kester 24-6337-6403 with water-soluble flux is suggested.[19]

I used 0.020-inch solder (Kester 24-6337-0027), set my iron temperature to 625°F and made very sure not to hold it to the pad for more than a couple of seconds so that the IC would not exceed 437°F. I allowed the IC time to cool before soldering each pin. Once you have soldered the transmitter IC to the PC board, it will be nearly impossible to remove it. For prototyping, I mounted the modules on small pieces of PC board (carriers) with traces extending beyond each pin by about 1/4 inch. I put those carriers on a bigger board, held in place with a piece of double-stick tape or hot-melt glue and connected wires to them. The carriers are easy to remove and it is easy to solder and unsolder their connections, being careful not to overheat the IC.

Module Capabilities and Restrictions

The LC-model transmitter and receiver block diagrams are shown in Fig 8. These modules use sophisticated SAW (surface-acoustic-wave) devices that operate at a tightly controlled frequency of 433.92 ±0.05 MHz. The receiver has a band-select filter, Gilbert-cell mixer, SAW-controlled local oscillator, ceramic filter, AM detector and more. The circuits are hybrids packaged in a single-module IC.[20] While the oscillator is tightly controlled, it turns out that the receiver input filtering is as wide as a barn door.[21] It is therefore possible for both in-band and out-of-band signals to cause unwanted responses: That is why the decoder circuit is necessary.

The datasheet claims a range of up to 300-feet for these modules. RF power is specified as 0 dBm (1 mW, with $V_{cc} = 3.0$) and as much as +3.5 dBm (about 2 mW, with $V_{cc} = 5.0$). The maximum data rate is 5 kbps—6000 WPM of CW! Considering the time it has taken me to reach 30 WPM, I don't think I will be pushing the module's capacity anytime soon. Nonetheless, this high data-rate capability offers a number of intriguing opportunities for use with faster digital modes.

As with all RF devices, board layout is very important. No conductive items should be within 1/4-inch of the module's top or sides.[22] A ground plane on either or both layers of the PC board is recommended. For antenna placements greater than 1/4 inch from the module, use a 50-Ω trace or coax.[23] In a future segment, I will how to make a 50-Ω feed line right on the PC board.

Finally, the datasheet notes: "The choice of antennas is one of the most critical and often overlooked design considerations." For us hams, this is a wonderful opportunity for experimentation. At 433.92 MHz, a 1/4 wave is a mere 6.5 inches long. Some gain antennas are shown in the ARRL literature.[24] In Part 2 of this series, I will show several low-profile antennas that have negative gain but offer the benefits of small size: a big advantage for remote work and a little-explored area for amateurs to the best of my knowledge.

What Next?

Did you notice the design of this UHF transceiver project presented quite a number of challenges, but that none of them required extensive radio-design expertise? Heck, I didn't even need to know what a SAW is, other than maybe it was something used to cut wood.[25] The skills needed to build wireless communication devices have changed. Instead of knowing about SAWs, I had to be able to write a program, find power sources that were appropriate, determine suitable antennas, make a PC board and solder the modules and SM parts. Developing these new skills takes a bit of time, but once obtained, the projects an amateur can build are

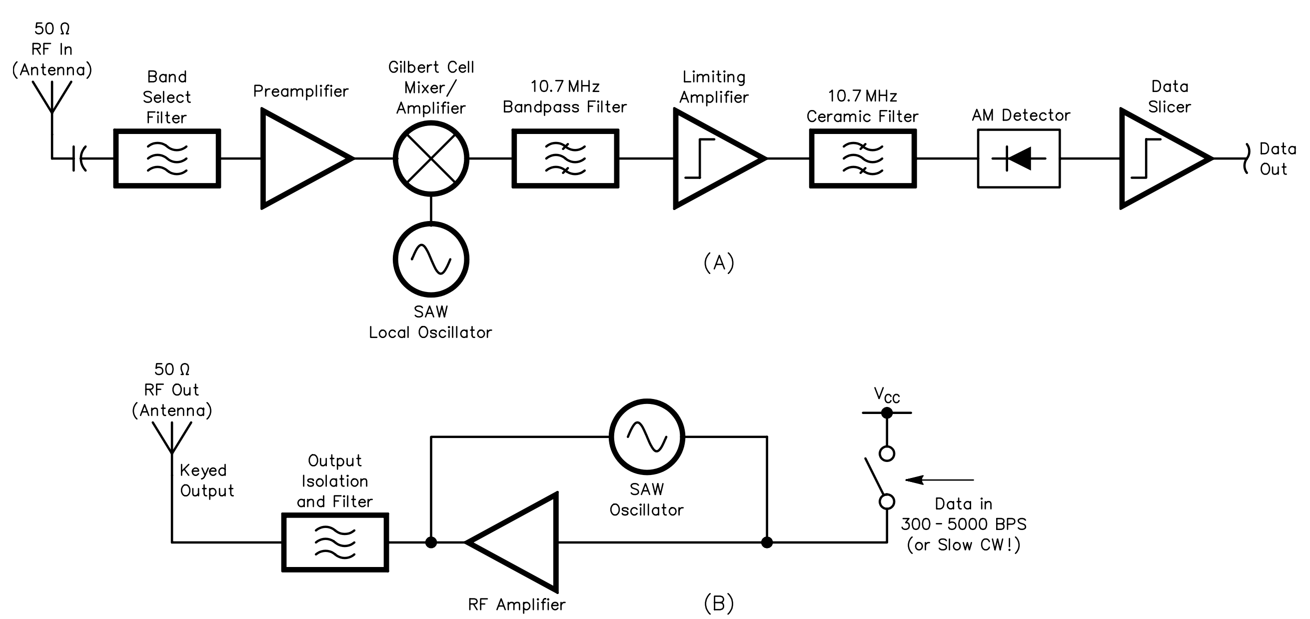


Fig 8—Block diagrams: (A) LC series receiver, (B) LC series transmitter.

far more sophisticated than those of "the good old days."

My remote keyer works great. I can now surf the Web and be active on the CW net at the same time. In fact, the project was so easy and worked so well, I wanted to do more. Next time, I will show you how I was able to have coffee with my wife and listen to the Waterway SSB net at the same time.

Notes

[1] S. Ulbing, N4UAU, "An Active Audio CW Filter You Can Build," *QST*, Oct 1992, pp 27-29; also "My All-Purpose Voltage Booster," *QST*, July 1997, pp 40-43.

[2] A personal comment: I believe it is time for Amateur Radio to petition FCC to expand their privileges to include all methods of noncommercial wireless communication. There is an enormous opportunity for learning and doing in the UHF area. Why shouldn't hams be able to build wireless home centers for music, video communications or appliance control? I think doing this would make Amateur Radio more interesting to the younger generation, thus helping revitalize our hobby. If you agree, please write to me.

[3] Linx Technology, Inc, 575 SE Ashley Pl, Grants Pass, OR 97526; tel 800-736-6677, fax 541-471-6251; **www.linxtechnologies.com**. Small quantities of Linx parts can be purchased from RF Digital Corp, 1160 N Central Ave Ste 201, Glendale, CA 91202; tel 818-500-1082, fax 818-246-9122; e-mail **info@rfdigital.com**; URL **http://www.rfdigital.com/**.

[4] RF Monolithics, Dallas Texas, **www.rfm.com**; Micrel, Inc, 1849 Fortune Dr, San Jose, CA 95131; tel 408-944-0800; **www.micrel.com** are just two companies involved in this area. Both of these sites and the Linx site have a wealth of technical information available for as free downloads.

[5] I have recently been invited to work with the Machine Intelligence Laboratory at the University of Florida to assist them apply such UHF modules to robotics. You can see what they are working on at **www.mil.ufl.edu**.

[6] S. Ulbing, "Uncle Albert's Unique Keyer," *QST*, Jan 1994, pp 42-44; also "The New and Improved Uncle Albert's Keyer," *The 1996 ARRL Handbook*, pp 22.21-22.24; "The N4UAU Super CW Station," *73 Amateur Radio Today*, June 1995, pp 10-16.

[7] While the '750 lacks some nice features like EEPROM and SM for small size, it offers benefits like a large number of ports to process external information, two interrupt ports, an internal clock, idle and sleep modes. Like all else, microprocessor selection has tradeoffs.

[8] The 87C750 clock runs at 1/12 of the crystal speed.

[9] To get a feel for the work involved in making a programmer, see the articles mentioned in the sidebar "Programming an 87C750."

[10] Two new lithium batteries could exceed 6.0 V.

[11] You could use a diode instead. The voltage at the transmitter will be greater, and the radiated power and current draw will increase as well. I found the keyer worked fine at the lower voltage.

[12] For 3-V operation, this pin can be connected directly to ground.

[13] Per RF Monolithics Application Note AN36, "Antennas For Low Power Applications," by Kent Smith, available at their Web site.

[14] It can be left floating for normal operation. Pulling the pin to ground shuts down the receiver and reduces the current draw to around 350 µA.

[15] I started with an SM inductor and capacitor that I happened to have on hand. It helped a lot, so I "designed" a two-element Butterworth filter using the tables in *The 1992 ARRL Handbook* (p 2-40). Because the UHF signal is so much higher in frequency than the HF signal, a more sophisticated filter is not needed.

[16] Pin 5 of U2 outputs modulated CW and pin 3 outputs real CW for applications that need it.

[17] S. Ulbing, "Surface Mount Technology—You Can Work With It," *QST*, April 1999, pp 33-39.

[18] The datasheet highlights this with comments like: "Since these pads are inaccessible during mounting, castellations that run up the side of the module have been provided to facilitate solder wicking" and ". . . the low profile of the module makes it difficult to contact the pins during hand assembly." As noted earlier, the skills needed to build equipment now are different from those of not so many years ago.

[19] The temperature limits are low because the IC is comprised of discrete elements soldered to a PC board. Overheating might cause the internal connections to unsolder.

[20] The LM2825 switching regulator I described in the SM article uses the same approach: Perhaps this is a trend in packaging technology.

[21] This is a quote from the technical support at Linx Technologies. Realize that the design goals for this chip were small size, low power, low cost. The receiver works well as long as you realize the design tradeoffs.

[22] This is for the receiver; for the transmitter, the distance is given as 0.15 inch.

[23] Keep in mind the high attenuation of coax at these frequencies. See *The ARRL Antenna Book* for actual figures.

[24] See *The ARRL Handbook*, *Antenna Book*, Antenna Compendiums, etc.

[25] Those interested in how SAW devices work, should look at *The ARRL Handbook* discussion of SAW filters (p 16.20) for an easy-to-understand explanation. SAW devices are basically much-improved crystals.

Sam Ulbing, N4UAU, studied electronics in the 1960s, but spent his career in the financial area. Since he retired in 1986, Sam has enjoyed exploring the opportunities offered to the amateur builder by the new ICs. He feels that electronic design for amateurs has become much easier than it once was. Sam recalls how in the '60s, he spent hours sweating over complex equations to design even simple circuits. Now although he has forgotten almost all of his math, the circuits he has built with the new electronics perform very sophisticated functions, and best of all—they work! Sam has a Web page at **http:/n4uautoo.home.sprynet.com**.

A PLL Spur Eliminator for DDS VFOs

Modern DDS systems can easily generate RF signals—perhaps too many. This setup ensures pristine VFO output.

By Rick Peterson, WA6NUT

If you've built a DDS VFO for your HF transceiver, you may be plagued by audio tones appearing at the receiver output that need to be cleaned up before the VFO is usable on the air. This PLL project, adapted from a readily available kit, can rid your DDS VFO output of those nasty spurs. It also indicates PLL lock and interrupts the VFO output if the PLL loop ever becomes unlocked while transmitting.

The popularity of PC-based transceivers, such as the Kachina 505DSP and the Ten-Tec Pegasus, has sparked new interest in using PC-controlled, direct-digital synthesizers (DDSs) to tune older rigs, such as the venerable Heath HW-101. Other DDS tuning schemes use PIC microprocessors, BASIC-stamp modules or other microcontrollers. But, however controlled, the DDS output is subject to spurs generated by mixing of the DDS signal with clock fundamentals and harmonics.[1, 2, 3]

Together with a 4.0-5.5 MHz PC-controlled DDS,[4] I use the PLL unit to replace the VFO in my Heath HW-101 transceiver. The PC parallel port and software provide several other functions:

- DDS control with frequency memories, split transmit/receive frequency operation and RIT
- CW keying from the keyboard or built-in iambic and non-iambic keyers
- received-CW decoding with text displayed on the PC monitor

All this runs in a single QBASIC program of less than 40 kB! Thus, the old Heath HW-101 transceiver now has some features in common with state-of-the-art transceivers.

The PLL unit described in this article is *not* required if your HF transceiver (such as the Drake TR-7A) was designed with a PLL following the internal 5.0-5.5 MHz analog VFO. When you replace the original analog VFO with your external DDS VFO, the internal PLL in the transceiver will clean up the DDS VFO spurs;[5] however, software controlling the DDS VFO should ensure that commanded frequency changes never exceed the PLL lock-in range.

Objectives

The PLL unit was designed with the following design goals in mind:

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wa6nut@arrl.net

[1]Notes appear on page 49.

• It should accommodate the popular VFO range of 5.0-5.5 MHz (requires DDS range of 4.5-5.0 MHz).
• It should allow out-of-band receiver operation with extended 4.5-6.0 MHz VFO range (if the receiver first-IF filter bandwidth and receiver-mixer spurs allow). This will require DDS range of 4.0-5.5 MHz.
• It should provide a lock detector to indicate PLL lock and interrupt the VFO output if the PLL loop ever becomes unlocked while transmitting.
• It should use readily available kits wherever possible to reduce construction time.
• It should use ac/dc wall adapters to simplify power-supply construction so that no 120 V ac wiring is required.

Block Diagram

The block diagram is shown in Fig 1. Readers of *The ARRL Handbook*[6] will recognize the similarity between Figs 14.51 and 14.52 and the PLL6 board. In fact, the PLL6 board is identical to the G3ROO/GM4ZNX design described on pp 14.45-14.49 of the *Handbook*, with a few improvements: R51 and D13 are replaced by a 78L05 regulator. Multiple crystal oscillators were added; the Q16/Q17 amplifier is deleted, and six bands are implemented on the PLL6 instead of twelve, as in the *Handbook* article. Readers can find a complete theory of operation and alignment procedure in the *Handbook*.

The PLL6 board is a kit manufactured by Hands Electronics[7] in the UK and distributed in the US by Kanga US.[8] Normally, the kit is sold with six "band packs:" the VCO and crystal-oscillator components for six amateur bands. The kit is normally sold as part of a six-band HF transceiver, but the manufacturer has made the kit available as a "Special PLL6" kit, with no band packs supplied, for special applications. The purchaser of a Special PLL6 kit must supply his or her own band-pack components; in this application, those are components for a 4.5-6.0 MHz VCO and 10-MHz crystal oscillator. The kit is not difficult to assemble, and is supplied with good documentation for assembly and alignment. When using the board with only one VCO, the loop compensation may be optimized for the sensitivity (K_v in Hz/V) of just the one VCO, instead of compromising the compensation for the K_vs of several different VCOs.

The lock-detector board has an amplifier to boost the PLL6 VCO signal to the level required by the HF transceiver VFO input. The lock-detector-board low-pass filter is a kit manufactured by Communication Concepts;[9] it reduces the harmonic content from the lock-detector-board amplifier. Although large—the filter is designed for use at a transmitter output—it is reasonably priced and works well.

The PLL tuning-diode voltage is applied to a window comparator on the lock-detector board. When the tuning voltage falls outside the comparator voltage window, the green **PLL LOCK** LED is extinguished, and the relay contacts open (only in transmit mode), interrupting the VFO signal to the transceiver. The logic also includes circuitry to briefly interrupt the VFO output immediately following a transmit-to-receive or receive-to-transmit transition. Thus, during split-frequency operation, the VFO output is inhibited while the VFO is changing frequencies.

PLL6 Board Modifications

Fig 2 is the schematic of the PLL6 board modifications to the VCO and crystal oscillator band-pack circuits. Sources for the required parts are shown in the figure captions; e-mail and Web addresses for each source are given in the Notes[10,11,12,13,14] at the end of this article. In the event that a component is no longer available from a listed source, try the Part Miner Web

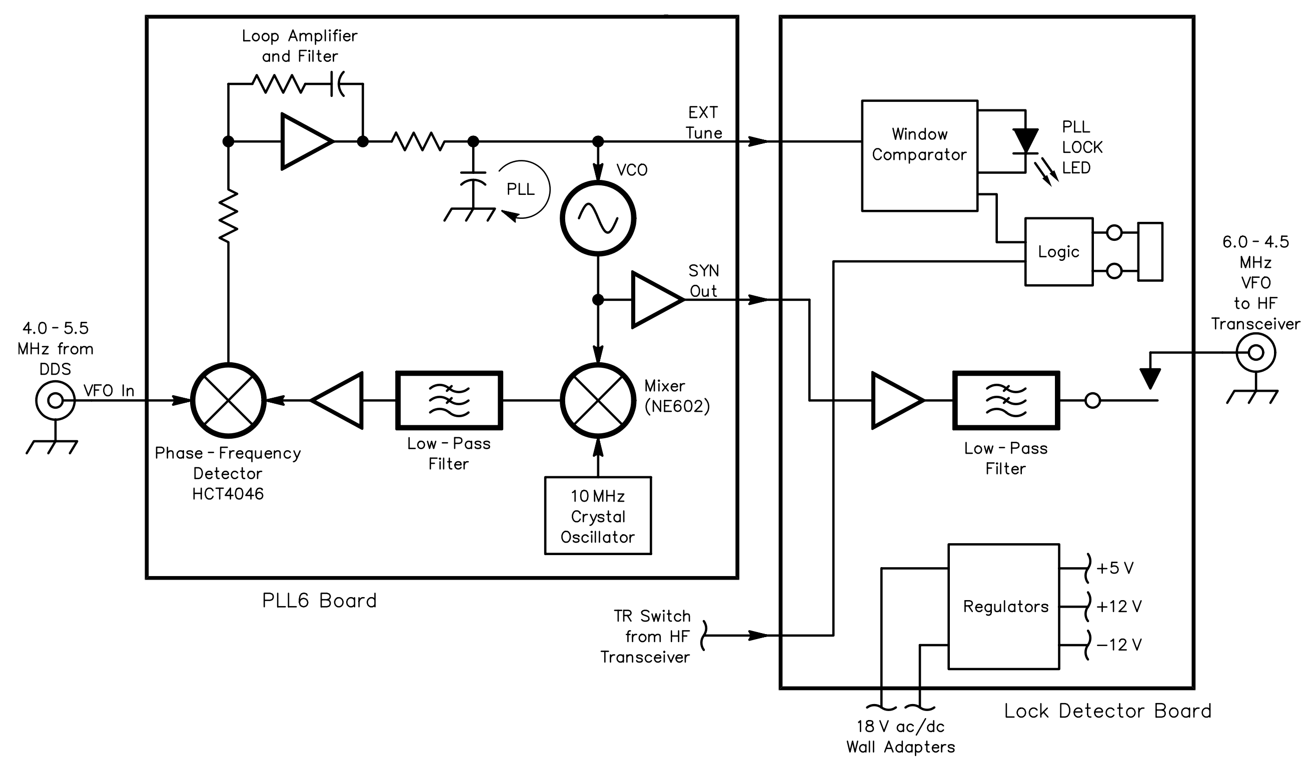


Fig 1—PLL unit block diagram.

site.[15] It's not only a great way to locate parts, but datasheets are available at the click of the mouse.

To mount D5A1-D5A4, I soldered three solid hookup-wire "posts" at the D5A location. I then soldered the four MVAM109 tuning diodes to the posts: two on one side, and two on the other, with all four soldered to the middle post. Be careful to observe the correct diode polarity.

Note that R17A is a "test-select" part. If you have an oscilloscope, check the **SYN OUT** waveform after assembling and aligning the PLL6 board. For my board, substituting a 27-kΩ resistor eliminated distortion of the waveform, resulting in a near-perfect sinusoid. Some experimentation may be required to find the optimum R17A value for your PLL6 board. If you're not able to observe the waveform, leave the R17A value at 100 kΩ.

Without a trimmer capacitor at C27A, the crystal oscillator used on the PLL6 board oscillates at a frequency slightly above the nominal crystal frequency—15 kHz higher in my unit. This was not a problem for my application because the PC-controlled DDS software provides for calibration to WWV for both "scale-factor" error (DDS clock-frequency error) and "offset" error (combined frequency errors in the PLL6 and HF transceiver's crystal oscillators). If hardware calibration of the crystal-oscillator frequency is preferred for your application, a 30-pF trimmer capacitor can be added at C27A on the PLL6 board.

C20A is also a test-select part. You may need to increase the value of C20A to mitigate close-in PLL phase-noise effects at low reverse-bias levels. These effects appear as raspy-sounding received CW signals (and heterodynes) heard interspersed among normal signals, once the PLL unit is installed between the transceiver and DDS unit.

Fig 3 is a schematic for the PLL6 board modifications to the loop amplifier and filter circuits. The new component values provide compensation for the VCO sensitivity, K_v, for the VCO redesigned with MVAM109 tuning diodes.

I obtained all the common-value, 1/4-W, ±5% resistors for this project from a Radio Shack carbon-film assortment (RS #271-312). Other less-common values were obtained from a Digi-Key carbon-film assortment (Digi-Key #RS225-ND). The assortments are

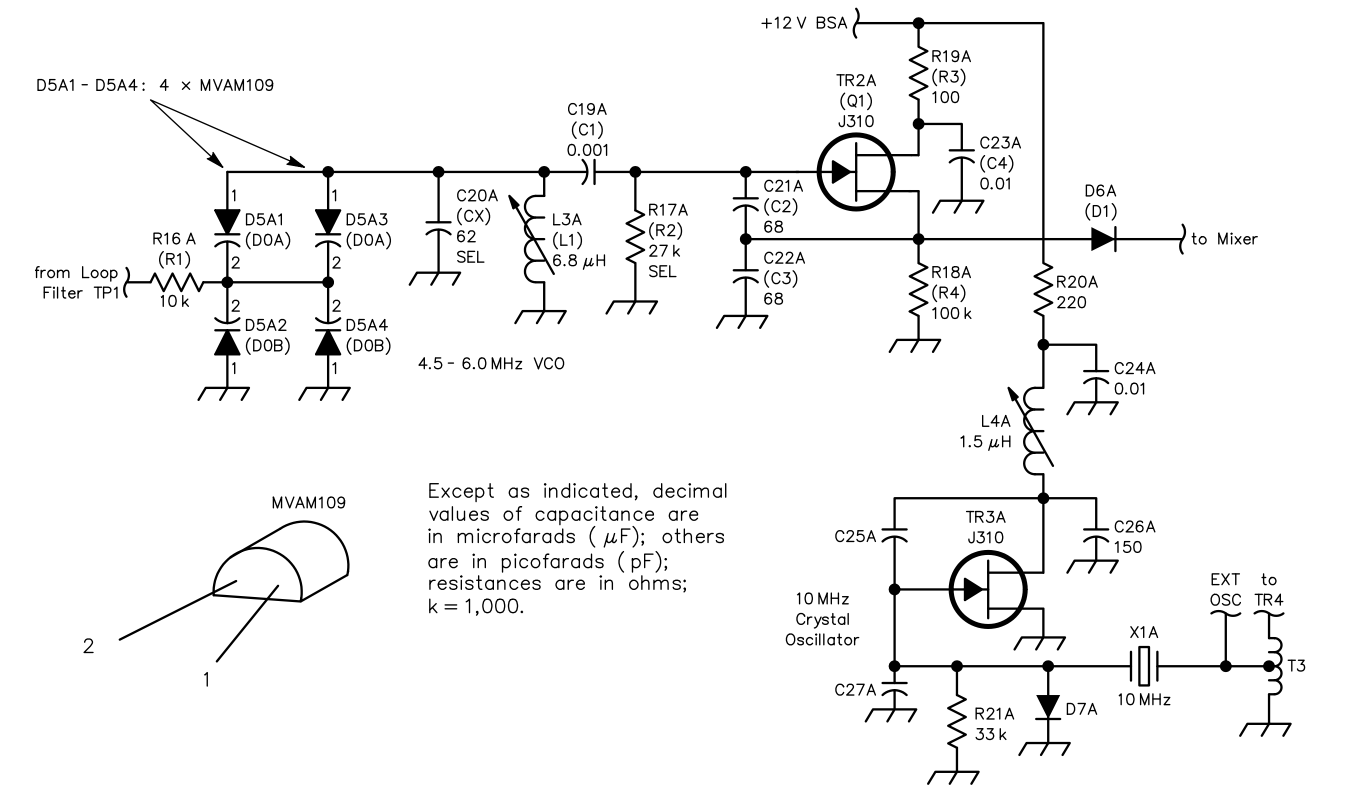


Fig 2—Hands Electronics PLL6 "band pack" schematic redrawn with new component values for VCO operation over the 4.5-6.0 MHz range. Note that, except for T3, these band-pack components are not supplied with the special PLL6 kit and must be obtained from other sources. Reference designators in parentheses are from the *2000 ARRL Handbook for Radio Amateurs*, Fig 14.52. Some part designations differ from ARRL style in order to comply with the Hands Electronics documentation. (Part sources: CS = Circuit Specialists; OSE = Ocean State Electronics, RS = RadioShack; DK = Digi-Key. See Notes for contact information.) Here is the Hands Electronics PLL6 kit band-pack parts list:

C19A—0.001 µF, 500 V (RS #272-126)
C20A—62 pF, 300 V, ±5% dipped mica, test-select value (see text, CS #DM10-620J)
C21A, C22A—68 pF, 300 V, 5% dipped mica (CS #DM10-680J)
C23A, C24A—0.01 µF, 500 V (RS #272-131)
C25A—0 pF (not used)
C26A—150 pF, 300 V, ±5% dipped mica (CS #DM10-151J)
C27A—0 pF (not used, see text)
D5A1-D5A4—MVAM109 tuning diode (4 required, OSE)
D6A, D7A—1N914 (RS #276-1620)
L3A—6.8 µH, ±6% adjustable Toko BTKANS-9441HM (DK #TK1416-ND)
L4A—1.5 µH, ±6% adjustable Toko BTKANS-9449HM (DK #TK1412-ND)
R16A—10 kΩ, 1/4 W, ±5%
R17A—27 kΩ, 1/4 W, ±5% (test-select, see text)
R18A—100 KΩ, 1/4 W, ±5%
R19A—100 Ω, 1/4 W, ±5%
R20A—220 Ω, 1/4 W, ±5%
R21A—33 kΩ, 1/4 W, ±5%
TR2A, TR3A—J310 (CS)
X1A—10.00 MHz, series, HC49, CTS (DK #CTX057-ND) see text

particularly useful for trial-and-error determination of circuit test-select values.

AC Stability of the PLL Loop

The VCO frequency is controlled by a feedback loop that compares the NE602 mixer output phase and frequency with those of the DDS output; it keeps the mixer output locked to the DDS. The dc component of the HCT4046 phase-frequency detector output is proportional to the phase difference between the input signal from the DDS and the feedback signal from the NE602 mixer. The loop includes a carefully designed filter to keep the feedback in the correct phase relationship with the input signal at loop frequencies where the open-loop gain exceeds 0 dB. Loops tend to oscillate if the feedback is in-phase (0°) and the open-loop gain exceeds 0 dB! Note that the filter response conditions the HCT4046 phase-frequency detector output to have a bandwidth of about 5 kHz, well below the 4.0-5.5 MHz range of the DDS and mixer output.

Loop stability is measured by phase margin (how close the open-loop phase is to 0° for 0 dB open-loop gain) and gain margin (how close the open-loop gain is to 0 dB for 0° open-loop phase). Ideal phase and gain margins are 45° and 30 dB or greater, respectively. If the open-loop gain is plotted versus frequency on a log-log graph, the designer will try to put the center of the –20 dB/decade portion of the curve at the 0-dB gain point.

A more complete discussion of loop design appears in *The ARRL Handbook* (pp 14.40-14.44). If you're interested in applying the *Handbook* discussion to the PLL6 board, or if you're redesigning the VCO and loop amplifier for another application, you'll need to calculate the loop parameters from the following equations:

$$f_{\text{pole }\#1} = 0\text{ Hz} \quad \text{(loop filter integrator)} \tag{Eq 1}$$

$$f_{\text{pole }\#2} = 0\text{ Hz} \quad \text{(VCO)} \tag{Eq 2}$$

$$f_{\text{pole }\#3} = \frac{1}{(2\pi)(R13)(C17)} \approx 12.94\text{ kHz in Fig 3} \tag{Eq 3}$$

$$f_{\text{pole }\#4} = \frac{1}{(2\pi)(R15)(C36)} \approx 159\text{ kHz in Fig 3} \tag{Eq 4}$$

$$f_{\text{zero}} = \frac{1}{(2\pi)(R12)(C15)} \approx 1.75\text{ kHz in Fig 3} \tag{Eq 5}$$

$$K_{\text{p}} = \text{phase detector gain} = \frac{V_{\text{cc}}}{4\pi} \approx 0.398\ \frac{V}{rad}\ \text{for HCT4046} \tag{Eq 6}$$

$$K_{\text{v}} = \text{VCO gain} = \frac{2\pi df}{dV} \approx 2.66\times10^{6}\ \frac{rad}{Vs}\ \text{in Fig 2, } C20A = 62\text{ pF} \tag{Eq 7}$$

$$\text{unity}-\text{frequency gain} = 20\log\left(\frac{K_{\text{p}}K_{\text{v}}}{(2\pi)^{2}(R10)(C15)}\right) \approx 139\text{ dB in Fig 3} \tag{Eq 8}$$

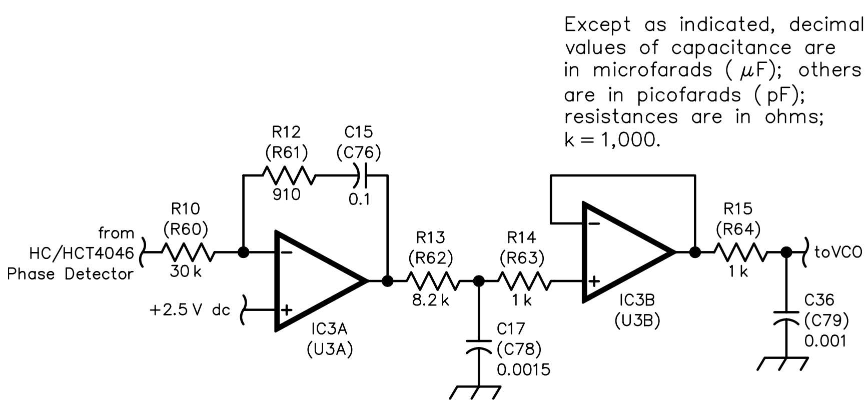


Fig 3—Redesigned PLL6 loop filter, with values calculated from KD9JQ *PLL3* program (reference designators in parentheses are from the *2000 ARRL Handbook for Radio Amateurs*, Fig 14.52). Some part designations differ from ARRL style in order to comply with the Hands Electronics documentation. Note that these components are not supplied with the special PLL6 kit, and must be obtained separately from other sources. Loop filter components not listed are supplied with the special PLL6 kit. (Part sources: RS = RadioShack. See Notes for contact information.)
C36—0.001 µF, 500 V (RS #272-126)
R10—30 kΩ, 1/4 W, ±5%
R12—910 Ω, 1/4 W, ±5%
R13—8.2 kΩ, 1/4 W, ±5%

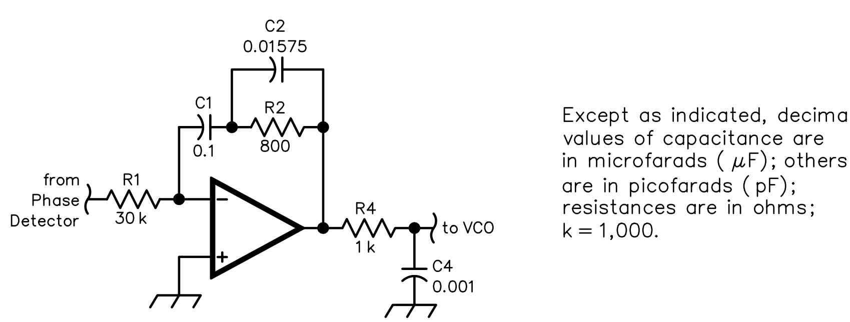


Fig 4—Values from KD9JQ PLL3 program (R4 and C4 values correspond to "modulation bandwidth" selected as 159 kHz).

Table 1—Loop parameters for entry to KD9JQ PLL3 program.

Desired Phase Margin in Degrees: 45
Loop Bandwidth (BW) in Hz: 5E3
VCO Tuning Sensitivity (K_v) Hz/V: 4.24E5 (measured with C20A = 62 pF)
Detector Constant (K_d) in V/R: 0.3979
Select Detector Type: 6 Tri-State Flip Flop P/F
Optional DC Gain (K_A) as Ratio: 1
Frequency Division Ratio: 1
VCO Modulation BW (B_v) in Hz: 1.59E5 (set by R4 and C4)
Op Amp DC Open Loop Gain (K_o): 2E5 (TL072 typical)
Op Amp 1st Pole Freq (B_o) in Hz: 20 (TL072 typical)
Reference Frequency (*FR*) in Hz: 4.75E6 (note that the actual DDS frequency range is 4E6 to 5.5E6)

The poles due to the low-pass filter at the output of the PLL6-board mixer can be neglected in the loop-stability analysis because the 5.5 MHz filter bandwidth is much greater than the 5 kHz loop bandwidth.

A graphing calculator can use the equations in Fig 14.47 of the *Handbook* article to plot the open-loop gain and phase response.[16] The *Handbook* equations assume ideal op-amp gain and bandwidth, and may be inaccurate when designing for wide loop bandwidths. The TL072 op amp has sufficient gain and bandwidth such that, for the 5 kHz loop bandwidth used in the modified PLL6 board design, the analysis is accurate.

Design Help from the Internet

If the preceding work seems too tedious, and if you have access to the Internet, you can download the very helpful *PLL3* program[17] from KD9JQ's Web site. After you enter the desired loop parameters (shown in Table 1), the program optimizes loop stability and calculates component values as shown in Fig 4. This results in calculated phase and gain margin values of 45° and 30 dB, respectively (in agreement with the TI-82 analysis). Like the stability analysis with the TI-82 graphing calculator, the *PLL3* program assumes that the phase shift in the low-pass filter at the output of the PLL6 board mixer is negligible. Unlike the TI-82 analysis, the program accounts for finite gain and bandwidth in the op amp. Note that the Fig 4 loop amplifier is different from the Fig 3 PLL6 circuit, so the following equations are required to convert the Fig 4 values for the Fig 3 circuit:

$$R10 = R1 \quad \text{(Eq 9)}$$

$$R12 = R2\left(1+\frac{C2}{C1}\right) \quad \text{(Eq 10)}$$

$$R13 = \frac{R2 \bullet C2}{C17} \quad \text{(Eq 11)}$$

$R14$ = not specified (select 1 kΩ) (Eq 12)

$$R15 = R4 \quad \text{(Eq 13)}$$

$$C15 = C1 \quad \text{(Eq 14)}$$

$C17$ = 0.0015 µF (select) (Eq 15)

$$C36 = C4 \quad \text{(Eq 16)}$$

The program also provides dynamic loop parameters, such as lock-in range and time (27.68 kHz and 54 µs, respectively, for Fig 3). Software controlling the DDS should restrict DDS frequency hops to less than the lock-in range; this is especially important for split (TX/RX) frequency operation. When necessary, the software can divide the commanded frequency change into a series of smaller hops, with each smaller hop meeting the lock-in-range requirement.

Help for the Mixer Design, Too

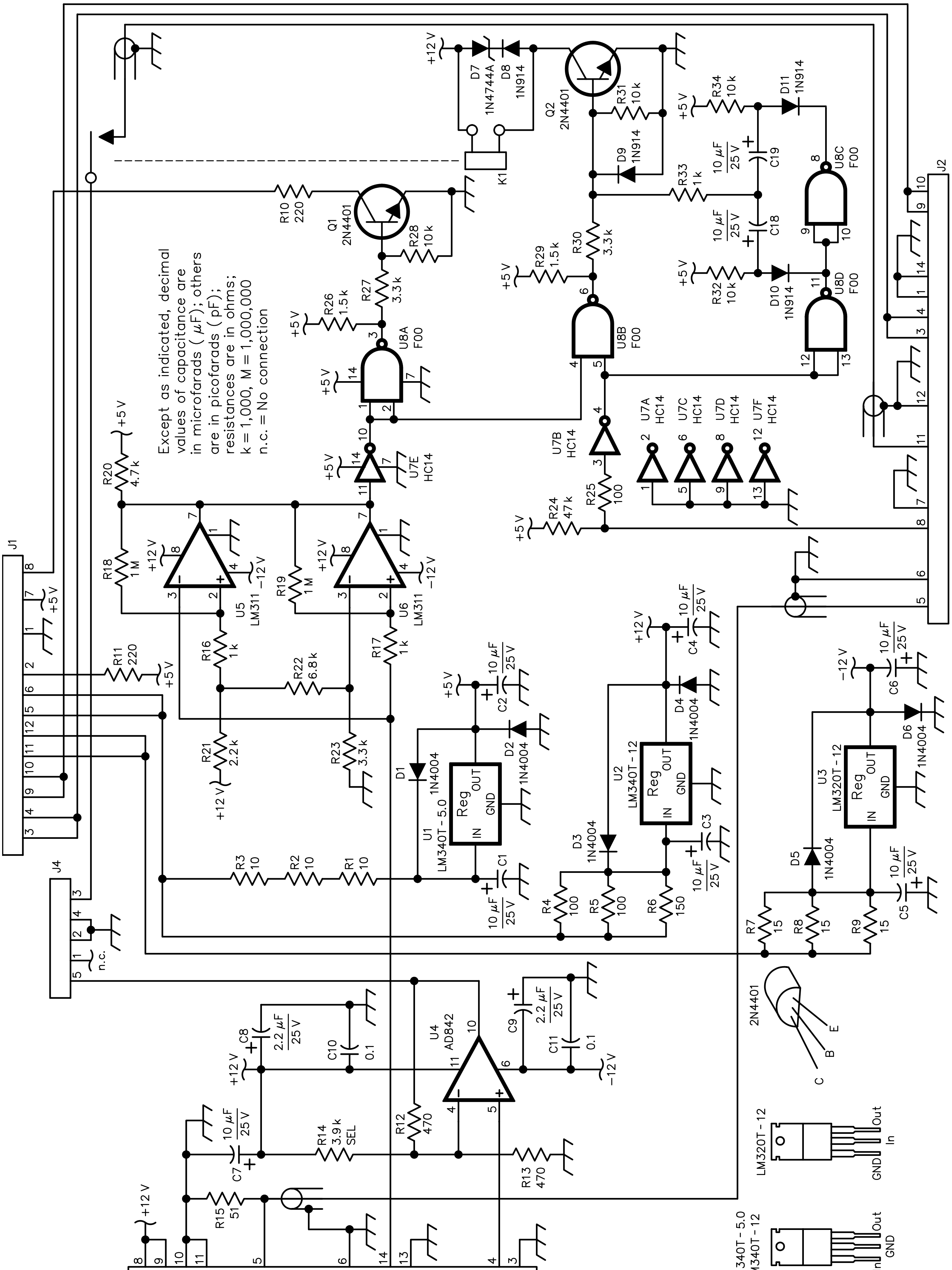
The NE602 mixer output contains not only the desired 4.0-5.5 MHz difference frequency (between the VCO and crystal-oscillator fundamentals), but also other difference frequencies (between the fundamental and various harmonic combinations) that fall within the mixer output low-pass-filter passband. Similar problems with undesired mixer products may exist in the HF transceiver if the VCO range is increased for extended receiver operation.

Mixer analysis can seem overwhelming because each of many interactions must be separately analyzed. Fortunately, Hittite Microwave Corp provides a fine Web site with a graphical tool, the *Mixer Spur Chart Calculator*.[18] Although the tool is designed for analysis of microwave mixers, it can also analyze products for the NE602 and within the HF transceiver.

After you reach their Web page, simply enter the VCO RF range (4.5-6.0 MHz). We'll ignore the legends that say "GHz" and assume a scale factor of 1000. Enter an IF range of 0-5.5 MHz and an LO frequency (crystal oscillator) of 10 MHz. Leave the LO power and RF power at their default values.

After clicking the "Calculate" button, the spurs will be displayed as red lines. Clicking on a line will display the power (dBc) and order (N×M) of that spur. The displayed power for each spur does not apply directly to the NE602, since the displayed power is based on a simulation of a particular

Fig 5—Lock Detector board schematic: Connect C12-C17 (not shown) between ground and each power pin of U5 through U8. (Part sources: Allied = Allied Electronics; CS = Circuit Specialists; RS = RadioShack. See Notes for contact information.)
C1-C7, C18, C19—10 µF, 35 V dipped tantalum (CS #TAC008)
C8, C9—2.2 µF, 35 V dipped tantalum (CS #TAC006)
C10-C17—0.1 µF, 50 V monolithic ceramic (Allied #1C10Z5U104M050B)
D1-D6—1N4004 (CS)
D7—1N4744A 15 V Zener diode, (RS #276-564)
D8-D11—1N914 (RS #276-1620)
J1-J3—14-pin DIP socket (CS #I314BOZ)
J4—8-pin DIP socket (CS #I308BOZ)
K1—SPST reed relay, 12 V (RS #275-233) Socket pins are made from a pin-line socket (CS #40-0518-10)
Q1, Q2—2N4401 NPN, TO-92, (CS). Sockets for Q1 and Q2 pins are made from pin-line sockets (CS #40-0518-10)
R1-R3—10 Ω, 1 W, ±5% (RS #271-151)
R4, R5—100 Ω, 1/2 W, ±5% (RS #271-1108)
R6—150 Ω, 1/2 W, ±5% (RS #271-1109)
R4-R6 test-selected, see text
R7-R9—15 Ω, 1/2 W, ±5% (RS #271-1102)
R10, R11—220 Ω, 1/4 W, ±5%
R12, R13—470 Ω, 1/4 W, ±5%
R14—3.9 kΩ, 1/4 W, ±5%, (test-select, see text)
R15—51 Ω, 1/4 W, ±5%
R16, R17, R33—1 kΩ, 1/4 W, ±5%
R18, R19—1 MΩ, 1/4 W, ±5%
R20—4.7 kΩ, 1/4 W, ±5%
R21—2.2 kΩ, 1/4 W, ±5%
R22—6.8 kΩ, 1/4 W, ±5%
R23, R27, R30—3.3 kΩ, 1/4 W, ±5%
R24—47 kΩ, 1/4 W, ±5%
R25—100 Ω, 1/4 W, ±5%
R26, R29—1.5 kΩ, 1/4 W, ±5%
R28, R31, R32, R34—10 kΩ, 1/4 W, ±5%
U1—5 V positive regulator, TO-220, LM340T-5.0 (Allied)
U2—12 V positive regulator, TO-220, LM340T-12 (Allied)
U3—12 V negative regulator, TO-220, LM320T-12 (Allied)
U1-U3 each mounted with heat sink: Aavid #577202B00000 (Allied) heat-sink hardware (RS #276-1373). Heat sink grease (RS #276-1372)
U4—op amp, 14-pin DIP, AD842JN (Allied)
U5, U6—voltage comparator, 8-pin DIP, LM311N (Allied). Sockets for U5 and U6 each 8-pin DIP socket (CS #I308BOZ)
U7—hex inverter, 14-pin DIP, 74HC14N (CS)
U8—quad NAND, 14-pin DIP, 74F00PC (Allied). Sockets for U4, U7 and U8 each 14-pin DIP socket (CS #I314BOZ)


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 n.c. = No connection
J1
J2
J3
J4
K1
Q1 2N4401
Q2 2N4401
U1 LM340T-5.0
U2 LM340T-12
U3 LM320T-12
U4 AD842
U5 LM311
U6 LM311
U8A F00
U8B F00
U8C F00
U8D F00
U7A HC14
U7B HC14
U7C HC14
U7D HC14
U7E HC14
U7F HC14
D7 1N4744A
D8 1N914
D9 1N914
D10 1N914
D11 1N914
R14 3.9 k SEL
2N4401
LM320T-12
LM340T-5.0
LM340T-12

Hittite microwave mixer; but they may be useful as relative indicators. Note that this program assumes the use of a doubly balanced mixer (all three ports are isolated from each other). The PLL6 board uses a singly balanced mixer (LO port isolated from the other two). Thus, the PLL6 board mixer may provide more VCO feedthrough to the output than indicated by the Mixer Spur Chart Calculator.

Also included at the Web site is a help page and a reprint of a *Microwave Journal* article giving the theoretical basis for its Java program. The program is for on-line use only; it cannot be downloaded.

For the 4.5-6.0 MHz VCO and 10-MHz LO used on the modified PLL6 board, the program designates the desired mixer product as (1)×(–1) at 0 dBc. The strongest spur that could fall in the 5-kHz loop bandwidth is designated (2)×(–3) and is shown as –84 dBc (see the preceding comment about indicated spur power). This spur would fall in the 5-kHz loop bandwidth when the (2)×(–3) spur frequency is within ±5 kHz or so of the desired (1)×(–1) mixer output, which is locked to the DDS. This occurs for VCO, mixer output and DDS frequencies in the neighborhood of 5.0 MHz.

Thus for the HW-101, the worst-case spur problems occur at TX/RX frequencies at the top of each band: 4.0 MHz (80 meters), 7.5 MHz (40 meters), 14.5 MHz (20 meters), 21.5 MHz (15 meters), 28.5 MHz, 29.0 MHz, 29.5 MHz and 30.0 MHz (10 meters). Qualitative receiver checks with the HW-101 do not indicate phase-noise problems at these frequencies. A better test would be to measure the PLL unit (or HW-101) output for phase noise using a spectrum analyzer. If a 9.8304 MHz crystal (series-resonant, HC-49, Digi-Key X421-ND) is substituted for the 10.00 MHz crystal at X1A on the PLL6 board, any spur-related phase noise would be moved from the HW-101 upper band edge to about 85 kHz above the upper band edge.

The Lock Detector Board: How It Works

Fig 5 is the schematic for the Lock Detector board. It interfaces to the front panel, rear panel, PLL6 board and the FL1-40 low-pass filter board via J1, J2, J3 and J4, respectively (see schematics of Figs 8, 9, 11 and 12). U1 through U3 are the regulators for the +5 V, +12 V and –12 V power supplies. Input power is supplied from two ac/dc wall adapters connected to the rear panel, and the rear panel is connected to the Lock Detector board via J2. The front-panel power switch and red **POWER** LED indicator are connected to the Lock Detector board via J1.

The U4 circuit is a voltage amplifier (6-dB gain) boosting the PLL6 **SYN OUT** signal to the 2-3 V (RMS) level required at my HW-101 transceiver's VFO input. The low-pass filter, mounted on the Lock Detector board, reduces the harmonic content of U4's circuit output. The filter is connected to U4's circuit via J4.

Window comparator U5/U6 monitors the PLL6 VCO **EXT TUNE** voltage via J3. Resistors R21 through R23 form a resistive voltage divider, setting the 3-10 V dc window. If the PLL becomes unlocked, the tuning voltage falls outside the window; the comparator outputs go low, removing base drive from Q1 and extinguishing the green **PLL LOCK** LED (connected via J1).

The signal at U7 pin 3 is provided from the contacts of the HF transceiver TX/RX relay (connected via J2); it is grounded when transmitting. The low signal at U7 pin 3 (transmit only) results in a high signal at U8 pin 5 that with a high signal at U8-4 (after PLL loses lock) opens the ground lead of relay K1. The contacts of K1 interrupt the VFO signal to the HF transceiver (also connected via J2).

The Lock Detector logic assumes semi-break-in CW keying, as used in the Heath HW-101 transceiver. In the CW mode, the transceiver does not switch back to receive until the end of a transmitted word or sentence. Thus, the logic is designed to momentarily open the K1 contacts (interrupting the VFO signal) immediately following an RX/TX or TX/RX transition. This feature ensures that the VCO frequency movements between RX and TX frequencies (when operating split) are not transmitted. While receiving, C18 is charged to +5 V, R29 provides base current to Q2, the coil of relay K1 is energized and the relay contacts are closed. Except during RX/TX and TX/RX transitions, there is no current in R33. Immediately after switching from receive to transmit, however, U7 pin 3 and U8 pin 11 go low, discharging C18 through R33 and forward-biased diodes D9 and D10. Q1 is switched off as long as D9 is forward biased (until C18 is fully discharged). Q1 is similarly switched off just after a transmit-to-receive transition, when C19 is discharged through R33 and forward-biased diodes D9 and D11. Clearly, this momentary interruption of the VFO is incompatible with full QSK at higher CW speeds, since the combined C18 (or C19) discharge and charge times (over 100 ms) may be longer than a Morse dot, dash or character. Components C18, C19, D9-D11 and R32-R34 can be deleted if the board is to be used with full break-in operation; however, split TX/RX CW operation would not be advised.

Building the Lock Detector Board

The Lock Detector board is hand wired on a 5.2×5.4-inch prototype board with four-hole solder pads and interleaved buses, and with holes centered on a 0.1-inch grid (Odyssey Marketing 2300-T). To ensure a low-inductance ground, the solder pads in every ninth row (rows 1, 10, 19, etc) are soldered together across the board (perpendicular to the buses) and soldered to the buses. All ICs (except regulators) are socketed, using 8- and 14-pin DIP sockets with machined pins (Circuit Specialists #I308BOZ and #I314BOZ). To keep lead lengths short and reduce the amount of hand wiring required, sockets are aligned with respect to each other to maximize the use of the solder pads for connections between components. Bypass capacitors are located as close to the IC power pins as possible. Ceramic bypass capacitors C10-C17 are located on the wiring side of the board, as are D7, D8 and R12. Except for C1-C9, D1-D6 and R1-R11, which are mounted directly on the component side of the board, all remaining passive components are mounted on socketed 8- and 14-pin DIP headers (Circuit Specialists 8-600-10 and 14-600-10). Sockets for K1, Q1 and Q2 pins are made from pin-line sockets (Circuit Specialists #40-0518-10). The photo in Fig 6, a top view of the PLL unit, shows both boards with the Lock Detector board components identified. Note the FL1-40 low-pass filter mounted at the upper end of the Lock Detector board; it is attached using insulated standoffs shortened to half-height (RS #276-1381).

Note that the parts list identifies R4 through R6 as test-selected resistors. Choose the parallel combination of these resistors (37.5 Ω in my board) so that the voltage at the input pin of U2 is +15-16 V with all boards connected, the PLL unit locked to the DDS and connected to the transceiver. R4 through R6 dissipate some of the power that would otherwise heat U2, thus reducing its junction temperature. The required parallel-resistor value will depend on the loaded output voltage of your wall adapter. I used Philmore MW 1830 ac/dc wall adapters, rated at 18 V dc/300 mA.

R14 is also a test-selected resistor. Choose this resistor for minimum dc offset of the 4.5-6.0 MHz PLL unit output waveform. My board required 3.9 kΩ at R14.

Connecting It All Together

The PLL6 and Lock Detector boards are mounted in a Sescom[19] MC-10A enclosure using 10-mm (0.39-inch) standoffs (RS #276-1381). This PLL enclosure matches the MC-12A enclosure used for my DDS unit (see Fig 7).

The Lock Detector board has four DIP sockets (J1-J4) to connect to the front panel, rear panel, PLL6 board and FL1-40 low-pass filter. The front-panel interconnect schematic is shown in Fig 8. Fig 6 shows the wiring and P101 connected to J1 on the Lock Detector board. Front-panel components are visible in Fig 7.

The rear-panel interconnect schematic is shown in Fig 9. Be sure to electrically isolate the dc-power input jacks (J203 and J204) from the rear panel. Fig 6 shows the wiring and P201 connected to J2 on the Lock Detector board. The rear-panel connectors are shown in Fig 10. *Important note:* To avoid arcing between the wall adapter plugs and J203-J204, always disconnect the adapters from the ac power line before connecting or disconnecting the plugs and jacks!

The PLL6 interconnect schematic is shown in Fig 11. Note that a 10-kΩ resistor is connected between the **SYN OUT** pin and ground on the PLL6 board. Fig 6 shows the wiring and P301 connected to J3 on the Lock Detector board. The output-filter interconnect schematic is shown in Fig 12. Fig 6 shows the wiring and P401 connected to J4 on the Lock Detector board (notice that R401 is located on the P401 header).

Checkout

Follow the instructions in the Hands Electronics documentation before powering up the PLL6 board.

Before installing ICs on the Lock Detector board, check all signal paths for continuity and isolation from ground and power supplies. Before applying power, connect all of the circuits: Lock Detector board to the front panel (J1 and P101), rear panel (J2 and P201), PLL6 board (J3 and P301) and FL1-40 board (J4 and P401).

Then check for short circuits on signal paths between assemblies and on the +12 V, –12 V and +5 V power buses. Also, check for continuity on signal paths between assemblies.

Disconnect P301 from J3, apply power to the Lock Detector board and check for the proper voltages on the power buses. Next, after reconnecting P301 to J3, align the PLL6 board using

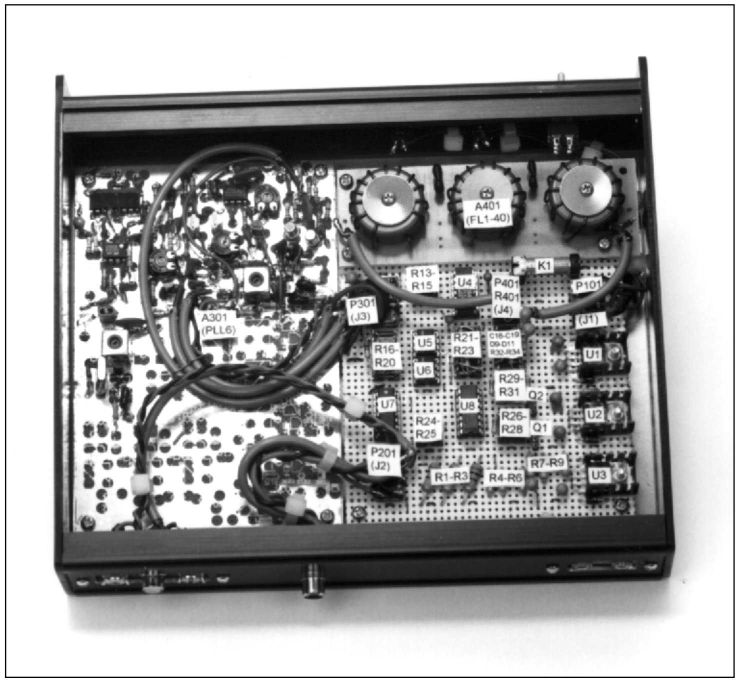


Fig 6—Top view, showing PLL6 board at left and Lock Detector board at right. Lock Detector board DIP headers are used to mount passive components and as plugs for connection to the PLL6 board, output filter and front and rear panels.

Fig 7—PLL unit atop matching DDS unit, with HW-101 HF transceiver below. On PLL unit front panel, from left to right, S101 (power switch), D101 (red POWER LED indicator) and D102 (green PLL LOCK LED indicator).

the instructions in the Hands Electronics documentation. While aligning the PLL6 board, the green **PLL LOCK** LED should illuminate while the VCO tuning voltage is in the 3-10 V dc window of the Lock Detector board window comparator. If you can monitor the VCO output waveform with an oscilloscope, R17A may be selected for minimum waveform distortion at the PLL6 **SYN OUT** pin. If you installed a 30 pF trimmer capacitor at C27A, adjust the PLL6 LO for zero beat to WWV (assuming there is a 10 MHz crystal at X1A).

While monitoring the FL1-40 filter output at J202, select R14 on the Lock Detector board for minimum dc offset on the output waveform. Then check the voltage at the U2 input pin. If necessary, select the parallel combination of R4-R6 such that the U2 input voltage is +15-16 V.

To assure that your transmitted signal meets the FCC requirements for spurious emissions, check the PLL unit output at J202 for spurs. (Current regulations require spurs to be at least –40 dB below the carrier for transmitted power levels up to 500 W.) The FL1-40 low-pass filter at the PLL unit's output keeps harmonics of the VCO frequency well below the FCC requirements, but doesn't ensure freedom from PLL phase-noise sidebands. If possible, monitor the output with a spectrum analyzer, looking for spurs up to about 100 kHz from the carrier. If a spectrum analyzer is not available, you can listen for PLL spurs on an HF transceiver, while using the PLL unit connected between the transceiver and the DDS. Tune the DDS to a strong signal (a CW code-practice station is ideal), and tune the DDS on both sides of the signal. Listen for a raspy replica of the signal (several may be present) up to 100 kHz on either side of the signal. It helps to have a second HF receiver left tuned to the signal while doing this, to verify that the raspy signal is actually associated with the strong signal.

Before installing the 62-pF capacitor at C20A, my own PLL unit exhibited strong spurs up to ±17 kHz from the VCO frequency, while tuning to either side of the W6ADO code-practice station frequency at 7098 kHz. In general, PLL spurs are indicated if raspy-sounding CW signals (or heterodynes) are heard among normal signals.

Certain types of PLL spurs may indicate an inadequate phase margin in the PLL. This may result from incorrect component values in the loop filter, unanticipated phase lag caused by stray capacitance, a poor choice of loop-compensation components or other causes. Other causes of PLL noise can include: pickup of ac-power or other signals, inadequate decoupling at IC power pins, noise in the tuning diodes or other components, VCO feedthrough in the mixer, insufficient loop bandwidth or jitter in the VCO, mixer LO or DDS clock.

PLL phase noise may be accentuated at low reverse-bias levels. This effect is more pronounced with increasing temperature. The 62-pF capacitor at C20A increases the tuning diode's reverse bias by about 1 V at the lower end of the VCO range. More capacitance at C20A will further increase its reverse bias, but will reduce the VCO K_v and tuning range. Lower capacitance at C20A will have opposite effects. Be sure to

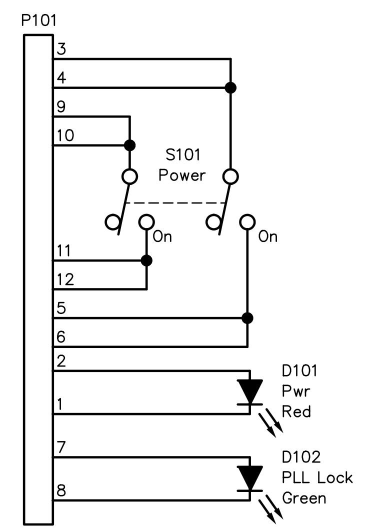


Fig 8—Front-panel interconnect schematic. (Part sources: CS = Circuit Specialists. See Notes for contact information.)
D101—LED indicator assembly, red (CS #35CA004)
D102—LED indicator assembly, green (CS #35CA005)
P101—14-pin DIP header (CS #14-600-10)
S101—DPDT miniature toggle switch (CS #8011)

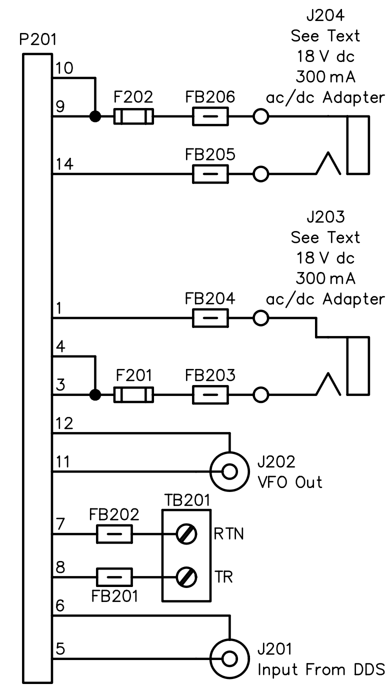
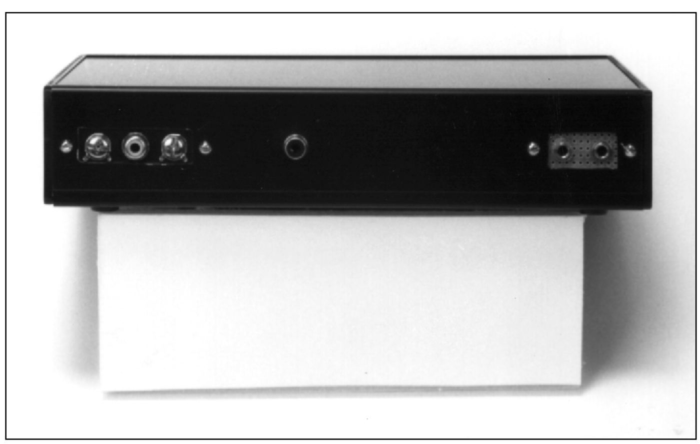


Fig 9—Rear-panel interconnect schematic. (Part sources: CS = Circuit Specialists; OSE = Ocean State Electronics, RS = RadioShack; DK = Digi-Key. See Notes for contact information.)
F201, F202—1/2-A, slow-blow fuse, 5×20 mm (RS #270-1061). In-line fuse holder, 5×20 mm (RS #270-1238)
FB201-FB206—ferrite beads (CS #FBPK-1)
J201, J202—phono jack (RS #274-346)
J203, J204—1/8-inch phone jack, two conductor (RS #274-251, both jacks are mounted on a small perf-board plate for electrical isolation from the rear panel.)
P201—14-pin DIP header (CS #14-600-10)
TB201—2-terminal bakelite terminal board (OSE #15-72) My unit used RS #274-620 (no longer available) in place of J201 and TB201.

Fig 10—Rear view showing, from left to right, combined J201/TB201 (input from DDS/TR from HW-101), J202 (VFO out), J203 and J204 (dc power from wall adapters).



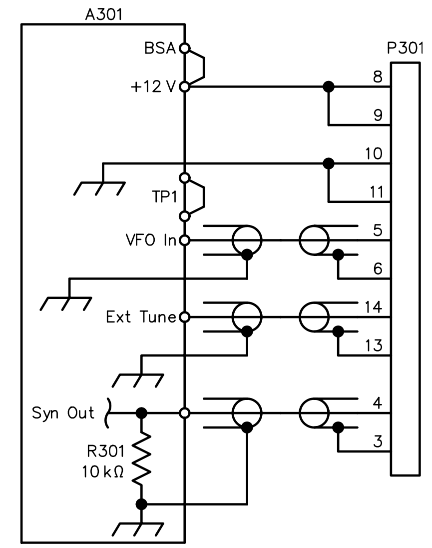


Fig 11—PLL6 board interconnect schematic. (Part sources: CS = Circuit Specialists; K = Kanga US. See Notes for contact information.)
A301—Hands Electronics special PLL6 board supplied without band packs (K)
P301—14-pin DIP header (CS #14-600-10)
R301—10 kΩ, ¹/₄ W, ±5%

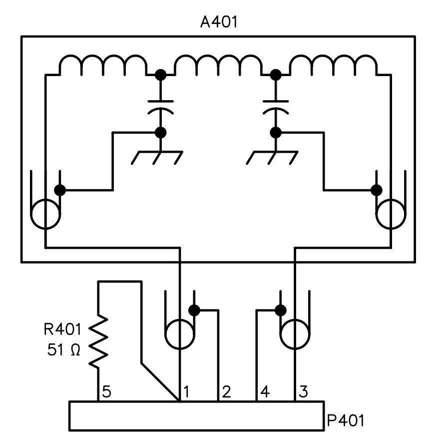


Fig 12—Output filter interconnect schematic. (Part sources: CC = Communication Concepts; CS = Circuit Specialists. See Notes for contact information.)
A401—40-meter low-pass filter (CC #FL1-40)
P401—8-pin DIP header (CS #8-600-10)
R401—51 Ω, ¹/₄ W, ±5%

check for PLL spurs and apply corrective measures before using the PLL unit for transmitting on the air. It could save you from a citation!

Summary

If I could grade the performance of my HW-101 transceiver with PC-controlled DDS before and after adding the PLL Spur Eliminator, the "before" report card would read "D-minus, marginally useful. Where did all these detestable tones come from?" The "after" report card would read "B-plus—Hey, this DDS really does work! What happened to those tones?"

The moral of this story: "PLL—don't try DDS without it!"

Notes

[1]U. L. Rohde, KA2WEU, *Microwave and Wireless Synthesizers—Theory and Design* (New York: John Wiley and Sons, 1997), pp 48-53, 117-118, 167-168.

[2]D. Smith, KF6DX, "Signals, Samples and Stuff: A DSP Tutorial (Pt 2)," *QEX*, May/June 1998, pp 28-29.

[3]Qualcomm, Inc, ASIC Products, "Hybrid PLL/DDS Frequency Synthesizers—Application Note," *Synthesizer Products Data Book*, pp 9-1 to 9-10; available at the Qualcomm Web site: **http://www.qualcomm.com/cdmatechnologies/vlsi/documents/SynthesizerDataBook.pdf**.

[4]B. Hodgkinson, VA3BH, "Julieboard—An Easy-to-Build DDS Synthesizer for the PC Printer Port," *73*, August 1993, pp 40-46.

[5]J. Craswell, WB0VNE, "Weekend DigiVFO," *QST*, May 1995, pp 30-32.

[6]R. D. Straw, N6BV, Editor, *The 2000 ARRL Handbook* (ARRL: Newington, Connecticut, 1999), "Frequency Synthesizers," pp 14.33-14.53 and pp 29.7-29.8.

[7]Hands Electronics, Tegryn Llanfyrnach Pembs, SA35 0BL, UK; tel +44 (0)1239 698427, fax +44 (0)870; e-mail **hands@rf-kits.demon.co.uk**; URL **http://www.rf-kits.demon.co.uk**.

[8]Kanga US (US distributor for Hands Electronics), 3521 Spring Lake Dr, Findlay, OH, 458401; tel 419-423-4604; e-mail **kanga@bright.net**; URL **www.bright.net/~kanga/kanga**

[9]Communication Concepts Inc, 508 Millstone Dr, Beavercreek, OH, 45434-5840; tel 937-426-8600, fax 937-429-381; e-mail **cci.dayton@pobox.com**; URL **http://www.communication-concepts.com**.

[10]Allied Corporate Headquarters, 7410 Pebble Dr, Fort Worth, TX 76118; tel 817-595-3500, fax 817-595-6444; URL **http://www.alliedelec.com/**

[11]Circuit Specialists Inc, 220 S Country Club Dr #2, Mesa, AZ 85210-1248; tel 800-528-1417 or 480-464-2485, fax 480-464-5824 (telephone hours: 8AM-6PM MST Monday-Friday); e-mail **info@cir.com**; URL **http://www.web-tronics.com**.

[12]Digi-Key Corp, 701 Brooks Ave S, Thief River Falls, MN, 56701-0677; tel 800-344-4539 or 218-681-6674, fax 218-681-3380; URL **http://www.digikey.com**.

[13]Ocean State Electronics, PO Box 1458, Westerly, RI 02891; tel 401-596-3080, fax 401-596-3590; e-mail **ose@riconnect.com**; URL **http://www.oselectronics.com**.

[14]RadioShack: See your local store or **http://www.radioshack.com**.

[15]Part Miner part-locating program for use on-line at: **http://www.partminer.com**.

[16]I have written an open-loop-gain phase-response-plotting program for the TI-82. You can download a command list from the ARRL Web **http://www.arrl.org/files/qex/**. Look for PETELOOP.ZIP.

[17]Download *PLL3* software from KD9JQ's site: **http://www.mbn.net/kd9jq/hamradio/kd9jq.html**.

[18]*Mixer Spur Chart Calculator* software, a Java program for use on-line at: **http://www.hittite.com/spurchart/SpurCalculator.htm**.

[19]Sescom Inc, 2100 Ward Dr, Henderson, NV 89015-4249; (orders only weekdays, 8AM to 4PM, PST) tel 800-634-3457, fax 800-551-2749; e-mail **sescom@sescom.com**; URL **http://www.sescom.com/**.

Rick Peterson, WA6NUT, has been interested in radio since boyhood, when his father helped him build his first crystal set (using a homemade crystal detector made from lead and sulfur). Rick was introduced to Amateur Radio by his uncle, Art Hendricks, W6BKU (later K4CTZ, now an SK). At the age of 12, Rick had access to the BC-348 receiver in his uncle's ham shack while Art was away at college, and enjoyed listening to local ragchewing on 75-meter AM. After SWLing for several years with shortwave receivers salvaged from trash barrels, Rick received his Amateur Radio license in 1960. He was active on 6-meter AM while in college, with a Heathkit "Sixer" transceiver, then a Globe Scout 680-A transmitter and a Hallicrafters 6-meter receiver

He received his BS in Engineering from UCLA in 1963. He retired as a Senior Design Engineer in 1995, after 30 years with General Motors (Delco Systems Operations) in analog circuit design (analog I/O design for aerospace guidance and control computers). He is now employed as a substitute middle- and high-school math and science teacher in Lompoc, California.

While Rick was employed as an engineer, he had little time for serious Amateur Radio but dabbled a little in RTTY (AFSK on 2-meter AM) and radio astronomy (as publisher of a newsletter, The Radio Observer*). Since his retirement, he has renewed his interest in CW (a member of ARRL and the Fists CW Club, he is building his code speed with help from the W6ADO 40-meter code-practice transmissions). Since reading VA3BH's 1993 article on PC-controlled DDS (see Notes), Rick's main technical interest has been developing hardware and software for PC control and display for his Heath HW-101 HF transceiver. The system is to include CW encoding from the keyboard and decoded CW displayed on the PC monitor.*

Science in the News

A New, High-Speed Data-Transmission Mode is Being Developed that may Shock Everyone

Telecommunications lines gushing data at terabits-per-second now link most American cities, but few homes have access to even a megabit of that torrent. The last-mile bottleneck continues to pinch the promise of the Internet. Even with DSL technology, telephone lines are still relatively slow and inefficient.

Coaxial (CATV) cables aren't a universal solution. Fiber-optic cables stop before they get close to most neighborhoods; even now, half the homes in Dallas and some entire communities in sprawling Los Angeles, for instance, have yet to be wired for cable. In addition, many such CATV Internet access schemes are high-speed in one direction only.

Hybrids of existing technologies are popular avenues for remedy. Almost daily, we hear of another multibillion-dollar merger as telephone, cable and entertainment entities try to find better ways to bring more information to us. One possibility may turn out to be as close and as universal as an ordinary wall socket.

A new high-tech company in Dallas has unveiled a technology that promises a solution to Internet anxiety by delivering multi-gigabit-per-second Internet access, including voice and video data, to schools, households and businesses via the most extensive wired network on earth: the electrical power grid. Almost 85% of the world is wired for electricity; compare that with the 12-15% wired for phone service.

The company, privately held Media Fusion, was founded in 1998 by former Microsoft scientist William "Luke" Stewart. Media Fusion's proprietary technology uses the electrical power grid to transmit communications signals at very high speed; this would mitigate the public funding needed to equip schools and other under-served areas with expensive CATV or fiber connections. Anyone with a simple electrical outlet would be able to plug into telephone, television and Internet services.

The technology's potential has created a growing buzz in utility and government circles. Top congressional officials say they will even seek government funding for the project if the private sector doesn't come through. "Their plans include a small device that plugs into any wall socket," US Representative Billy Tauzin (Republican, Louisiana), chairperson of the House Commerce Telecommunications Subcommittee, said in a speech last year. "You could simply connect your television, your telephone, and your PC to them and immediately get those services under this system."

Individual consumers could get network connections of 2.5 gigabits per second; in its own estimation, the company calls this "highly conservative." Even at that "modest" speed, one could download the entire contents of a 10 GB hard drive in a few seconds, have interactive "telemedicine," high-quality, real-time video conferencing or easily watch a movie downloaded from the Web.

Plug It In, Plug It In

Stewart calls his chip-laden devices "night lights." Several of them would cost about $60. This technology could help bridge the growing "digital divide," finally offering the Internet—and all the medicine, education and e-commerce that implies—to rural America (where it's too expensive to string fiber-optic lines or cable to just a few homes per square mile) as well as to underdeveloped countries, most of which the Information Age has skipped entirely.

"We are developing a system with near-limitless capacity that will increase data transfer rates to the exabit-per-second [10^{18}] region," Stewart said. "With these incredible transmission capabilities, you'll also see dramatic advances in hardware and software products that will further facilitate operating speeds for computers and computing—including the way computers talk to each other and share work."

Scientists and engineers have been trying to discover a way to send high-speed information over power lines for 100 years, but line noise, electrical load imbalances and intervening transformers prevented much success. Stewart found a way to avoid the distortion and signal loss that occurs when you try to send information through copper wires that are also carrying high voltage and current. He says it works in the same way a laser uses optics to amplify signals during stimulated emission.

"The only other technologies that might qualify as competition are those trying to increase bandwidth for communications delivery while simultaneously attempting to reduce costs associated with infrastructure build-out," said Stewart, whose expertise with signal processing, microwave technologies, supercomputing and neural networks has contributed to numerous US military defense projects and advanced private-sector innovations. "Most are working to push the limits of traditional transport media with only sporadic success."

Last year, a consortium led by Nortel abandoned a European test project that was attempting to determine the feasibility of using the electrical grid to communicate. While the technology itself was deemed feasible, high cost and other concerns forced the group to "pull the plug." Media Fusion's technology is in alpha test at the John C. Stennis Space Center in Mississippi.

Some Doubts

"If it works, it changes the world," says John Fike, Associate Professor of Engineering Technology at Texas A&M. Fike has written two books on electronics and data communications. "People have been looking at all that copper out there for a long time. I hope he's right." The technology has attracted its share of scoffers, who think the notion of unlimited bandwidth on power lines is little more than a surge of fantasy. Media Fusion might also have to seek changes in FCC rules that limit conducted emissions on the ac mains and deal with potential interference to other services caused by radiation.

Intentional Part-15 radiators now share industrial, scientific and medical (ISM) bands with the Amateur Radio Service; these are already being fielded to provide high-speed Internet access (11 Mbits per second and higher) around the world. Such systems also promise to reach outlying areas not served by any wired network at a reasonable subscriber cost. Those of us interested in preserving our Amateur Radio resources should question the impact of both these schemes on our microwave bands.—*Douglas Page, 608 Esplanade – 7, Redondo Beach, CA 90277-4174;* ***dpage@arrl.org***

A Calibration Source for DC and AC Voltmeters

Reliable measurements require test equipment that is adequately calibrated. Build this source of stable and dependable ac and dc calibration voltages for your test bench.

By Wayne J. Stanley, W4RDG

The voltage-calibrator circuit shown in Fig 1 enables periodic verification of the measurement accuracy of voltmeters to within 0.5%(dc) and 1.0%(ac). This includes most general-purpose analog or digital multimeters normally found in the ham shack.

The calibrator consists of five building blocks as follows:

1. A precision 10-V dc source
2. A precision 5-V RMS 1-kHz sine-wave ac source
3. A 10-step attenuator and dc amplifier for use with either source
4. A dual-polarity, 107-V regulated dc power supply
5. A dual-polarity, 15-V regulated dc power supply

PO Box 10308
Winston-Salem, NC 27108

DC Reference Source

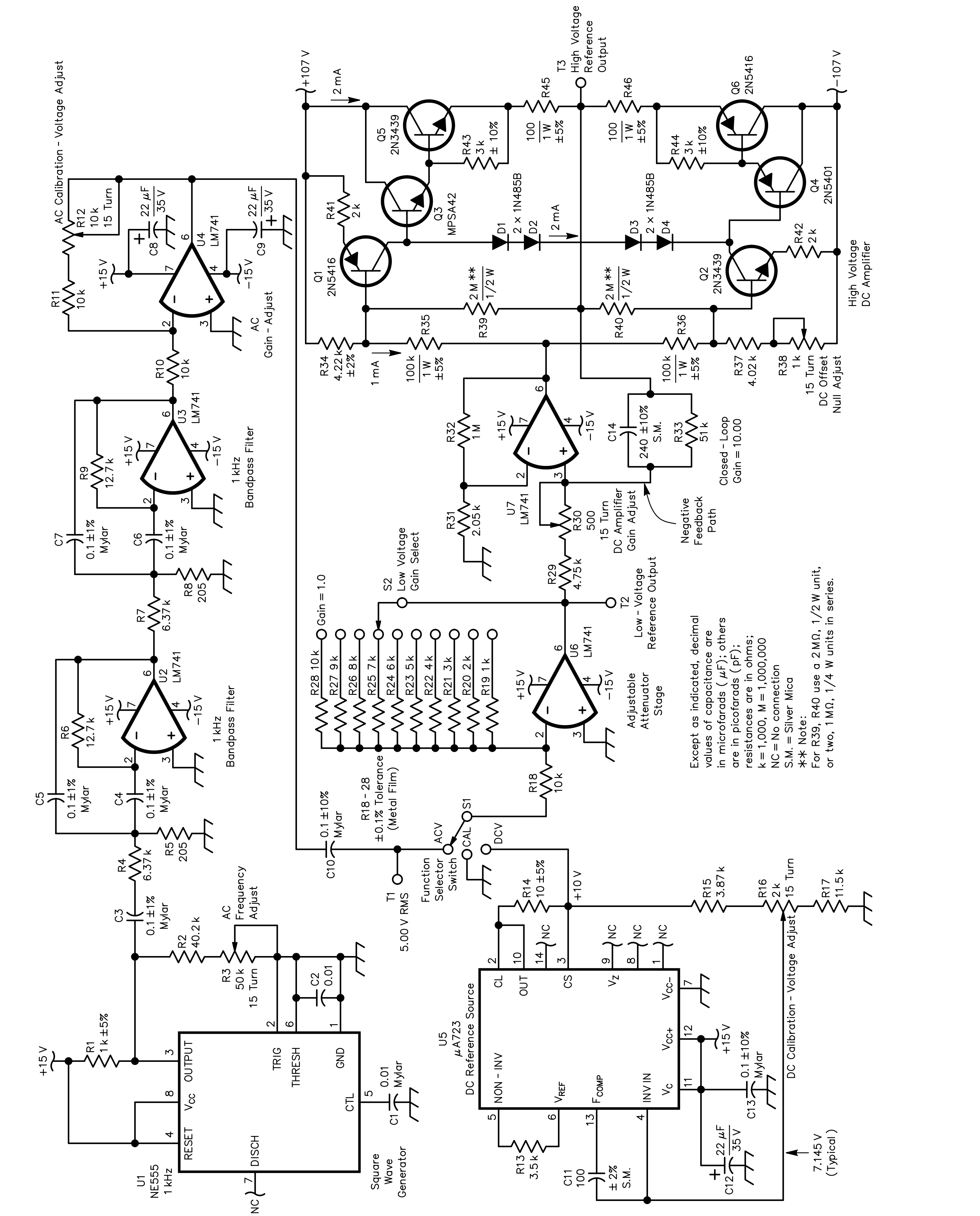
The precision 10-V dc reference source, U5, is a μA723 integrated-circuit device in a 14-pin dual in-line (DIP) package. This device is advertised as having 0.02% input regulation and 0.03% load regulation,[1] which is more than sufficient to meet the requirements of the calibrator circuit. The output reference voltage at pin 3 of this device is adjusted by means of potentiometer R16. Accurate adjustment of the source is discussed below.

AC Reference Source

For the 5-V RMS precision reference source, I initially tried to use a Wien-bridge oscillator circuit incorporating an operational amplifier and a miniature lamp to control the amplitude of the output voltage. Unfortunately, the amplitude of this reference voltage was not sufficiently stable for use as a precision source.

I then tried an NE555 timer in an 8-pin DIP package to generate a 1-kHz square wave. This waveform would then be filtered to provide the sine-wave reference. This timer package was wired to insure a 50%-duty-cycle square waveform, as described in an article by F. N. Cicchiello.[2] The amplitude stability of the generated reference voltage depends upon the stability of the associated power-supply voltage and R-C components. Amplitude stability also depends on tracking of the NE555 frequency with the filter's mid-band frequency. This combination of an NE555 and band-pass filters provides a very stable ac reference source.

[1]Notes appear on page 54.



AC Calibration - Voltage Adjust
U1 NE555 1 kHz
Square Wave Generator
AC Frequency Adjust
1 kHz Bandpass Filter
U2 LM741
U3 LM741
U4 LM741
AC Gain - Adjust
T1 5.00 V RMS
Function Selector Switch
ACV
CAL
DCV
S1
R18 - 28 ±0.1% Tolerance (Metal Film)
S2 Low Voltage Gain Select
Gain = 1.0
Adjustable Attenuator Stage
U6 LM741
T2 Low - Voltage Reference Output
U7 LM741
15 Turn DC Amplifier Gain Adjust
Negative Feedback Path
Closed - Loop Gain = 10.00
T3 High Voltage Reference Output
+107 V
−107 V
High Voltage DC Amplifier
15 Turn DC Offset Null Adjust
U5 μA723 DC Reference Source
DC Calibration - Voltage Adjust
7.145 V (Typical)
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
NC = No connection
S.M. = Silver Mica
** Note:
For R39, R40 use a 2 MΩ, 1/2 W unit, or two, 1 MΩ, 1/4 W units in series.

Fig 1

Fig 1—Voltage calibrator circuit schematic diagram. Unless otherwise indicated, resistors are ±1%-tolerance, 1/4-W metal-oxide or metal film units. Equivalent parts may be substituted. Suppliers are indicated in parentheses. Those parts with no supplier listed are available from Concord Components, 111 Broadway, Concord, NE 68728; tel 800-871-1749, fax 402-584-2615; concord@nntc.net; http://www.conc.com. Jameco Electronics, 1355 Shoreway Rd, Belmont, CA 94002; tel 800-831-4242, fax 800-237-6948; info@jameco.com; http://www.jameco.com. Newark Electronics, 4801 N Ravenswood Ave, Chicago, IL 60640-4496; tel 800-463-9275, 773-784-5100; http://www.newark.com/. Circuit Specialists, 220 S Country Club Dr #2, Mesa, AZ 85210; tel 800-528-1417, 602-464-2485 (8 AM-6 PM MST), fax 602-464-5824; jr@cir.com; http://www.cir.com/.

C1—0.01 µF, ±20% mylar
C2—0.01 µF, ±5% mylar
C3-C7—0.10 µF, ±1% mylar
C10, C13—0.10 µF, ±10% mylar
C11—100 pF, ±2% silver mica
C8, C9, C12—22 µF, 35 V electrolytic
C14—240 pF, ±10%, silver mica
D1-D4—1N485B rectifier, 180 V, 0.2 A
Q1, Q6—2N5416 PNP transistor (350 V, 1 A, Newark)
Q2, Q5—2N3439 NPN transistor (350 V, 1 A, Newark)
Q3—MPSA42 NPN transistor (300 V, 0.5 A)
Q4—2N5401 PNP transistor (150 V, 0.6 A)
R1—1 kΩ, ±5%, 1 W
R12—10 kΩ, 15-turn cermet potentiometer (Jameco)
R14—10 Ω, ±5%
R16—2 kΩ, 15-turn cermet potentiometer (Jameco)
R18, R28—10 kΩ, ±0.1% 1/4 W metal film (see text regarding tolerances for R18 through R28)
R19—1 kΩ, ±0.1% 1/4 W metal film
R20—2 kΩ, ±0.1% 1/4 W metal film
R21—3 kΩ, ±0.1% 1/4 W metal film
R22—4 kΩ, ±0.1% 1/4 W metal film
R23—5 kΩ, ±0.1% 1/4 W metal film
R24—6 kΩ, ±0.1% 1/4 W metal film
R25—7 kΩ, ±0.1% 1/4 W metal film
R26—8 kΩ, ±0.1% 1/4 W metal film
R27—9 kΩ, ±0.1% 1/4 W metal film
R30—500 Ω, 15-turn cermet potentiometer (Jameco)
R34—4.22 kΩ, ±2%
R35, R36—100 kΩ, ±5% 1 W
R38—1 kΩ, 15-turn cermet potentiometer (Jameco)
R39, R40—2 MΩ, ±1% 1/2 W
R43, R44—3 kΩ, ±10%
R45, R46—100 Ω, ±5% 1 W
R47-R50—75 kΩ, ±5% 1 W
U1—NE555 timer
U2-U4, U6,U7—LM741 op-amp
U5—µA723 voltage regulator
SW1—4p3t rotary switch (Circuit Specialists)
SW2—1p12t rotary switch (Circuit Specialists)

The 1-kHz square wave from pin 3 of U1 is fed into two tandem-connected, 1-kHz bandpass filters, U2 and U3. The design of these filters is also discussed in *The ARRL Handbook*.[3] Two identical filter sections were chosen because a single section would have required an excessively large pole-Q value to provide the filtering needed. Pole-Q values exceeding five would result in greater sensitivity of the filter's transfer function to variations in R-C component values and would result in an unacceptably large spread in these values. For this simple filter configuration, pole Q is equal to the ratio of the mid-band frequency (1 kHz) to the 3-dB bandwidth. A pole Q of four was chosen for each filter section and was found to result in a good sine wave at pin 6 of U4.

As noted in the ARRL article,[4] capacitors having good Q—such as those with polystyrene, mica or mylar dielectrics—are suitable for use in these filters. Do not use low-Q ceramic capacitors. The resistors must be stable units such as those using metal-oxide or metal-film technology. Do not use carbon-composition resistors. Both R and C component tolerances should be ±1% or better. Components with greater tolerances can be measured and series-parallel connected to achieve the correct values. If you do not have suitably accurate test equipment to measure R and C, it should be available at local high school or college physics laboratories. To compensate for any shift in the mid-band frequency of the bandpass filters caused by aging of R-C components, R3 of the U1 oscillator circuit can be adjusted periodically.

The dc high-voltage amplifier circuit includes operational amplifier U7 as well as transistors Q1 through Q6. This amplifier is similar to a circuit described in one of the Burr-Brown handbooks.[5] Current drain from the dual-polarity 107-V power supply is less than 6 mA. The Darlington-connected output stage, Q3 through Q6, is biased by diodes D1 through D4 to operate at a quiescent current of 2 mA. This eliminates crossover distortion of the output waveform.

To minimize drift in the quiescent dc output voltage at the T3 terminal, the circuit was designed to avoid heat build-up in transistors Q1 through Q6. For the same reason, a relatively large value of loop gain is employed. The closed-loop gain is 10.00 and is equal to the ratio of R33 to the sum of R29 and R30.

Step Attenuator

Potentiometer R38 is provided so that the quiescent dc voltage at terminal T3 can be adjusted to zero. The U6 stage allows the input reference voltage to be multiplied by a factor of 0.1 through 1.0 in 10 equal steps. Thus, the dc voltage appearing at terminal T2 can range from 1 V through 10 V in 1-V steps, depending on the setting of switch SW2. At the same time, the dc voltage appearing at terminal T3 is 10 times that at terminal T2. Pad R18-R28 to realize 0.1% resistor tolerance.

Power Supplies

To power the ac reference source, I use a dual-polarity, 15-V supply that happened to be on hand. Any IC regulator, such as the µA723, will serve this purpose. These IC regulators are discussed in *The ARRL Handbook*.[6]

The dual-polarity, high-voltage pow-er supply (shown in Fig 2) is of conventional design. It is adjusted to provide the proper output voltage by means of the 100-kΩ potentiometers R53 and R54.

Construction

I built the circuitry of Fig 1 on a vector board using wire-wrap sockets and push-in terminals. The R and C components were mounted on the "push-in" ends of these terminals. Any convenient component mounting and wiring methods could be employed, but do not "daisy-chain" circuit ground paths. I ran a separate ground lead from each stage to a common point near the low-voltage power supply. If one were to build this circuit on a PC board, I would recommend minimum 0.1-inch spacing between each high-voltage land and all other lands, to prevent the possibility of land-to-land voltage breakdown.

Calibrating the Calibrator

The accuracy of this source can be no better than that of the instrument used to calibrate it. To meet stated accuracy goals, adjustments on the source must be carried out with a digital voltmeter having accuracy no worse than 0.1% over the specified output-voltage range. Builders may need to relax accuracy goals to be consistent with whatever equipment is available. Community college physics labs are a good place to look for the temporary use of highly accurate voltmeters.

After a 10-minute warmup, the order of circuit calibration adjustment is as follows:

1. Set SW1 to the CAL position.

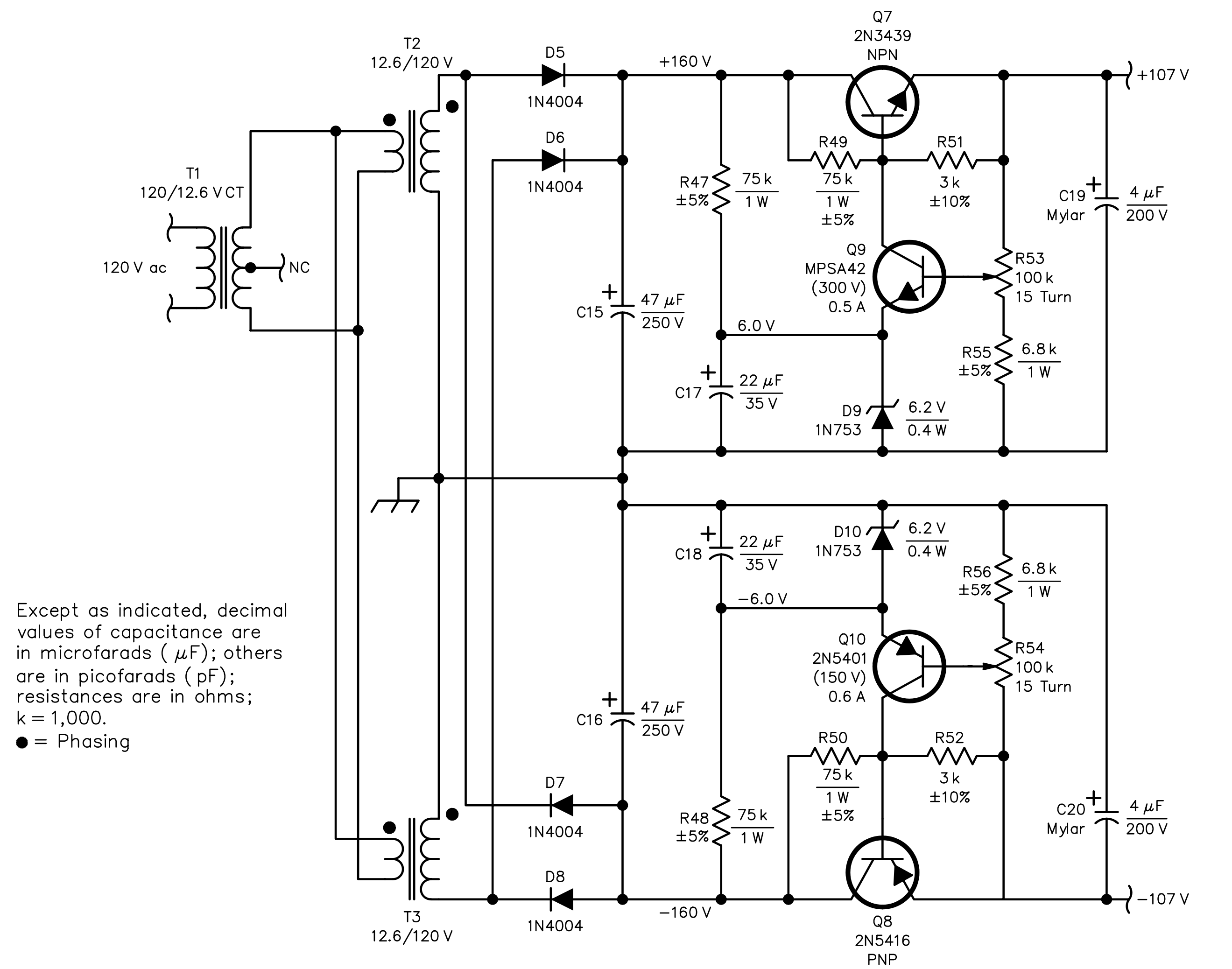


Fig 2—Dual-polarity, high-voltage power-supply schematic diagram. Unless otherwise indicated, resistors are ±1%-tolerance, 1/4-W metal-oxide or metal film units. Equivalent parts may be substituted. Suppliers are indicated in parentheses. Those parts with no supplier listed are available from Concord Components. Supplier contact information appears in the caption for Fig 1.

C15, C16—47 µF, 250 V electrolytic
C17, C18—22 µF, 35 V electrolytic
C19, C20—4 µF, 200 V mylar
D5-D8—1N4004 rectifier, 400 V, 1 A
D9, D10—1N753A, 6.2 V, 0.4 W Zener diode
Q7—2N3439 NPN transistor (350 V, 1 A, Newark)
Q8—2N5416 PNP transistor (350 V, 1A, Newark)
Q9—MPSA42 NPN transistor (300 V, 0.5 A)
Q10—2N5401 PNP transistor (150 V, 0.6 A)
R51, R52—3 kΩ, ±10%
R53, R54—100 kΩ, 3/4 W, 15-turn cermet potentiometer (Jameco)
R55, R56—6.8 kΩ, ±5% 1 W
T1—transformer, 120-V:12.6 V center tapped 1.2-A secondary (RadioShack 273-1352)
T2, T3—transformer 120:12.6 V, 0.45 A secondary (RadioShack 273-1365)

2. Adjust R38 to produce 0.00 V at terminal T3.

3. Set SW1 to the **DCV** position.

4. Set SW2 to the **10-K** (unity gain) position of the U6 stage.

5. Adjust R16 to produce 10.00 V at terminal T2.

6. Adjust R30 to produce 100.00 V at terminal T3.

7. Set SW1 to the **ACV** position.

8. Adjust R3 to maximize ac voltage at terminal T1.

9. Adjust R12 to produce 5.00 V RMS at terminal T1.

After these adjustments have been made, 5.00 V RMS will appear at terminal T2, and 50.00 V RMS will appear at terminal T3. The calibrator is now ready for use.

Notes

[1] *Voltage Regulator Data Book*, Texas Instruments, 1983, pp 2-153, 2-157.

[2] F. N. Cicchiello, "IC Timer Circuit Yields 50% Duty Cycle," Geometric Data Corp, Wayne, Pennsylvania. The publisher and publication date of this paper are unknown.

[3] *The ARRL Handbook* (Newington, CT: ARRL, 1990 edition) p 28-4.

[4] *ibid*, p 28-5.

[5] J. G. Grame, *Applications of Operational Amplifiers: Third-Generation Techniques*, (New York: McGraw-Hill, 1973), p 47.

[6] *The ARRL Handbook*, 1990 edition, p 26-4.

RF

By Zack Lau, W1VT

Proving the Conjugate Matching Power Theorem

This theorem has provoked quite a bit of controversy in amateur circles in recent years. I think it might be useful for advanced amateurs to see a mathematical proof. A good proof often adds to one's understanding of concepts.

Let's start by defining our system: The power source is represented by a Thevenin equivalent, which is a source voltage and a series source impedance, represented by Vs and $R \pm jX$, respectively (see Figure 1). Radio books use j to represent the imaginary operator because the usual imaginary operator, "*i,*" is already used to represent current. Math books use "i" to represent the imaginary operator; that is, $i^2 = -1$. All of these are constants, invariant with time. Now we seek to determine the optimum load impedance, Z_l (that subscript is a lower-case "L"—*Ed*), which is composed of a resistor R_l and series impedance X_l. By optimum, we mean that the power lost in R_l is maximized.

Since we have a series circuit, the power lost, P_l, is most easily calculated in terms of $|i|^2 \bullet R_l$. Since R_l is defined as a resistor, we just need to know the magnitude of the current through it. Thus, we can calculate $|i|^2$ as $(i) \bullet (\bar{i})$, where $\bar{i}$ is the complex conjugate of i.

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$$i = \frac{V_s}{R + jX + jX_l + R_l} \qquad \text{(Eq 1)}$$

$$\bar{i} = \frac{V_s}{R - jX - jX_l + R_l} \qquad \text{(Eq 2)}$$

$$\begin{aligned} i \bullet \bar{i} &= \frac{V_s^{\,2}}{\left(\left(\left(R+R_l\right)+j\left(X+X_l\right)\right)\left(\left(R+R_l\right)-j\left(X+X_l\right)\right)\right)} \\ &= \frac{V_s^{\,2}}{\left(R+R_l\right)^2-\left(j^2\right)\left(X+X_l\right)^2} \\ &= \frac{V_s^{\,2}}{\left(R+R_l\right)^2+\left(X+X_l\right)^2} \end{aligned} \qquad \text{(Eq 3)}$$

Taking the complex conjugate results in the sum of squares and cancellation of the imaginary cross product.

Thus, the power lost, P_l, is

$$P_l = \frac{V_s^{\,2} \bullet R_l}{\left(\left(R+R_l\right)^2+\left(X+X_l\right)^2\right)} \qquad \text{(Eq 4)}$$

Now, we wish to see how to maximize P_l by optimizing Z_l. Since Z_l is actually $R_l + jX_l$, we need to optimize P_l with respect to R_l and X_l. Calculus was invented for just this purpose. For example, if you were given an equation of the distance traveled from a point by a car, you could determine its maximum distance by just looking at the times the speed drops to zero, as well as looking at its starting and stopping times. This is quite useful—an infinite number of values is reduced to just a few points on a graph that can be easily calculated. Similarly, we can use calculus to determine the key points of complex electronic equations, to better understand how circuits work.

To do this, we differentiate P_l with respect to R_l and X_l. Remembering that if the equation is of the form

$$\frac{f(x)}{g(x)} \qquad \text{(Eq 5)}$$

the derivative is

$$\frac{f'(x)g(x) - g'(x)f(x)}{\left(g(x)\right)^2} \qquad \text{(Eq 6)}$$

One of the tricks to doing math is to simplify the equations as much as possible. I like to look for terms that cancel or become zero. Thus, it makes a lot of sense to do the differentiation with respect to X_l first. This way, $f(x)$ is a constant, which means that $f'(x) = 0$.

$$f(x) = V_s^2 \bullet R_l \qquad \text{(Eq 7)}$$

$$g(x) = \left(R + R_l\right)^2 + \left(X + X_l\right)^2 \qquad \text{(Eq 8)}$$

$$f'(x) = 0 \qquad \text{(Eq 9)}$$

$$g'(x) = 2 \bullet \left(X + X_l\right) \qquad \text{(Eq 10)}$$

Thus,

$$\frac{dP_l}{dX_l} = \frac{\left(0 \bullet g(x) - 2 \bullet \left(X + X_l\right)\left(V_s^2 \bullet R_l\right)\right)}{\left(\left(R + R_l\right)^2 + \left(X + X_l\right)^2\right)^2} \qquad \text{(Eq 11)}$$

It isn't necessary to expand $g(x)$—we are just going to cross it out.

$$\text{Setting } \frac{dP_l}{dX_l} = 0, \quad V_s^2 \bullet R_l \bullet 2\left(X + X_l\right) = 0$$

$$X_l = -X \qquad \text{(Eq 12)}$$

Keep in mind that $(R + R_l)^2$ cannot equal zero, or we would have division by zero. Very bad mistakes can occur when you divide by zero. Thus, R_l is not equal to $-R$.

Alternatively, we can look at the power equation with R_l set to an arbitrary constant. Since the numerator is fixed, we can maximize the power by minimizing the denominator. This occurs when $X_l + X$ is set equal to zero. Thus, $X_l = -X$. This approach is not as rigorous, but a knowledge of calculus isn't required. We can use $X + X_l = 0$ to simplify the equation,

$$P_l = \frac{V_s^2 \bullet R_l}{\left(R + R_l\right)^2 + \left(X + X_l\right)^2} \qquad \text{(Eq 13)}$$

$$P_l = \frac{V_s^2 \bullet R_l}{\left(R + R_l\right)^2} \qquad \text{(Eq 14)}$$

We can now take a partial derivation with respect to R_l.

$$f(x) = V_s^2 \bullet R_l \qquad \text{(Eq 15)}$$

$$g(x) = \left(R + R_l\right)^2 \qquad \text{(Eq 16)}$$

$$f'(x) = V_s^2 \qquad \text{(Eq 17)}$$

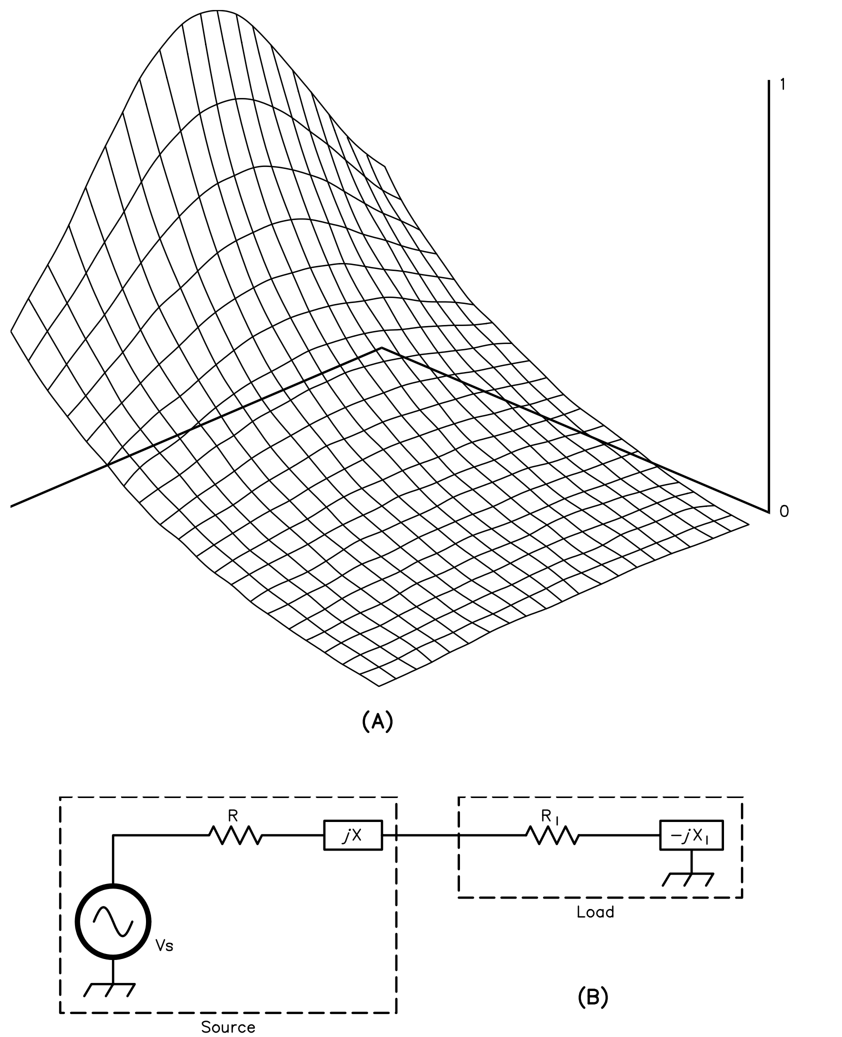

Fig 1—At A, a graph of power lost (P_l) as a function of R and X. See Table 1 for the conditions associated with this graph. At B, a diagram of the source and load circuits.

Table 1—Conditions associated with Fig 1

$R_l = 1$	$R_n = 0.1n$
$X_l = 1$	$X_m = 0.1m$
$N = 20$	$f(R, X) = \dfrac{R_l}{\left(R + R_l\right)^2 - \left(X + X_l\right)^2}$
$n = 0 \cdots N$	
$m = 0 \cdots N$	$PL_{(m,n)} = f\left(R_n, X_m\right)$

$$g'(x) = 2(R+R_1) \qquad \text{(Eq 18)}$$

$$\frac{dP_1}{dR_1} = \frac{\left(V_s^2 \bullet (R+R_1)^2 - 2(R+R_1)V_s^2 \bullet R_1\right)}{\left((R+R_1)^2\right)^2} \qquad \text{(Eq 19)}$$

Setting $dP_1 / dR_1 = 0$ and noting that $R + R_1$ cannot equal zero (lest we divide by zero):

$$V_s^2 \bullet (R+R_1)^2 = 2(R+R_1)V_s^2 \bullet R_1 \qquad \text{(Eq 20)}$$

$$R^2 + 2RR_1 + R_1^2 - 2RR_1 - 2R_1^2 = 0 \qquad \text{(Eq 21)}$$

$$\mathrm{R}^2 - \mathrm{R}_1^2 = 0 \qquad \text{(Eq 22)}$$

Factoring, $(R + R_1)(R - R_1) = 0$

Remember that we previously noted that R is not equal to $-R_1$, to avoid division by zero. Thus, the only solution is $R_1 = R$.

Since the real parts of the source and load impedance are equal, the source and load lose the same amount of power, setting the system efficiency at 50%.

Now to be very rigorous, you could do second differentiation, to determine whether the point we found is a maximum, a minimum or an inflection point. Alternatively, we know from practical experience or calculations that either endpoint—$Z_1 = 0$ or $Z_1 = \infty$ results in zero power transfer to the load. Also, power increases as the real part increases from zero or decreases from infinity. Thus, it is quite obvious from practical experience that the single point obtained is the desired maximum. Even if you do perform the extra math, looking at equations from this viewpoint is always a good idea. Remember that the equations are intended to simulate reality.

Thus, we have proven that if the source impedance is $R + jX$, the load impedance that results in maximum power transfer is $R - jX$. Thus, the optimum load impedance is the complex conjugate of the source impedance.

So why the controversy? I think it results from the misapplication of the theorem to modern transmitters. Most amateurs want to maximize their output power, so they look to the conjugate-matching theorem for guidance. However, modern transmitters are designed to work into specific load impedances, usually 50 Ω, rather than to be conjugately matched. While you may get more power with a conjugate match, you may overstress the final amplifier parts or create excessive distortion. Thus, even though I use a meter that automatically subtracts the reverse power from the forward power to give actual power measurements, I still adjust my Transmatch for minimum SWR, rather than maximum actual power.[1] An SWR meter measures how closely you match a desired impedance with a single number, which makes it quite convenient as a tuning indicator.

Many thanks to Kevin Schmidt, W9CF, who looked at an early draft of this manuscript and provided useful comments.

Why are SWR Meters affected by Power Level?

Contrary to what some believe, the impedance of the source has little to do with what an SWR meter reads. An SWR meter reads reflections that return from the load, not the source. A reflection from the source would be indistinguishable from "forward" power generated by the source.

The greatest source of erroneous SWR readings is nonlinearity of diode detectors. Consider an oversimplified model of an ideal –32-dB directional coupler with diodes having an exact 0.3 V conduction voltage. (See Fig 2.) The –32-dB coupling factor means that the forward power detector samples 32 dB less signal than is applied to the coupler. For instance, a 100-W signal will result in 50 dBm – 32dB = +18 dBm of signal, or 63 mW applied to the forward power detector. On the other hand, the 0.3-V voltage drop means that $(0.3\ \mathrm{V})^2/(2 \times 50\ \Omega)$ must be generated before the diode will conduct. This is 0.9 mW or –0.5 dBm. Thus, the reverse power can be has high as 0.9 mW before *any* reflected power will be indicated. At that level, the actual SWR would be

$$SWR = \left(\frac{1 + \left(\frac{P_R}{P_F}\right)^{0.5}}{1 - \left(\frac{P_R}{P_F}\right)^{0.5}} \right) = \left(\frac{1 + \left(\frac{0.9\ \mathrm{mW}}{63\ \mathrm{mW}}\right)^{0.5}}{1 - \left(\frac{0.9\ \mathrm{mW}}{63\ \mathrm{mW}}\right)^{0.5}} \right) = 1.27 \qquad \text{(Eq 23)}$$

where P_F is the forward power and P_R is the reflected power. Thus, the meter reads a 1.0 SWR when the SWR may actually be as high as 1.27.

If the power is boosted to 1500 W, the forward-power detector sees 61.8 dBm – 32 dB, or 29.8 dBm, or 955 mW.

The reverse power detector sees 0.9 mW × 1500 W / 100 W = 13.5 mW. This is 0.821 V (RMS) across a 50-Ω load, or 1.16 V (peak). Subtracting the 0.3 V, one gets an apparent reflected power of 7.4 mW.

Similarly, the apparent forward power is 897 mW, due to the diode drop.

$$SWR = \left(\frac{1 + \left(\frac{7.4\ \mathrm{mW}}{897\ \mathrm{mW}}\right)^{0.5}}{1 - \left(\frac{7.4\ \mathrm{mW}}{897\ \mathrm{mW}}\right)^{0.5}} \right) = 1.20 \qquad \text{(Eq 24)}$$

Thus, the SWR may appear to jump from 1.0 to 1.20 when an amplifier boosts the power from 100 W to 1500 W.

The situation becomes worse when running QRP. At the 5-W level, the forward power available from the directional coupler is just 37 – 32 = +5 dBm, or 3 mW. Since the reverse power meter reads zero for signals up to a threshold of 0.9 mW, the SWR could get as high as

$$\frac{1 + \left(\frac{0.9}{3}\right)^{0.5}}{1 - \left(\frac{0.9}{3}\right)^{0.5}} = 3.4 \qquad \text{(Eq 25)}$$

[1]"The Tandem Match—An Accurate Directional Wattmeter, John Grebenkemper, KI6WX, January 1987 *QST* and recent ARRL Handbooks p.22.34 to 22.40 of the *2000 Handbook*.

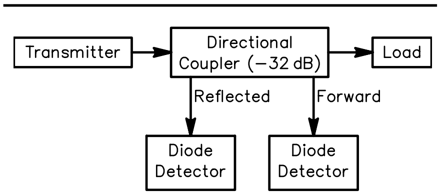

Fig 2—Block diagram of a typical SWR meter.

Table 2—Summary of SWR calculations

Transmit Power (W)	*Forward Power at Detector*	*Reverse Power at Detector*	*Apparent Forward power*	*Apparent Reflected Power*	*Apparent SWR*
100	63 mW	0.9 mW	49 mW	0 mW	1.00
1500	955 mW	13.5 mW	897 mW	7.4 mW	1.20

Thus, the SWR could be 3.4:1 before the reflected power begins to register, even if the forward scale is properly calibrated. This is summarized in Table 2.

A solution is to draw separate SWR scales for 5- and 100-W levels that compensate for the diode nonlinearity. Alternately, one can dispense with the SWR scale and mentally compute the degree of mismatch as the ratio of the forward and reverse powers. Bird Corporation suggests this approach with their analog power meters.

Another approach is to use a better detector—one that corrects for the nonlinearity of the diode detector. Roy Lewallen, W7EL, published an excellent example of this approach in "A Simple and Accurate QRP Directional Wattmeter," (*QST*, Feb 1990, pp 19-23). Roy discovered that not only is the diode drop a problem, but that best results require calibration with ac, rather than dc, reference signals. His article also gives a more precise model of diodes that allows more-accurate calculations of circuit performance.

Trepanning Large Holes

At the 1999 Microwave Update, Ed Krome, K9EK, talked about one of the most hazardous modern Amateur Radio activities: fly cutting a large hole in a sheet of metal. A sharp metal bit is held in a fixture that allows it to rotate in a large circle, cutting a grove in piece of metal firmly attached to a drill press. This can be quite dangerous if the bit gets stuck—the metal can easily be sent flying through the air. Often, the cutting operation has enough vibration to loosen any clamps, allowing the metal to fly free. A poor solution is to carefully smooth the edges and round the corners of the sheet metal. This will prevent the projectile from acting like a sharp knife.

A better solution is to use a milling machine and a rotary table to trepan the holes. Instead of whirling around a big cutter, the sheet metal is slowly rotated with the rotary table. The milling machine just cuts a little 1/16 or 7/64 inch hole as the plate is rotated. Granted, a little four-inch Sherline rotary table won't allow exact duplication of a big 432 MHz plate line, but will often do a fine job at higher frequencies. It may also be practical to make precision plate collets that can be screwed onto low-frequency plate lines. I suspect that little four-inch plates are much easier to work with, particularly if you are attempting to preheat the work with a hot plate. The milling machine is also useful for empirically optimizing your work. It isn't too difficult to mill off the used solder and enlarge a hole, if you want to try a slightly larger hole.

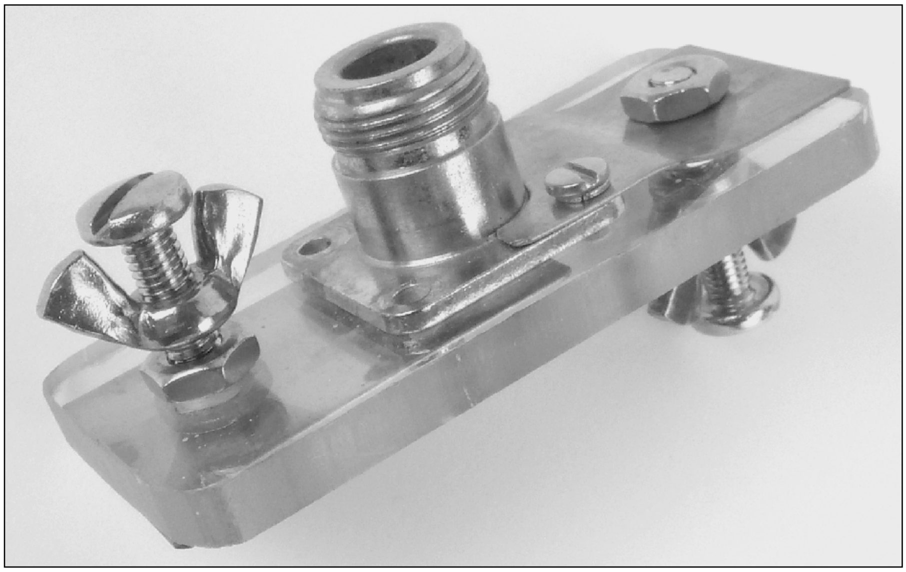

Fig 3—W3IRZ's technique to capture wing nuts at connections.

More on Center Insulator for Dipoles

Zack—That's an okay idea on the wing nuts for binding posts, but here's one better: For field use, you should prevent the loss of the wing nuts by reversing the screw 180°. By doing this, the wing nut will contact the screw head when it unscrews about 3/8 of an inch. The screw is held onto the material (plastic in this case) with a nut on both sides. I picked up a box of #8-32 wing nuts years ago and have been using them in this manner for some time.—*Mike Branca, W3IRZ, Conyers, Georgia;* **w3irz@att.net**

Next Issue in QEX/Communications Quarterly

Bar-Giora Goldberg gives us a review of popular frequency-synthesis techniques and his vision for the future of that field. He discusses tradeoffs among power consumption, spectral purity, cost and complexity. Interesting information regarding loop noise shaping and all-digital fractional-N methods is presented. Giora also concentrates on characterizing signals as narrow-band noise and points out a prime goal of frequency synthesis: cleaning up the signal.

As we've seen recently, interest in homebrewing high-dynamic-range receivers has persuaded some designers to rethink the use of broadband front ends. Bill Sabin, W0IYH, brings us his design of a bank of narrow band-pass filters that may be used ahead of receivers to limit input bandwidth, thus assuaging second-order IMD and other problems. CAD is used extensively in design and analysis of the filters. To achieve close linearity, Bill paid careful attention to the inductor-core material and flux density.

Charles Kitchin, N1TEV, updates us on some new super-regenerative receiver techniques. He draws on his experience with these circuits over the last few years to bat down some false—but common—notions about them. Charles details traditional drawbacks of the "super-regen" and tells how he and others have developed ways to avoid them. Designs are discussed for VHF and NBFM.

R. P. Haviland, W4MB's series on quad antennas continues. Part 3 covers the care and feeding of multi-element designs.

Letters to the Editor

Practical HF Digital Voice (May/June 2000)

Hello Doug,

I am most impressed with your *QEX/Communications Quarterly* readership. I have had far more well informed e-mails in response to our digital-voice article than I have had from any of the readers of the other magazines that have carried a similar article.—*Charles Brain, G4GUO,* **chbrain@dircon.co.uk**

Dear Doug,

Thanks! It looks like this [digital voice] is finally on the radar screen! Perhaps we could start a small, private chat group to collect background materials, formulate topics, and allocate time and other resources.—*Bob Bernacki, N9LQV, PO Box 3188, Bloomington, IN 47402;* **bob@bernacki.com**

PTC: Perceptual Transform Coding ... (May/June 2000)

Hi there Doug,

I read your article with great interest. Many decades ago, I recall Bell Labs running voice tests with what I think they called meaningless mono symbols.

My reason for this note, however, is my interest in HF CW and that of my pal Dave, KO6RS, in moonbounce. Both arenas involve very weak-signal CW. We have talked about how the human ear/brain works under these conditions, but we have come to no conclusions. We would be interested in any insights you might have.

Referring to your Fig 1, it suggests that there would be a penalty of a couple decibels or so reading CW around the normal listening tone of 500-750 Hz. Is this so? I have tried listening to CW at higher notes but find it tiring. Where does fatigue become a factor?

Your Fig 2 is intriguing, as it suggests that an external filter bandwidth of around 100 Hz for 500 Hz tones would match that of the "internal" filter. Is this a valid interpretation of your graph? I have found 100 Hz to be the best all-round audio filtering bandwidth, but this might just be coincidence.

I would like to better understand the implications of "masking" when the wanted signal tends to be "masked" in noise. One can play with the CW tone and the spectrum of the audible noise, and it seems to me that noise concentrated at lower pitches than that of the CW tone gives better results. I would like to know more about the ideal shape of the CW envelope and resultant sidebands relative to readability. It seems that sometimes a slightly rough note is better to copy than a pure tone.

So please, in your quest for optimizing signals using auditory psychophysics, don't leave us CW folks out in the cold!—*73, Ron Skelton, W6WO,* **rskelton@surfnetusa.com**

Hi Ron,

Thanks for the note. Although I don't know how to evaluate the effects of fatigue on the criterion-level problem, I know the higher-frequency tone inherently sounds louder; thus it is more tiresome. Other effects may be in play.

One way to interpret Fig 2 is to relate it to polyphonic sounds that fall partly inside the critical bandwidth, partly without. Phase changes of one or more tones (or noise) inside the critical bandwidth tend to be masked.

I've been preparing more material on the criterion-level business because this is of obvious interest to weak-signal fans: Have I received the signal, or not? This discipline of detection theory is a large one in and of itself; I'm not that familiar with it yet. Try checking out some of the references I gave.

In speaking with moonbounce fans at Dayton this year, I learned about the problem of "Doppler smear," caused by the motions of the Moon and Earth, which tends to broaden a CW signal and make it sound "raspy." It's not yet clear to me whether DSP adaptive-equalizer techniques can be applied to this problem to eliminate the smear, but I'd say there is hope.—73, Doug Smith, KF6DX, **kf6dx@ arrl.org**

Doug,

I like the article in May/June issue and look forward to the rest of the parts. I have one complaint, however: You defined acoustic intensity as the "physical measure of sound pressure level." Acoustic intensity is the product of the pressure and the particle velocity, or in a simple free field, it is equal to the pressure squared divided by the specific acoustic impedance of air (415 *rayls*). It is a vector quantity (where sound pressure is scalar) that describes the energy flow associated with a sound (W/m^2).

In a reverberant environment, the pressure is typically high but the intenity is essentially undefined; modern intensity systems measure near zero. Also, when the term *sound pressure* is used, it normally means the units are in *pascals*. If the term *sound pressure level* is used, the unit is decibels referenced to 0.000020 *pascals.—73, Bob Hand, W8WQS,* **bobhand@kuntrynet.com**

Hi Bob,

Thanks for that clarification.—73, Doug

A High-Performance Homebrew Transceiver: Part 3 (Nov/Dec 1999) and A High-Performance AGC System for Homebrew Transceivers (Oct 1995)

Doug,

Here are a few corrections to Part 3 of my 1999 article series: At the upper right of Fig 3, the trimpot label "IF SHIFT" should be deleted. At the lower left, the USB IFS transistor should be inverted (emitter should connect to the diode and collector to the trimpot). At the right in Fig 8, there should be a blocking capacitor between the output of BPF1 and the first buffer. At the upper left of Fig 10, the network surrounding C1 (near the σ40 terminal) should be labeled BPF1.

In my 1995 article, at lower right of Fig 5, the lower meter should be labeled "M2—SQUELCH/CLIPPING" and M1 should be labeled "SIGNAL/ALC." Unfortunately, the M1-M2 meter labels in the 1999 article series are reversed from those in this earlier article. There are several corrections to Fig 7: To the right of U202's output, the unlabeled terminal should be labeled "Va—to S-Meter Section," and the unlabeled resistor should be 100 k. At the lower left, U204 should have a dot at the output—this specifies an LM339 (see the caption of Fig 6). At the upper left, the IN and OUT terminal labels of the 7815 are reversed. I hope these have caused no inconvenience for readers. Thanks to Paul Bardell, N2OTD, for pointing out the errors in the AGC article.—*Mark Mandelkern, K5AM,* **k5am@zianet.com**; **k5am@arrl.net**.

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CLOVER-2000 sends data over standard HF SSB radio channels at 3000 bps. Including error correction and ARQ overhead, CLOVER-2000 will deliver up to 2000 error-corrected bits/second over an HF radio link. The -50dB occupied bandwidth of the transmitted signal is only 2000 Hz. CLOVER-2000 may be used with *any* good-quality HF SSB transmitter and reciever.

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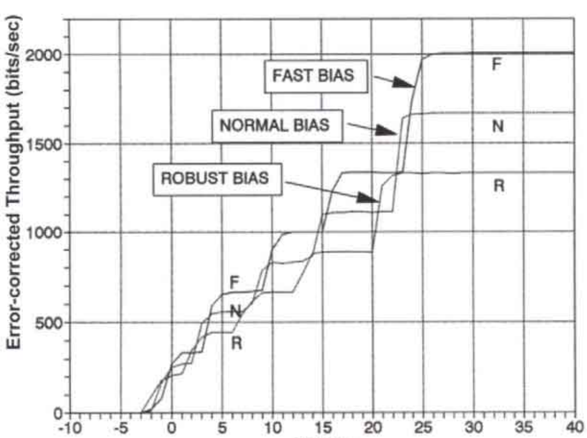

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The CLOVER ARQ protocol allows real-time adaptive transmission of data *in both directions* on the ARQ link *without the need for special "over"* commands. CLOVER is also data transparent; the modem will send any 8-bit stream provided without modification and without requiring special command sequences. CLOVER-2000 is a low-cost and high throughput solution to your HF data communications problems. It is the "waveform of choice" for thousands of users throughout the world.

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SS-12	10	12	1⅜ x 6 x 9	3.4
SS-18	15	18	1⅜ x 6 x 9	3.6
SS-25	20	25	2⅞ x 7 x 9⅜	4.2
SS-30	25	30	3¾ x 7 x 9⅝	5.0


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SS-30M*	25	30	3¾ x 7 x 9⅝	5.0


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SRM-30	25	30	3½ x 19 x 9⅝	7.0

WITH SEPARATE VOLT & AMP METERS

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25M	20	25	3½ x 19 x 9⅝	6.5
SRM-30M	25	30	3½ x 19 x 9⅝	7.0


MODEL SRM-30M-2

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MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
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SRM-30-2	25	30	3½ x 19 x 9⅝	11.0

WITH SEPARATE VOLT & AMP METERS

MODEL	CONT. (Amps)	ICS	SIZE (inches)	Wt.(lbs.)
SRM-25M-2	20	25	3½ x 19 x 9⅝	10.5
SRM-30M-2	25	30	3½ x 19 x 9⅝	11.0


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CUSTOM POWER SUPPLIES FOR RADIOS BELOW	*NEW SWITCHING MODELS*
EF JOHNSON AVENGER GX-MC41	SS-10GX, SS-12GX
EF JOHNSON AVENGER GX-MC42	SS-18GX
EF JOHNSON GT-ML81	SS-12EFJ
EF JOHNSON GT-ML83	SS-18EFJ
EF JOHNSON 9800 SERIES	SS-10-EFJ-98, SS-12-EFJ-98, SS-18-EFJ-98
GE MARC SERIES	SS-12MC
GE MONOGRAM SERIES & MAXON SM-4000 SERIES	SS-10MG, SS-12MG
ICOM IC-F11020 & IC-F2020	SS-101F, SS-121F
KENWOOD TK760, 762, 840, 860, 940, 941	SS-10TK
KENWOOD TK760H, 762H	SS-12TK OR SS-18TK
MOTOROLA LOW POWER SM50, SM120, & GTX	SS-10SM/GTX
MOTOROLA HIGH POWER SM50, SM120, & GTX	SS-10SM/GTX, SS-12SM/GTX, SS-18SM/GTX
MOTOROLA RADIUS & GM 300	SS-10RA
MOTOROLA RADIUS & GM 300	SS-12RA
MOTOROLA RADIUS & GM 300	SS-18RA
UNIDEN SMH1525, SMU4525	SS-10SMU, SS-12SMU, SS-18SMU
VERTEX — FTL-1011, FT-1011, FT-2011, FT-7011	SS-10V, SS-12V, SS-18V

*ICS – Intermittent Communication Service