$5

QEX

INCLUDING: COMMUNICATIONS QUARTERLY

Forum for Communications Experimenters

November/December 2001

OK1DNZ's Two-Meter Reflective SP4T PIN-Diode Switch

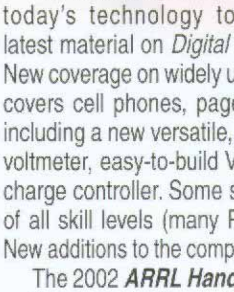
ARRL The national association for AMATEUR RADIO
225 Main Street
Newington, CT USA 06111-1494

THE ARRL
HANDBOOK
2002

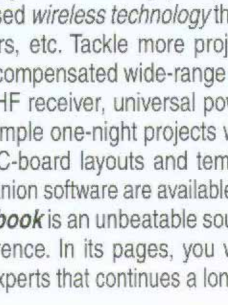
The 79th edition of ***The ARRL Handbook*** puts today's technology to work. Discover the latest material on *Digital Signal Processing* (DSP). New coverage on widely used *wireless technology* that covers cell phones, pagers, etc. Tackle more projects, including a new versatile, compensated wide-range modular RF voltmeter, easy-to-build VHF receiver, universal power supply, and solar charge controller. Some simple one-night projects will appeal to builders of all skill levels (many PC-board layouts and templates are included). New additions to the companion software are available for free downloading.

The 2002 ***ARRL Handbook*** is an unbeatable source as an electronics and communications reference. In its pages, you will find the collective wisdom of generations of experts that continues a long tradition of technical excellence. Over twelve hundred pages and thirty chapters help translate theory into practice. Radio amateurs have always turned to ***The ARRL Handbook*** for the most current, practical information on the Amateur Radio state of the art. **Who** else uses ***The ARRL Handbook***?— Electrical/Electronic Technicians and Engineers; Instructors and Students; Government and Private Researchers. **Where** else will you find ***The ARRL Handbook*** used?— Engineering Laboratories; Product Development Teams (Electronics); Research Institutions; School and University labs.

For more information, ordering, or to locate an ARRL dealer, call **1-888-277-5289** (US). Or visit us at **www.arrl.org/shop**.



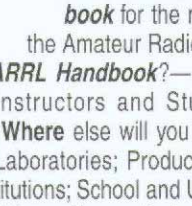
The ARRL Handbook for Radio Amateurs



Softcover.
ARRL Order No. 1891 .... **$34.95***
*plus shipping $7 US (UPS)/$9 International

Hardcover.
ARRL Order No. 1913 .... **$49.95***
*plus shipping $8 US (UPS)/$10 International

The ARRL Handbook CD for Radio Amateurs



Version 6.0—for *Windows* and Macintosh**
View, **Search** and **Print** from the entire 2002 edition book! CD-ROM.
ARRL Order No. 1905 .... **$39.95***
*plus shipping $5 US (UPS)/$7 International
**Some supplementary software utilities included, for *Windows* and DOS only.

ARRL *The national association for AMATEUR RADIO*

225 Main Street, Newington, CT 06111-1494
tel: 860-594-0355 fax: 860-594-0303

In the US call our toll-free number 1-888-277-5289 8 AM-8 PM Eastern time Mon.-Fri.
www.arrl.org/shop

QEX 11/2001

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: Send address changes to: QEX, 225 Main St, Newington, CT 06111-1494
Issue No 209

Mark J. Wilson, K1RO
Publisher

Doug Smith, KF6DX
Editor

Robert Schetgen, KU7G
Managing Editor

Lori Weinberg
Assistant Editor

Peter Bertini, K1ZJH
Zack Lau, W1VT
Contributing Editors

Production Department

Steve Ford, WB8IMY
Publications Manager

Michelle Bloom, WB1ENT
Production Supervisor

Sue Fagan
Graphic Design Supervisor

David Pingree, N1NAS
Technical Illustrator

Joe Shea
Production Assistant

Advertising Information Contact:

John Bee, N1GNV, *Advertising Manager*
860-594-0207 direct
860-594-0200 ARRL
860-594-0259 fax

Circulation Department

Debra Jahnke, *Manager*
Kathy Capodicasa, N1GZO, *Deputy Manager*
Cathy Stepina, *QEX Circulation*

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
Telex: 650215-5052 MCI
Fax: 860-594-0259 (24 hour direct line)
e-mail: **qex@arrl.org**

Subscription rate for 6 issues:

In the US: ARRL Member $24, nonmember $36;

US by First Class Mail: ARRL member $37, nonmember $49;

Elsewhere by Surface Mail (4-8 week delivery): ARRL member $31, nonmember $43;

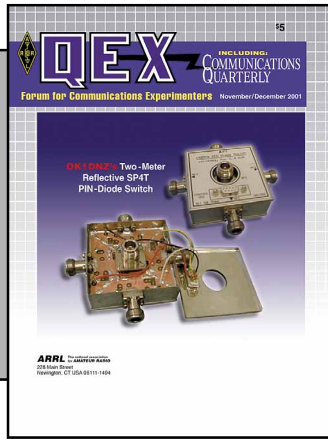
Canada by Airmail: ARRL member $40, nonmember $52;

Elsewhere by Airmail: ARRL member $59, nonmember $71.

Members are asked to include their membership control number or a label from their *QST* wrapper when applying.

In order to ensure prompt delivery, we ask that you periodically check the address information on your mailing label. If you find any inaccuracies, please contact the Circulation Department immediately. Thank you for your assistance.

Copyright ©2001 by the American Radio Relay League Inc. For permission to quote or reprint material from *QEX* or any ARRL publication, send a written request including the issue date (or book title), article, page numbers and a description of where you intend to use the reprinted material. Send the request to the office of the Publications Manager (**permission@arrl.org**)



About the Cover

Pavel Zanek's solid-state VHF switch. Details begin on page 3.

Features

Columns

Nov/Dec 2001 *QEX* Advertising Index

THE AMERICAN RADIO RELAY LEAGUE

The American Radio Relay League, Inc, is a noncommercial association of radio amateurs, organized for the promotion of interests in Amateur Radio communication and experimentation, for the establishment of networks to provide communications in the event of disasters or other emergencies, for the advancement of radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

ARRL is an incorporated association without capital stock chartered under the laws of the state of Connecticut, and is an exempt organization under Section 501(c)(3) of the Internal Revenue Code of 1986. Its affairs are governed by a Board of Directors, whose voting members are elected every two years by the general membership. The officers are elected or appointed by the Directors. The League is noncommercial, and no one who could gain financially from the shaping of its affairs is eligible for membership on its Board.

"Of, by, and for the radio amateur, "ARRL numbers within its ranks the vast majority of active amateurs in the nation and has a proud history of achievement as the standard-bearer in amateur affairs.

A bona fide interest in Amateur Radio is the only essential qualification of membership; an Amateur Radio license is not a prerequisite, although full voting membership is granted only to licensed amateurs in the US.

Membership inquiries and general correspondence should be addressed to the administrative headquarters at 225 Main Street, Newington, CT 06111 USA.

Telephone: 860-594-0200
Telex: 650215-5052 MCI
MCIMAIL (electronic mail system) ID: 215-5052
FAX: 860-594-0259 (24-hour direct line)

Officers

President: JIM D. HAYNIE, W5JBP
3226 Newcastle Dr, Dallas, TX 75220-1640

Executive Vice President: DAVID SUMNER, K1ZZ

The purpose of *QEX* is to:

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

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

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

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the Editor or the League. While we strive to ensure all material is technically correct, authors are expected to defend their own assertions. Products mentioned are included for your information only; no endorsement is implied. Readers are cautioned to verify the availability of products before sending money to vendors.

Empirically Speaking

Recent terrible events have left us shocked and appalled. We are glad to see unity around the world in the face of tragedy, though; we are exploiting our best opportunities to prevent further disaster. Again, Amateur Radio has shown itself to be a global treasure—not just because of our tactical value, but also because we represent millions who care. While we do not always make the headlines, our presence is surely felt by those in need. Our ability to promote international goodwill is especially important in this time of tribulation.

We are in a situation that threatens to undermine our most-cherished beliefs of freedom and justice. Beyond the immediate threat lie two other stealthy and even greater opponents: ignorance and apathy. To win the war against those sleepy behemoths is to defeat all other enemies at once. We pledge to battle ignorance but individual motivation must come from within. Get on the air and think about what you are saying.

This issue marks the 20th anniversary of *QEX*. Editing this magazine for the last three and a half years has been a delightful experience. We have enjoyed outstanding support from you, and if that continues, *QEX* will continue to grow. I am proud to be part of a forum that holds the free exchange of ideas and information to be so important. It is an honor for me, but you are the real heroes. Were it not for you, we would not exist. You are sustaining a wonderful legacy of technical excellence: One need not dig too deeply to prove that. Just parse some past issues to find many outstanding examples of innovation.

The 20th is the "china" anniversary. I cannot resist drawing an analogy between that beautiful, yet fragile material and what we do here. The beauty of the technical arguments presented on these pages resides not only in their novelty, but also in that they are presented here at all. We engage in a fragile business since we often deal with intangibles that defy definition. Your writing and comments about new things are what others seek for motivation in the technical arena. So, how did we get started and what keeps us going?

All healthy human beings seem to have an inherent curiosity. Further, we like to share with others what we have learned. In his final statement to the Boy Scouts of the world, Robert Baden-Powell left us with a significant plea: "Try and leave the world a better place than before you have found it." It is a good slogan that represents a worthy goal: that of continual improvement. Our philosophy at *QEX* has been that many little improvements amount to significance in the end. Big improvements are good, too; but persistence is what pays in the finish.

In This Issue

Pavel Zanek, OK1DNZ, brings us a two-meter antenna switch. Its unique features may be useful to you in many circumstances.

Lilburn Smith, W5KQJ, shows how to get contacts on an amateur band that largely is neglected. He introduces you to a bit of photonics, which field is a hot topic in communications these days. You say you want some bandwidth?

Mark Mandelkern, K5AM, delivers his promised segment about the HF "front ends" for his homebrew transceiver. This is the "icing on the cake" for homebrewers and contesters who want top-notch performance.

Steve Best, VE9SRB, concludes his tutorial on transmission-line mechanics. In it, he shows how impedance matching is achieved by wave superposition.

In January, Paul Wade, W1GHZ, presented some ideas on circular waveguide. Cal Tech's William Bridges, W6FA, adds a few observations of his own that you will want to consider. His analysis reveals several areas where we can benefit if we are careful.

In RF, Zack describes a simple 10-meter turnstile antenna for use with satellites. *73, Doug Smith, KF6DX;* **kf6dx@arrl.org**

A Two-Meter Reflective SP4T PIN-Diode Switch

Could you use a convenient electronic switch that selects one of four VHF inputs? If so, build one like this!

By Pavel Zanek, OK1DNZ

This reflective PIN-diode SP4T switch was designed for switching among four receive antennas to get very quickly the best RF signal on two meters. It is a high-performance unit with excellent RF parameters. We can use it in many applications, such as a controlled antenna multiplexer/demultiplexer, for laboratory tests and measurement and so forth. The design procedure and important, practical points in that process are illustrated. Based on this article, you may be able to build a similar switch according to your own requirements—perhaps with a different number of poles or in a new frequency range.

Slovenska 518
Chrudim
Czech Republic, 537 05
Zanek.pavel@worldonline.cz

Design Parameters

Goals for this design include:

- Characteristic impedance=50 Ω.
- Operating frequency range of 144-147 MHz
- Very fast switching, electronically controlled, unlimited lifetime
- Very low insertion loss (IL), typically 0.35 dB
- All RF ports very well matched, SWR less than 1.05:1
- High isolation, worst case 84.5 dB
- Low harmonic distortion, second harmonic = –58 dBc for input power of +29.3 dBm, –40 dBc for input power of +34.6 dBm. Higher harmonic products have better suppression.
- No damage for CW input power of +37 dBm (5 W) at one RF port, other ports connected to 50 Ω.
- Excellent output third-order intercept point, IP3 = +52.4 dBm
- SWR < 1.20 between 128-167 MHz at all RF ports
- IL less than 0.40 dB between 115 and 160 MHz, all directions
- Fully shielded construction
- Easy to produce, low-cost solution; minimized number of element values

All of the above-mentioned parameters were measured in the laboratory on the first sample of the SP4T switch, and they are all determined by the worst-case value measured in the range 144-147 MHz. All transmission and impedance parameters were measured with a Hewlett Packard HP-8714B vector network analyzer at an output level of 0 dBm. Two additional 10-dB pads for transmission measurement were used to avoid mismatch error when IL was measured. The HP-8714B was calibrated to correctly measure low SWRs. The analyzer was set to zero-span mode and an RF signal generator was used to measure and check isolation exactly.

Intercept point was computed for the following input data: input drive power levels, P= +23 dBm (for each tone), IMD_3 product = –59.5 dBc. These frequency sets were used to find the worst IP3: 144.0 and 144.1 MHz, 145.0 and 145.1 MHz and 146.0 and 146.1 MHz.

Optimizing the Switch for Two Meters

This article fully describes only the design and calculations for a SP4T switch. If you are interested in how PIN diodes work or how PIN diodes are usually used as switches, see Note 1.

A block diagram of the SP4T switch is shown in Fig 1. Table 1 describes the control of the switch. All control voltages are referenced to common ground. The schematic diagram is shown in Fig 2. The switch has been designed for these basic parameters inside the 2-meter band: IL < 0.5 dB, isolation > 80 dB and SWR < 1.15 using *Super Compact* software. Control current should not exceed 40 mA for every section of the SP4T switch. The absolute value of control current for an open or closed section must be the same. This same current can be measured for simple testing purposes—it can tell you whether the section is okay or not. I have chosen my favorite PIN diodes: MA4P1250 from M/A-COM Inc, USA.[2]

We obtain from a simple linear RF analysis that three PIN diodes are necessary (T structure of one switching section). Here are the results of the RF linear analysis for one section: The capacitance of the shunt diode must be less than 1.0 pF to satisfy the isolation (> 85 dB) requirement. Series resistance must be lower than 0.7 Ω to get the desired low IL of the complete switch. See below for a discussion of IL.

On the basis of these computed data and from Note 2, we can determine the bias conditions for the PIN diodes when they are on and off:

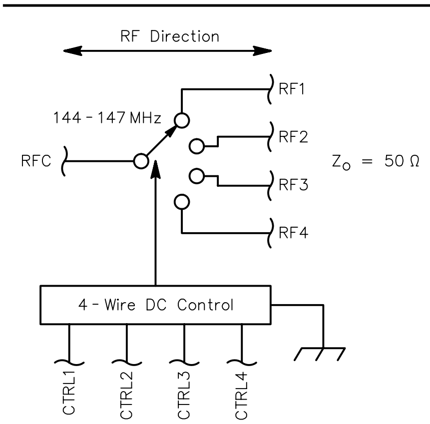
Forward current I_f = 17 mA
Series resistance R_s = 0.6 Ω
Reverse voltage V_r = 0.7 to 4.5 V
Capacitance C_r = 0.9 pF

Note: Series PIN diodes are open with I_f = 17mA and shunt PIN diodes short with $I_f = 2 \times 17$ = 34 mA. The 34-mA value results from the requirement that the control current be the same within one section. Reverse voltage V_r for the shunt diode is not the same as V_r for both series diodes. However, the capacitance of the active PIN diode doesn't depend upon V_r for frequencies above 50 MHz (see Note 1). It changes only the parallel resistance. This change is not so important to the isolation of one section.

[1]Notes appear on page 10.

Table 1—The control of SP4T switch. V_{pos} = +12.0 V, V_{neg} = –8.3 V. Control current $I_{pos} = -I_{neg}$ = 34 mA at every control pin CTRL1-CTRL4.

Connected ports	*CTRL1*	*CTRL2*	*CTRL3*	*CTRL4*
RFC - RF1	V_{pos}	V_{neg}	V_{neg}	V_{neg}
RFC - RF2	V_{neg}	V_{pos}	V_{neg}	V_{neg}
RFC - RF3	V_{neg}	V_{neg}	V_{pos}	V_{neg}
RFC - RF4	V_{neg}	V_{neg}	V_{neg}	V_{pos}



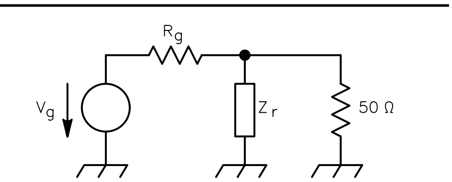


Fig 1—(A) Functional schematic of SP4T switch. (B) Model of circuit used for Table 2.

Theory of Operation

First, let's study only one section for RF open and closed conditions. The complete SP4T switch will be discussed later.

One Open Section

The principle of operation is very simple (see Fig 2). Let's assume that direction RF1-RFC (section 1) is open by applying positive voltage U_{pos} (+12 V) at control pin CTRL1. The other directions RF2-RFC (section 2), RF3-RFC (section 3) and RF4-RFC (section 4) must be closed by the negative voltage U_{neg} (–8.3 V) at control pins CTRL2, CTRL3 and CTRL4. Thus D10 and D12 are open by the same I_f = 17 mA. Their series resistances are low. PIN diode D11 is reverse-biased by the potential delivered by R10, R11, R12 and the V_f of D10 or D12. In that case, it is approximately +4.5 V and D11 now presents a parallel connection of approximately 30 kΩ and capacitance 0.9 pF. The actual IL of one section is only 0.13 dB (measured, 0.11 dB computed) when the other three sections are disconnected from the RFC port.

One Closed Section

Closed sections work in the following way: Let's study only closed section 2, direction RF2-RFC. The negative voltage V_{neg} opens shunt diode D21 and its resistance is low. Its forward current is two times higher (34mA), compared with the I_f of D10 or D12 (see the design criteria). A –0.79-V drop across D21 forces series diodes D20 and D22 to be reverse-biased. The parallel resistance of D20 or D22 is approximately 18 kΩ with a capacitance of 0.9 pF. So there is much isolation between the RF2 and RFC ports. Practically, this isolation is more limited by mechanical construction (RF bypassing of signal) than by RF parameters of the PIN diodes used. Thus, shielding partitions must be

Table 2—Insertion loss given by real parallel impedance Z_r

Z_r(Ω, X= 0Ω)	*IL (dB)*
1.000	0.21
2.000	0.11
5.000	0.04
7.000	0.03
10.000	0.02

added (see Figs 3, 6 and 14). The theoretical isolation of only one closed section is 88.8 dB.

Why are we saying this is a reflective SP4T switch? Well, closed port RF2 is not matched to 50 Ω, so total reflection occurs.

Complete SP4T Switch

Let's now study the complete SP4T switch for the situations described above. Each closed section 2, 3 and 4 has a parallel capacitance of about 3.3pF and a parallel resistance of about 4.2 kΩ (measured). The capacitance is that of open D22 (for example, section 2) and its small transformation by a short 50-Ω microstrip line between D22 and the RFC port. The capacitance of the center PC board point of the RFC port to ground has a value of approximately 4 pF. The capacitance overall at the RFC port is 3 × 3.3 + 4 = 13.9 pF. This is a large capacitance at 145 MHz, and it must by eliminated by using a parallel inductor, L1, to achieve resonance at 145 MHz. By Thomson's formula, we get value of L1 = 86.7 nH. Parallel connection of three resistances of 4.2 kΩ, in parallel with the resistance of coil L1, brings additional IL. The IL of the 50-Ω microstrip line (FR4 PC board of thickness 1.5 mm) for one section is only 0.01 dB. Several hundredths of a decibel of IL loss are caused by the finite parallel resistance of the RF chokes used (see below). All losses considered, IL = 0.35 dB.

What happens when compensating L1 is not used? The RFC port will have an SWR of 1.7 at 145 MHz. Thus, there will be additional loss from mismatch error, which increases overall loss to 0.62 dB. *One of the most important points is the use of properly designed RF chokes.* Each choke must achieve high isolation between the dc side and the RF side to yield a low IL. What's

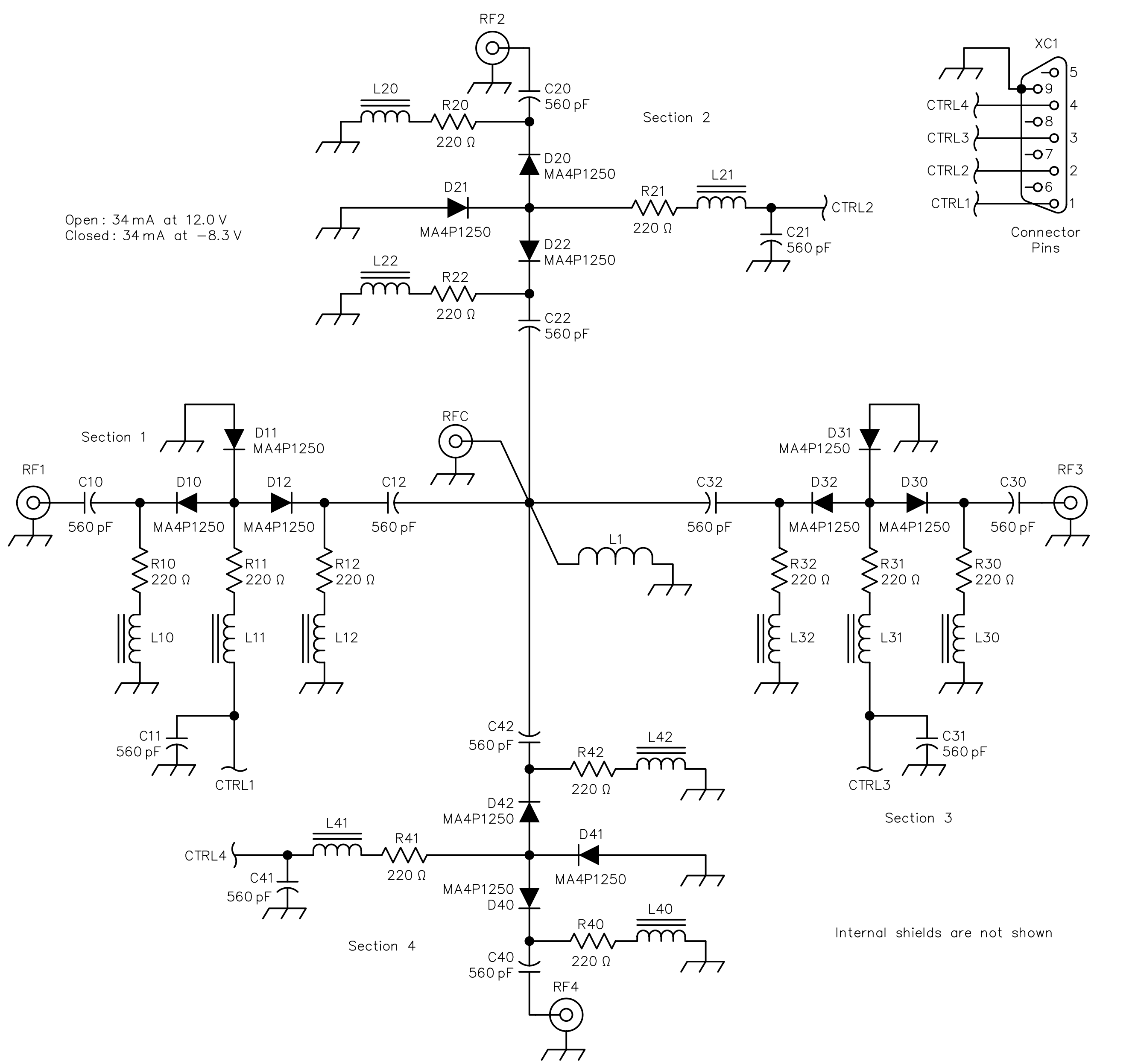


Fig 2—Schematic diagram of 145MHz PIN diode switch.

the necessary value of choke impedance Z_r? Let's assume a loss-less connection at the RF source (R_i = 50 Ω) and 50-Ω load. Table 2 shows the additional insertion loss caused by inserting an RF choke parallel to the 50-Ω load. You see we must have very high Z_r to achieve a low IL. From the circuit topology, we have the parallel connection of three RF chokes with series resistors (for example: L11 + R11 parallel to L10 + R10 parallel to L12 + R12) between RF and ground.

How do we get high isolation for this dc bias circuit simply? This requirement is met by parallel resonance of the RF chokes and a little higher Q factor. In this way, you get high real impedance and zero imaginary impedance at the resonance frequency.

The next inductor requirement is for very small external magnetic fields to avoid parasitic transmission between RF chokes (for example: between L10

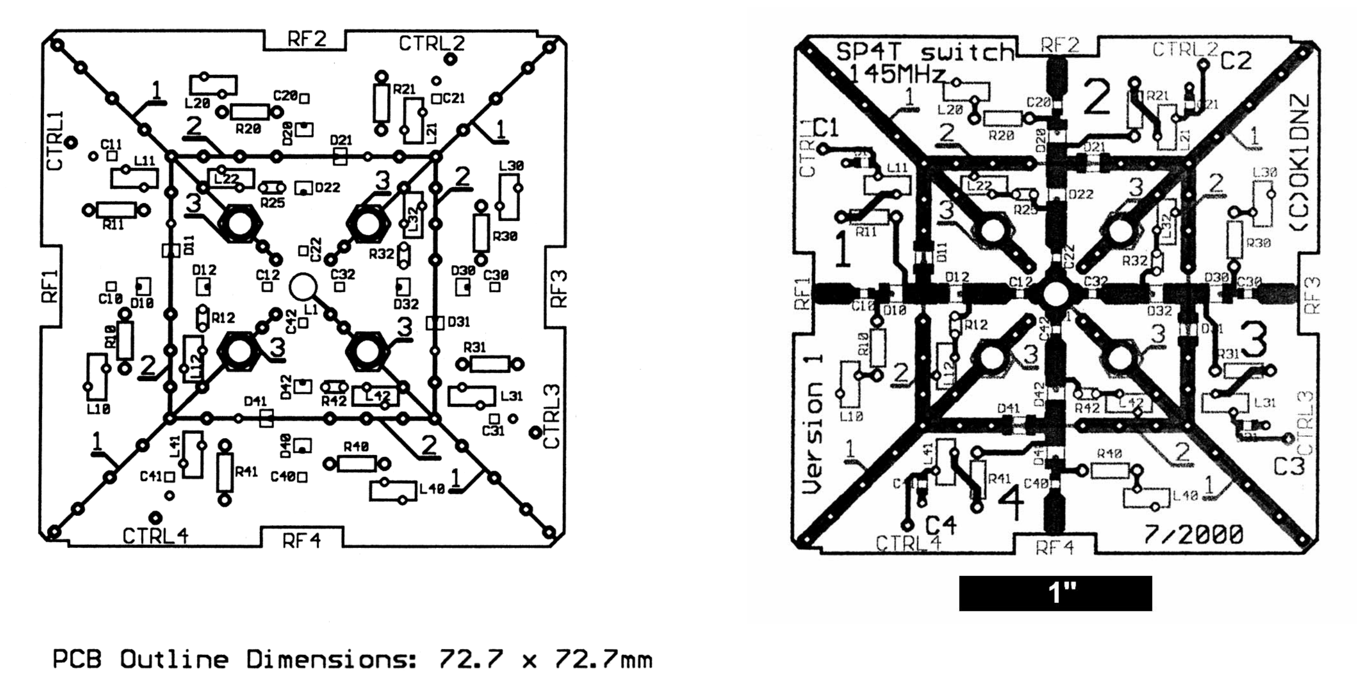


Fig 3—Two views of component layout, side A. At right shows traces as well as parts.

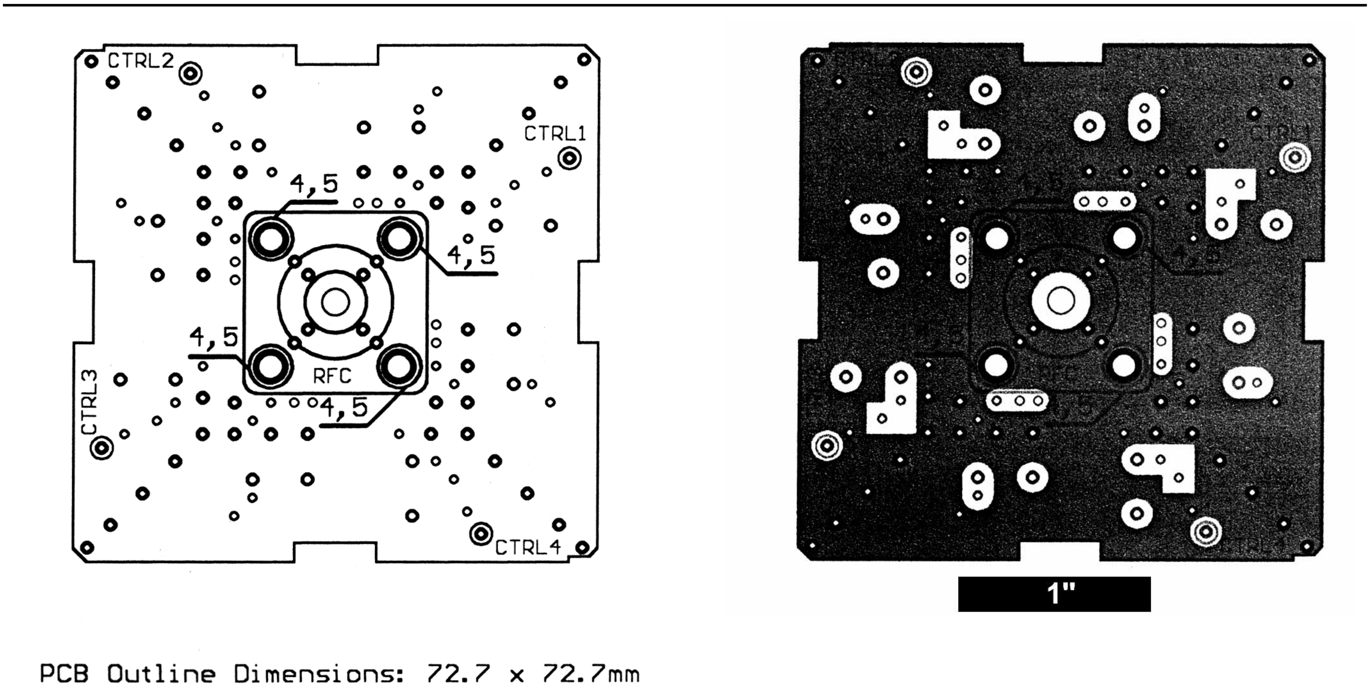


Fig 4—Two views of component layout, side B. At right shows the ground-plane foil.

and L12) when the section is closed. Toroid chokes with appropriate core materials must be used and designed based on vector impedance measurement. It's not so easy to mathematically predict the self-resonance frequency (SRF). All RF chokes used in SP4T switch are the same. Each parallel resistance is about 10 kΩ at 145.5 MHz.

Setting the DC Conditions

Symbol "x" below indexes the actual section of the switch: 1-4.

Calculating Resistor Values

Series PIN diodes Dx0 and Dx2 have the same I_f = 17 mA. Let's assume Rx0=Rx1=Rx2 (positive potential will be about 1/3 of $V_{pos} + V_f$ at the cathode of shunt diode Dx1, open section) and Dx0 and Dx2 have the same dc characteristics. We can write for Rx1:

$$Rx1 = \frac{\left(V_{pos} - V_f\right)}{3I_f} \quad \text{(Eq 1)}$$

where

V_f=forward voltage of diode Dx0 or Dx2 at I_f. In our case, 0.73 V

I_f=forward current through diode Dx0 or Dx2. In our case, 17 mA

V_{pos}=positive control voltage for open section (12.0 V)

From Eq 1, we obtain Rx1= 221 Ω. The power dissipation of Rx1 for the open section is given by:

$$P(Rx1)_{open} = Rx1(I_f)^2 \quad \text{(Eq 2)}$$

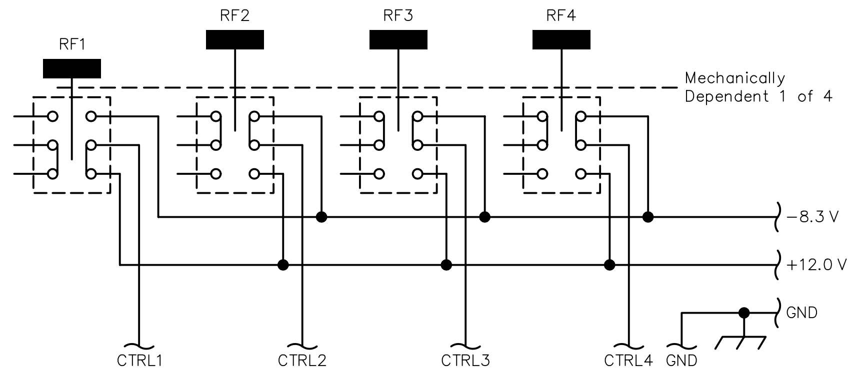


Fig 5—A simple four-button SP4T controller.

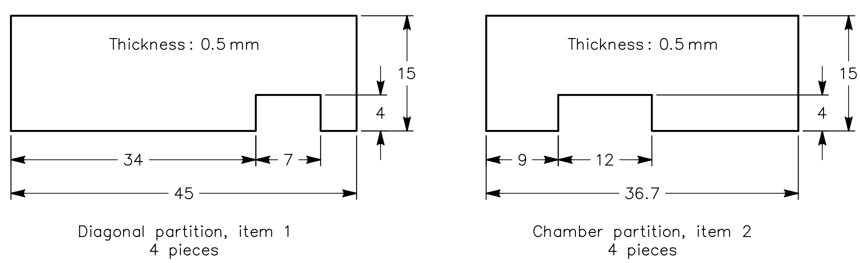


Fig 6—Construction details of the shielding partitions.

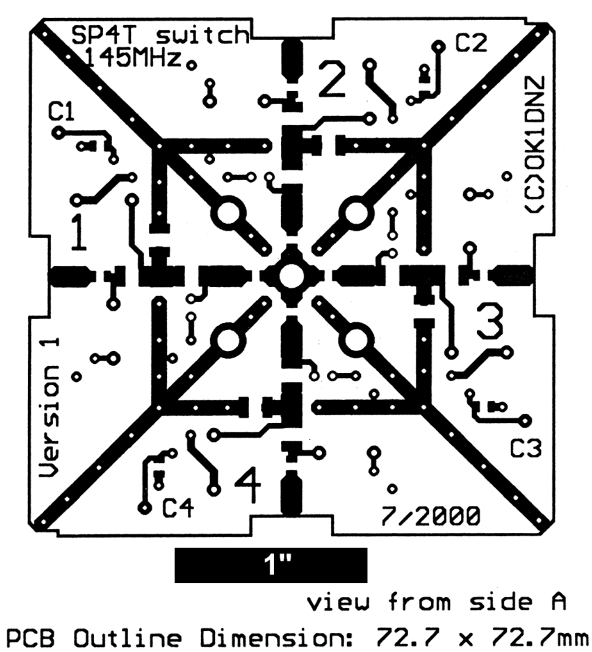


Fig 7—Etching pattern for the PC board, side A.

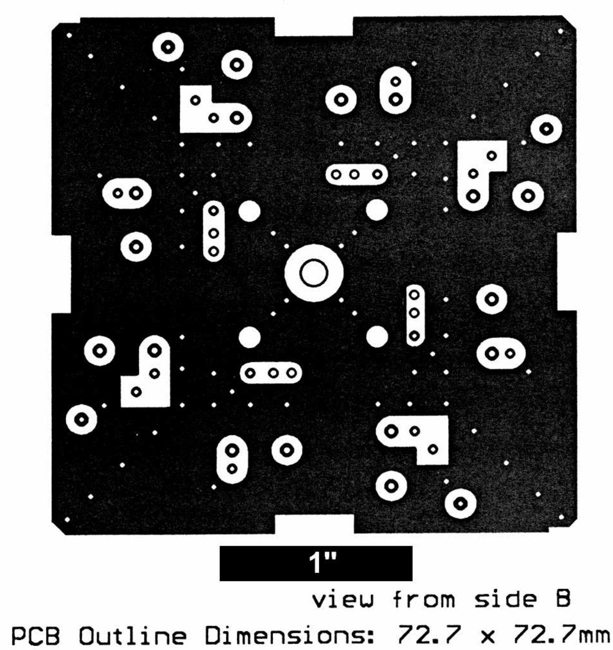


Fig 8—Etching pattern for the PC board, side B.

Calculating power dissipation, we get 63.6 mW.

Let's now find the closing V_{neg}. It will be derived from initial condition: $I_{neg} = -I_{pos}$ at pin CTRLx. It comes to:

$$V_{neg} = -(2Rx1)(I_f + V_f{}') \quad \text{(Eq 3)}$$

where

$V_f{}'$=forward voltage of diode Dx1 for forward current $I_f{}' = 2I_f = 34$ mA. In our case, it is 0.79 V.

The computed V_{neg} is –8.27 V. Power dissipated by Rx1 for the closed section is:

$$P(Rx1)_{closed} = \frac{\left(\left|V_{neg}\right| - V_f{}'\right)^2}{Rx1} \quad \text{(Eq 4)}$$

For Rx1 = 220 Ω, we get 254.3 mW from Eq 4. Power dissipated by Rx0 and Rx2 is lower than the computed cases above. At least a 0.6-W power resistor must be used for this application.

Construction

The switch is built into a cheap, tinned steel box (WBG 38 DONAU, 74×74×30 mm, thickness 0.5 mm). All RF ports use female N connectors. RF1-RF4 have N connectors with a smaller base (17.5×17.5mm). The N connector for the RFC port is a common type (base 25.4×25.4 mm). Control port XC1 has a D-subminiature 9-pin female connector. All components are placed on one side of the PC board, side A (see Figs 3 and 14). On the other side, side B, are placed the RFC N connector and wires, which connect XC1 to the PC board (see Figs 4 and 15). The PC board is double-sided with *vias*, which connect the ground of both PC-board sides. Here is a step-by-step procedure for making the entire switch assembly.

1. Solder all elements except L1 (see Fig 3).
2. Tin the ground area near live pin of RFC connector at the B side.
3. Fix RFC connector from side B using four screws (see Figs 4 and 13 and 15).
4. *Very important*: Solder the ground of RFC connector to ground of the PC board. Use 100-W soldering gun with a builder-modified tip to reach and solder as shown in Fig 13.
5. Mount all RF1-RF4 connectors to the box.
6. Insert the PC board into the box. Solder the four live pins of RF1-RF4 on side A. These pins must be shortened to 2 mm in length (see Fig 15).
7. Solder PC board (ground) to box along complete circumference. The ground of the RF1-RF4 connectors is connected to PC board ground by a thin copper strip.
8. Solder four diagonal partitions (Fig 6, item1 in Fig 3). The four nuts, item 3, for the RFC N-connector must be soldered too. Solder partitions to PC board and to the box.
9. Solder four "chamber" partitions (Fig 6, item 2 in Fig 3). Solder them to PC board and to diagonal partitions.
10. Drill a 24-mm-diameter hole at the center of upper cover for the RFC connector. Stick a label on the upper cover. Also, fix the XC1 connector there.
11. Place air inductor L1 (see Figs 3 and 14). Connect the Control XC1 connector using wires to control pins on the PC board, side B.
12. Clean all solder points.
13. Fix all ferrite chokes to PC board using silicone adhesive (RTV3140, DOW Corning). Paint both sides of PC board with a protective varnish like conformal coating 1-2577 (DOW Corning).
14. Make electrical adjustment (see below).
15. Place upper and lower cover. Solder them at several points.

Electrical Adjustment

Use a simple SP4T switch controller to check and adjust the RF performance. The simplest controller is shown in Fig 5. It's formed from four, interdependent switches (when one is pressed, the other three are disengaged). The adjustment is very simple.

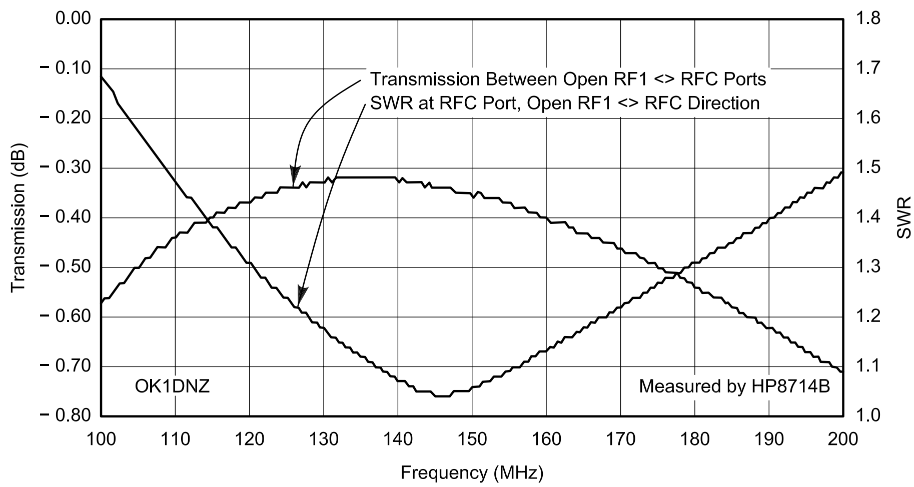


Fig 9—An Excel chart of the HP-8714B measurements shows insertion response and SWR for the closed path from RF1 to RFC.

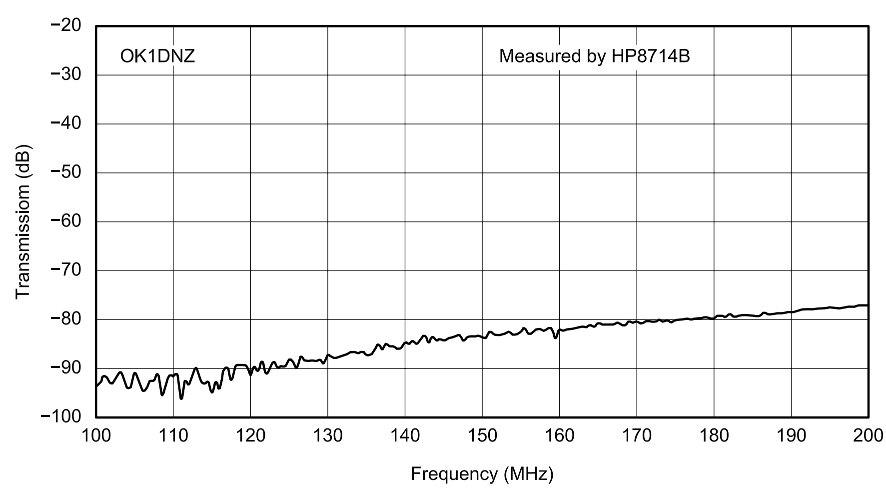


Fig 10—An Excel chart of the HP-8714B measurements shows insertion response for the open path from RF1 to RFC.

Connect a laboratory 50-Ω, male N-connector load on RF1 port. Activate RF1-RFC to be open.

Perform impedance calibration of network scalar/vector analyzer before measurement at frequency range 140-150 MHz. You must calibrate it in the measuring impedance plane.

Connect measuring cable of network analyzer to RFC port.

Adjust L1 to get the best SWR from 144 from 147 MHz.

Close both covers and check SWR again.

What happens if you don't perform these adjustments? You get a little worse SWR, typically 1.3:1; IL will be also a little higher, typically 0.40 dB. Nevertheless, you won't notice any difference in practical use.

Conclusion

This SP4T switch with a micro controller is used at our club radio station, OK1KCR, for all two-meter contests. Simply speaking, we are using four receive antennas pointed at desired directions. When in a very big pileup, we comfortably and very quickly pick up desired weak calling stations. This switch was designed only for indoor use. Only a compact, weatherproof housing is necessary for outdoor use.

Measured RF parameters are shown in Figs 9 and 10. They are very close to computed values. The front label of the

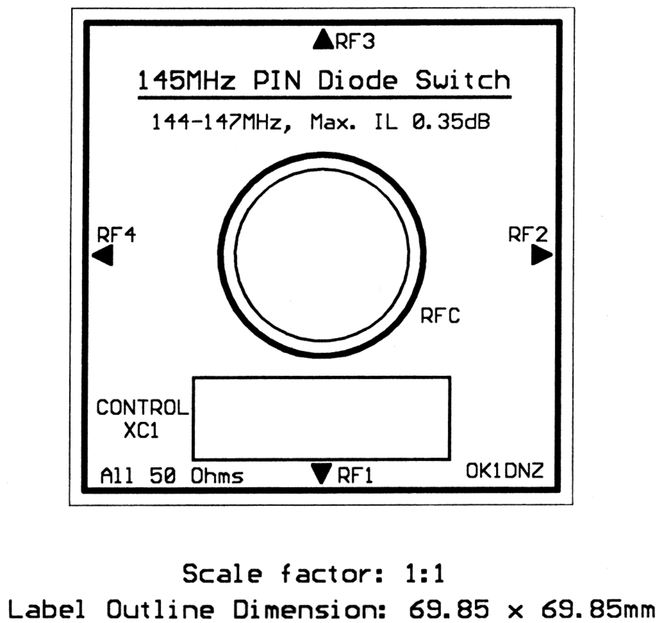


Fig 11—Drawing of the switch front label.

Fig 12—Front view of the SP4T switch.

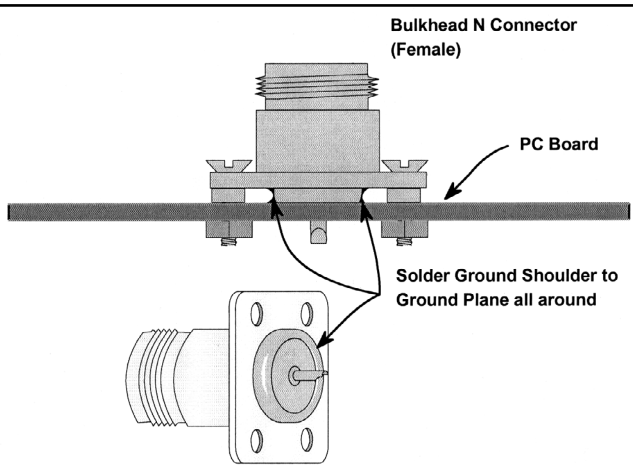


Fig 13—It is important that the shield of the incoming transmission line is maintained through to the component side of the circuit board. This is done by soldering the protruding ground shoulder on the back of the connector flange to the ground plane on all sides. To accomplish this requires a soldering tip capable of reaching between the connector flange and the circuit board. The prototype was constructed with a bent-wire extension to the tip of a 100-W soldering gun.

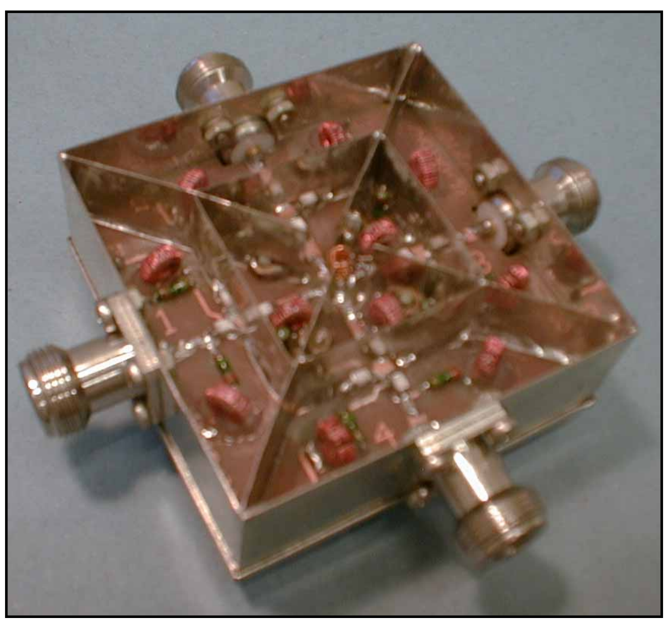


Fig 14—A view of the completed switch, side A.

Parts List

R10-R12, R20-R22, R30-R32, R40-R42—220 Ω, 0. 6 W metal-film resistor, high-frequency SMA0207, TK50ppm//C, ±1%, size 0207 SMA0207HF/MK1HF 50 220R 1% VISHAY-DRALORIC
C10-C12, C20-C22, C30-C32, C40-C42—560 pF 100 V ±5% SMD capacitor, size 0805, dielectric NP0, VJ 0805 A 561 J X B A T VISHAY-DRALORIC
L10-L12, L20-L22, L30-L32, L40-L42—145-MHz RF choke (2.03 µH @ 10 MHz): 31 turns of enameled copper wire, 0.20-mm diameter (AWG# 32) on Amidon toroid core: T25-10 (black). The winding is arranged uniformly around the whole core and coated with silicone rubber RTV3140, Dow Corning SRF = 145.5 MHz
L1—Air-core coil 7 turns of enameled copper wire, 0.40 mm diameter (AWG# 26) on 3-mm-diameter drill bit. Approximately 0.3 mm of space between turns.
D10-D12, D20-D22, D30-D32, D40-D42—MA4P1250 surface-mount PIN diode from M/A-COM Inc, USA (cathode indicated by a red dot)
RF1-4—Panel-mount female N connector, Flange jack, 53 K 403-200 A3 Rosenberger
RFC—Panel-mount female N connector, Flange jack, 53 K 401-200 A3 Rosenberger
XC1—D subminiature 9-pin female connector, Straight panel assembly, ITT Cannon
Item 1—4 diagonal partitions, tinned-steel (see Fig 6)
Item 2—4 chamber partitions, tinned-steel (see Fig 6)
Item 3—4 brass metric nuts, M3, DIN934 (see Fig 3)
Item 4—4 spring pads, 3.2, DIN6798A (see Fig 4)
Item 5—4 metric screws, M3, 10 mm long, DIN 7985 (see Fig 4)
PC board—FR4, double sided, 1.5 mm thick, $\varepsilon r = 4.3$, $tg\delta = 0.02$

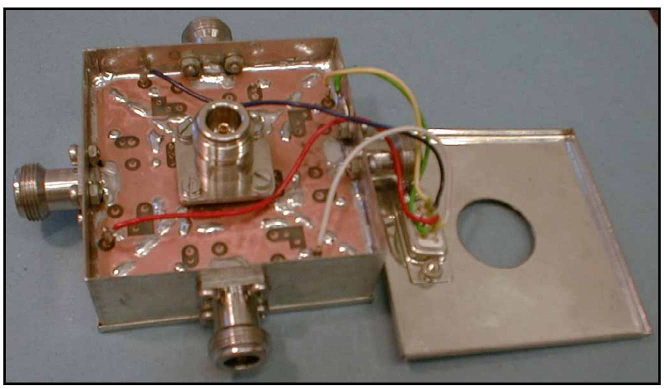


Fig 15—A view of the completed switch, side B.

SP4T switch is depicted in Fig 11. Photos (see Figs 12, 14 and 15) clearly show the complete SP4T switch. If you have any questions, please contact me by e-mail.

Notes

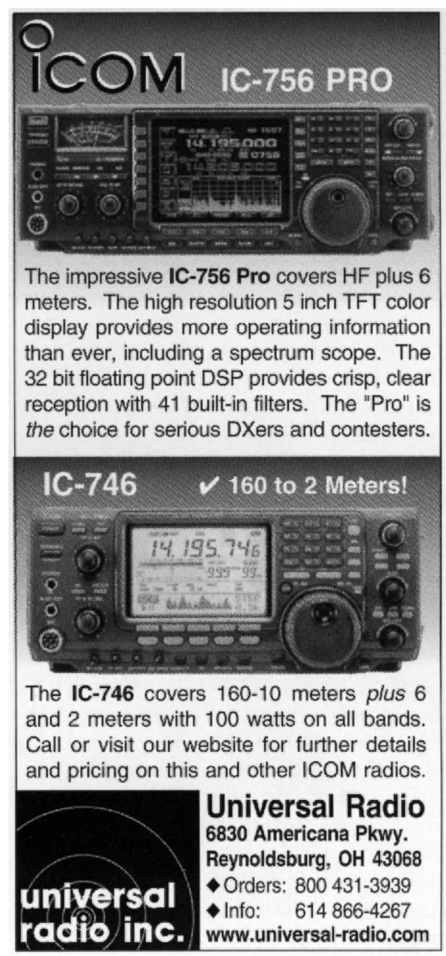
[1]G. Sabbadini, I2SG, "Microwave TR Switches Using Low-Cost PIN Diodes," *DUBUS 1/1998*, published by DUBUS Verlag GbR, Grutzmuhlenweg 23, D-22339 Hamburg, FRG, pp 30-38; **www.marsport.demon.co.uk/dubus.htm**.

[2]M/A-COM, an AMP company: Surface Mount PIN Diodes MA4P1250, MA4P1450 "SMQ." These products are presently missing from the M/A-COM site, but they are listed as available. You can download this datasheet from the ARRL Web at **www.arrl.org/qexfiles/**. Look for ZANEK1101.ZIP.

Pavel has been interested in radio communication since his twelfth year, when he began operating on the 160-meter band as OL5BAR. He became OK1DNZ in 1982.

In 1989 he graduated in electrical engineering and radioelectronics from the Czech Technical University at Prague. Pavel now works as an RF circuit designer at RCD Radiokomunikace, Ltd.

His current interests include designing and building RF Amateur Radio gear and operating on the HF bands, but he always prefers the technical side of the hobby. □□



A Laser Transceiver for the ARRL 10-GHz-and-Up Contest

An optical transceiver you can build to get on the 6500-Å band.

By Lilburn R. Smith, W5KQJ

This paper describes the construction of a laser transceiver designed specifically for the ARRL 10-GHz-and-Up contest. Earlier versions of this transceiver were presented in *Microwave Update Proceedings* for 1998 through 2000. This version has been greatly simplified. The previous amplifier incorporated automatic gain control to prevent saturation, which was found unnecessary. Also, the detector-bias dc-dc converter has been eliminated; the detector is now biased from the negative battery supply. The result is a simplified transceiver with improved performance because of noise reduction from the above changes. The transceiver is becoming popular as an entry into laser experimentation. The beginner can build this transceiver and then use the electronics with a larger telescope for DX, perhaps graduating to a higher-power laser diode for even greater range. A PC board is now available from FAR Circuits that lets experimenters easily duplicate the transceiver.

290 Robinson Rd
Weatherford, TX 76088
w5kqj@arrl.net

System Block Diagram

The system block diagram is shown in Fig 1. The receiver consists of an optical low-pass filter followed by a lens that focuses the laser energy on a silicon detector diode. The preamplifier and the audio amplifier (which drives the headphones) amplify the signal. The transmitter consists of a speech amplifier and modulator that amplitude modulates a laser diode. A single-lens telescope collimates the laser energy.

Receiver

A collecting lens is not required for direct-beam reception at one kilometer if the detector diameter is 1.5 mm or greater. However, the inclusion of a small focusing lens has a number of desirable features. The field of view and aperture become well defined, and the additional collecting area allows some gain margin to insure meeting the range requirement with a practical audio amplifier. A 25-mm (1-inch) lens with a focal length of 38 mm is available new from Edmund Scientific or Surplus Shed at low cost.[1, 2]

[1]Notes appear on page 19.

The laser detector is a large-area silicon detector from Edmund Scientific. The detector is available in several sizes. The price goes up as the size increases. The chosen detector has an area of 20.3 mm^2, a diameter of 5 mm (≈0.2 inches). The detector has a noise equivalent power of $8\times10^{-13}\ W\sqrt{H_z}$, which insures that the system will not be detector-noise limited.

The system will be dominated by sun noise in the daytime. Although a band-pass filter could be incorporated that would allow operation in daylight, the system would still be marginal for a one-kilometer range. In addition, a band-pass filter matched to the laser diode is not readily available at an affordable price. A deep red Wratten #29 filter is incorporated to remove the 60-Hz noise from incandescent or fluorescent lights. The optical-system parts chosen are shown in Table 1.

Audio Amplifier

The requirements placed on the audio amplifier in a laser receiver are severe because of the dynamic range required to accommodate the variations of the signal with range. The noise performance and gain required at minimum signal are set by the desire to have the receiver limited, in range, by the internal detector noise and the aperture, not by the shortcomings of the audio amplifier. The gain allowable at maximum signal is set by the requirement that the audio amplifier not saturate. Of course, saturation is allowable if the objective is to receive only CW, FM or PM signals. If AM voice is included in the requirements, the amplifier cannot saturate or compress excessively. The requirements for a low-noise amplifier and for high dynamic range conflict. If the receiver is to be capable of maximum DX, the audio amplifier should be designed for the least noise possible.

The amplifier presented here represents a compromise solution. The amplifier will saturate at close range. For test purposes, though, a neutral-density filter can be used to reduce the signal; the elimination of automatic gain control greatly simplifies the amplifier. The compromise audio amplifier has a gain of 96 dB and an input noise floor of about 10 µV. The dynamic-range requirements are met with a combination of switched gain and a manual audio-gain control. The transimpedance of the first amplifier is manually switched between values of 4.7 kΩ and 47 kΩ. The lower value is used for close-up tests and demonstrations, the higher value for DX. The audio-amplifier block diagram is shown in Fig 2.

The detector is operated in the photoconductive mode. A back bias is applied through a large-value load resistor so that the detector operates as a current source. The signal is capacitively coupled to a JFET op amp operated as a transimpedance amplifier followed by a fixed gain, 30-dB amplifier that has a low-pass filter incorporated. At maximum signal, the output of this stage is 10 V, which is applied to the audio-gain control. The remaining amplifiers have a gain of 66 dB.

Transmitter

The transmitter is constructed from a laser salvaged from one of the ubiquitous laser pointers available for a few dollars. The laser pointer consists of a battery power supply, a constant-current source, the laser diode and a laser detector used to control the constant current source. The laser diodes used in

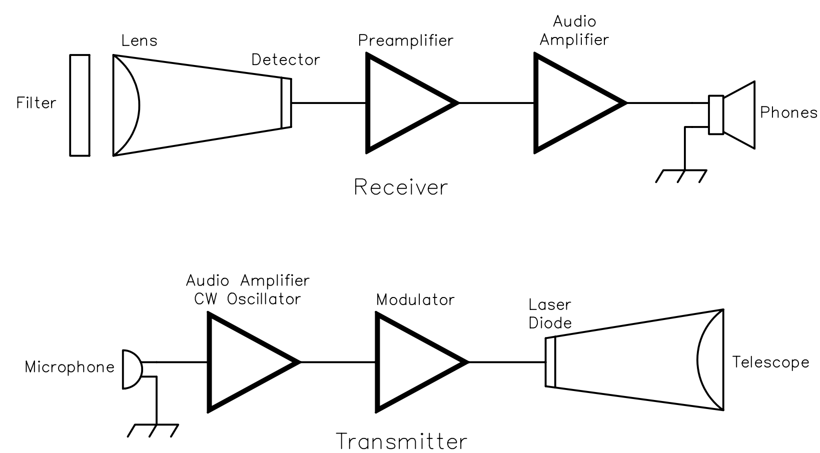


Fig 1—Transceiver block diagram.

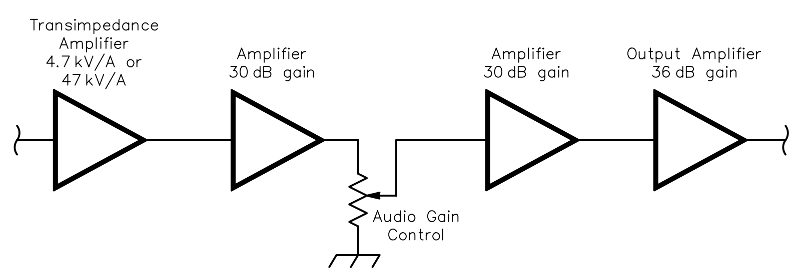


Fig 2—Audio amplifier block diagram.

Table 1—Optical Components

Lens: 25 mm diameter x 38mm focal length. F=1.5 Uncoated biconvex lens. Edmund Scientific Part Number H94822 at $10.10 each or Surplus Shed Part Number L1456 at $4.00 each.
Filter: Wratten No. 29, 620 nm low pass. Surplus Shed #L54-463, $21.20 each.
Detector: 5 mm diameter. Area = 20.3 mm^2, NEP = $8\times10^{-13}\ W\sqrt{H_z}$. Surplus Shed #L54-034, $33 each.

The chosen components yield an aperture of 25 mm and a field of view of 7.6°.

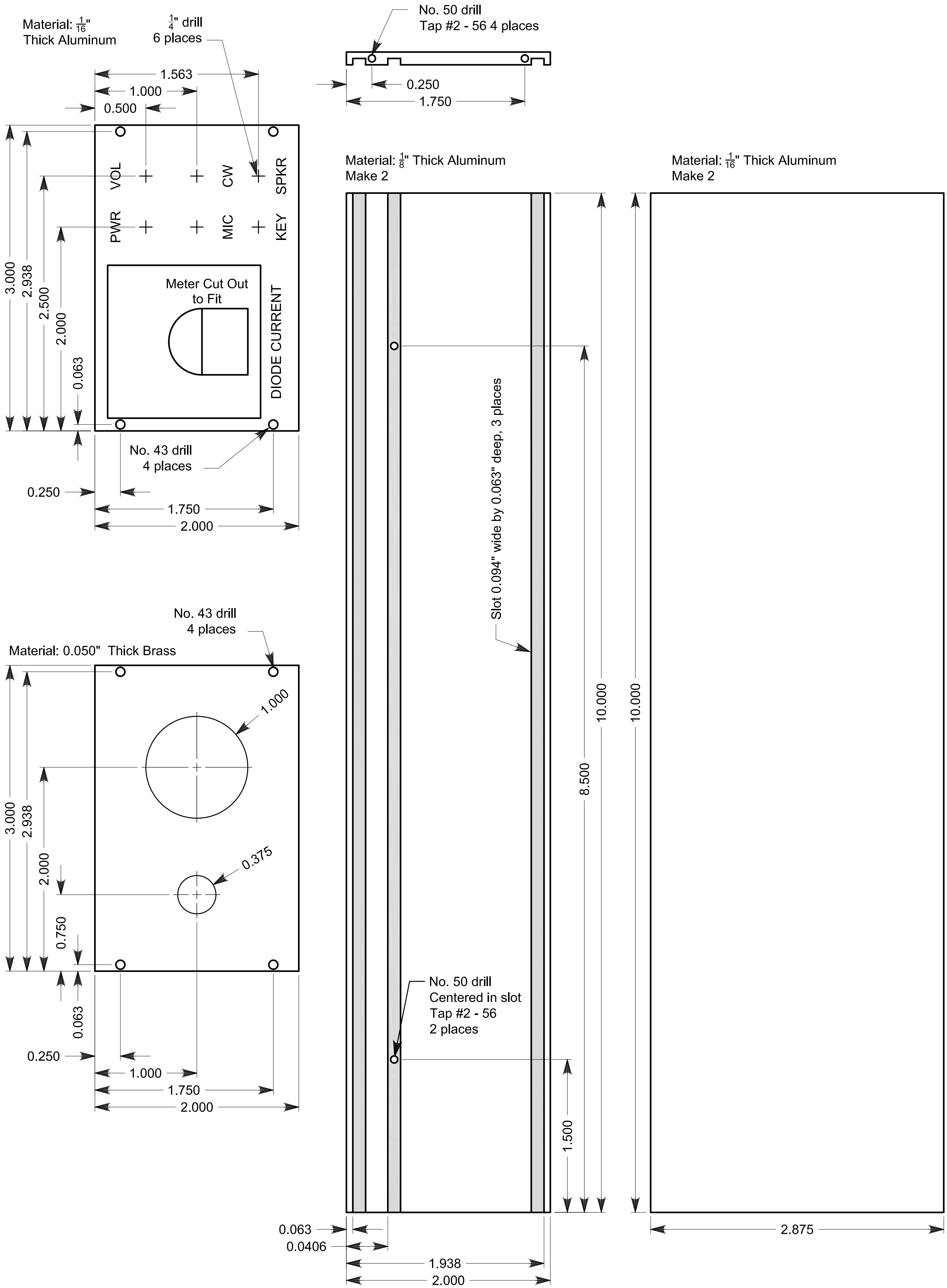


Fig 3—Details of sheet-metal parts.

the pointers are not high-quality parts. As current through a laser diode is increased, it emits light as a light-emitting diode until the lasing threshold is crossed. At that point, it emits laser light. If current is increased further, it goes into thermal runaway and destroys itself. In a good-quality diode, the threshold and thermal-runaway points are separated somewhat. In the cheap diodes, the two points are right on top of each other. A small silicon detector is added to the laser package to reduce the current and prevent thermal runaway. I have not been able to successfully drive the pointer diodes without the detector feedback. The diodes inevitably destroy themselves.

The pointer used for the prototypes was purchased from Harbor Freight Tools.[3] Their part number is 37431 and the item description is "Hi-Output Red Laser." Although any 3-V laser pointer will work, this particular one uses Sony diodes and the lens assembly is glass and brass instead of plastic. The price depends on the catalog from which you order. I have paid from $5.95 to $19.95 at various times. The latest catalog (551) lists the laser as part number 37431-6UEH and the price is $11.99. They go on sale often for half price.

Use the entire laser assembly, including the current-limiter PC board. Otherwise, you will surely experience laser-diode thermal runaway.

The transmitter electronics consist of an op amp that is used as an audio amplifier for voice operation and as an oscillator for CW operation. The output of the audio amplifier is centered on –3 V dc, which is the voltage the pointer uses. The modulation swings the voltage from ground to –6 V. The current supply in the original pen circuit limits the laser diode current. The limiter is such that it allows voice or CW modulation.

Transceiver Specifications

A brief summary of the transceiver specifications is given in Table 2. The transceiver circuit could be used with a better telescope to achieve DX operation. The 3-mW laser diode and low-noise detector will allow a maximum range of about 100 km with a large receiving lens. The transceiver is intended for contest operation at much shorter ranges. Long-range operation requires extremely good pointing accuracy and a line-of-sight path. The pointing requirements take up too much time for contest operation. The idea is to design a transceiver that can be easily duplicated in large numbers for contacts over the minimum path required by ARRL rules, which is 1 km for the 10-GHz-and-Up contest.

Construction

Construction of the transceiver should proceed in subassemblies followed by final assembly. The transceiver is built in a sturdy metal case. The sheet-metal parts are shown in Fig 3. Most of the pieces are aluminum; but the front panel and optical assembly are made of copper, since the optical housing is soldered to the front panel. Make two each of the sides and top pieces and one each of the front and rear panels. The top, bottom and printed circuit board are all mounted in slots in the side pieces, which are held captive by the front and back panels.

The optical assembly detail is shown in Fig 4. The lens holder is made of a 1-to-3/4-inch copper pipe-reducing coupling. The coupling is carefully soldered to the front panel, keeping the parts square with each other. The detector is mounted on a 3/4-inch pipe cap. The short section of pipe allows the image to be focused by sliding the cap back and forth. The filter is bonded to a ring made of 1-inch pipe. All solder work is done before mounting the optical parts,

Table 2—System Specifications

Receiver	
Lens Diameter	25 mm
Clear Aperture	22 mm
Field of View	7.6°
Noise Equivalent Power	10^{-8} W
Optical Passband	620-1200 nm
Optical Transmission	0.8
Detector Diameter	5 mm
Audio Gain	96 dB
Audio Passband	500-1500 Hz
Audio Output	5 V (RMS)
Headphones	32 Ω
Transmitter	
Wavelength	640 nm
Power	3 mW
Beam Divergence	0.5 × 1 milliradians
Modulation	AM, CW, data
Modulation %	100
Microphone	Electret
System	
Power Supply	12 AA cells
Size (including projections)	12×3×2.5 inches (LWH)

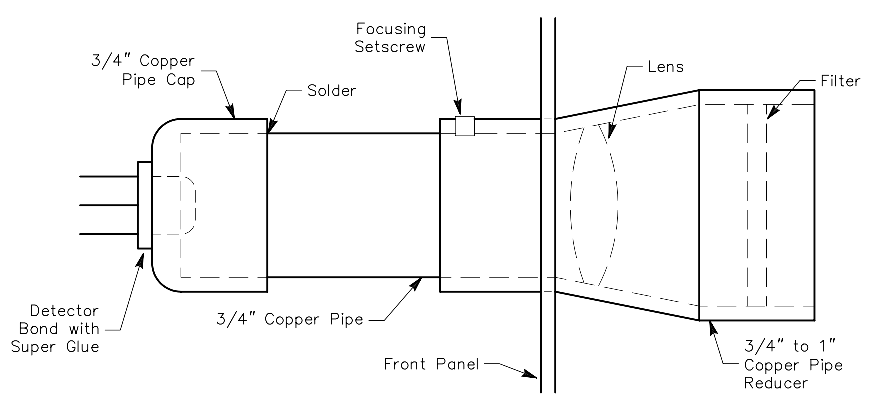


Fig 4—Details of the receiver optical assembly.

which would be destroyed by the heat.

The laser pointer is modified by carefully removing the plastic housing. Unsolder the battery clips. Remove the subminiature switch. Insert a jumper to replace the switch so that the laser is always on. Solder a wire to the "B–" pad and a wire to the ground pad. The laser is mounted to the front panel using a hardware-store bushing that is soldered to the brass or copper front panel before inserting the laser diode assembly. Shim the lens with metal tape for a tight fit.

The rear-panel assembly contains several controls (listed in Table 3), connectors and a meter. The rear panel layout is shown in Fig 5.

The transmitter and receiver are built on one PC board. The schematic is shown in Fig 6. The PC board is etched from G-10 $^1/_{16}$-inch-thick material and carefully trimmed to the exact dimensions shown, which fit in the slots in the sides. The component layout is shown in Fig 7.

The chassis wiring is shown in Fig 8. Use shielded wire for the **VOLUME** control and audio-output wiring. Dress all the short leads to the front panel so that the transceiver can be assembled without stressing the wires. The detector leads fit directly into pads on the PC board. Form the leads for minimum stress. The 3.3-Ω meter shunt is mounted on the meter terminals. The 0.22-µF capacitor is mounted between the switch and the jack terminals. The 47-kΩ resistor should be mounted directly on the switch with short leads. A 10-kΩ, RV6 variable resistor is used for the **VOLUME** control. The meter is a 250-µA movement bought surplus. Any suitable meter can be used, provided it can be shunted to provide an 80-mA full-scale deflection.

The recommended order of construction is as follows:

1. Make the side rails of aluminum. Mill or saw the slots. Tap the holes where required.
2. Make the front and back blank panels. The front panel is made from sheet copper and the rear panel is aluminum. Drill or punch all holes.
3. Make the top and bottom covers.
4. Do a trial assembly of the case. The front and rear panels should hold the rails squarely. The top and bottom covers should slide freely into their slots. If required, sand and file the case parts for proper fit.
5. Disassemble the case.
6. Paint the side rails on the outside surfaces only and lay aside to dry.
7. Make a temporary jig to hold the front panel, lens pipe reducer and laser bushing square. Solder carefully, using an absolute minimum of solder. Clean the solder flux and any excess from the assembly. Mask the inside surfaces of the reducer and bushing, and the back surface of the panel. Paint the remaining surfaces with good-quality gloss enamel and allow them to dry. Bake in a kitchen oven (not a gas oven!) at a low (warm) setting for two hours. Mask the painted surfaces and paint the inside of the reducer with flat-black enamel as sold for barbecue grills. Allow them to dry and then bake.
8. Mount the lens in the reducer, with the side that has the most curvature forward. Use "super glue" or model-airplane cement. Do not get cement on the lens surfaces. Model-airplane cement can be cleaned off with acetone but super glue is there for good.
9. Make a filter-mounting ring of 1-inch copper pipe. Paint only the inside surface with the black barbecue enamel. Cut the filter carefully to shape. Mount the filter on its mounting ring. Be careful to partially insert the filter ring into the lens holder before cementing to avoid wrinkling the filter. Use the absolute minimum cement and do not get it on the filter surfaces. Push the assembled filter into the lens reducer.
10. Make a detector holder from a $^3/_4$-inch pipe cap and a short pipe. Drill or punch the hole for the detector. Cut the cap for a total length of $^1/_2$ inch and solder it to the $^3/_4$-inch pipe. Wait to bond the detector to the holder. Insert the detector holder onto the pipe. Take the entire optical assembly outside and hold a piece of thin paper to the back of the detector holder. Slide the cap back and forth while observing the sun image on the paper. *CAUTION*: *This step can be dangerous. Do not focus the sun's rays onto your eyes or any other part of your body.* Cut off the pipe if required to obtain focus. When the image is sharp, lock down the setscrew.
11. Bond the detector to the detector holder. Be careful to align the leads to fit properly in the circuit board.
12. Prepare the laser assembly as described above and mount it in the bushing using copper tape as a shim for a press fit. Be neat with the shim or the laser will not be aligned properly. Lay aside the optical assembly in a safe place.
13. Build and test the PC board. Notice that the holes are not plated through and solder the leads on both top and bottom. Some components must be mounted slightly raised to solder the leads. The board is tested with an audio signal generator and variable attenuator capable of 100-dB attenuation. Apply the audio signal to the detector input through a 4700-Ω resistor and check the board for gain.

Table 3—Rear Panel Controls and Connections

PWR	Controls power to receiver and transmitter
VOL	Controls receiver gain
MIC	Microphone input for transmitter
CW	Switch between voice and CW
KEY	Straight-key input to transmitter
SPKR	Receiver audio output
DIODE CURRENT	Laser-diode current meter

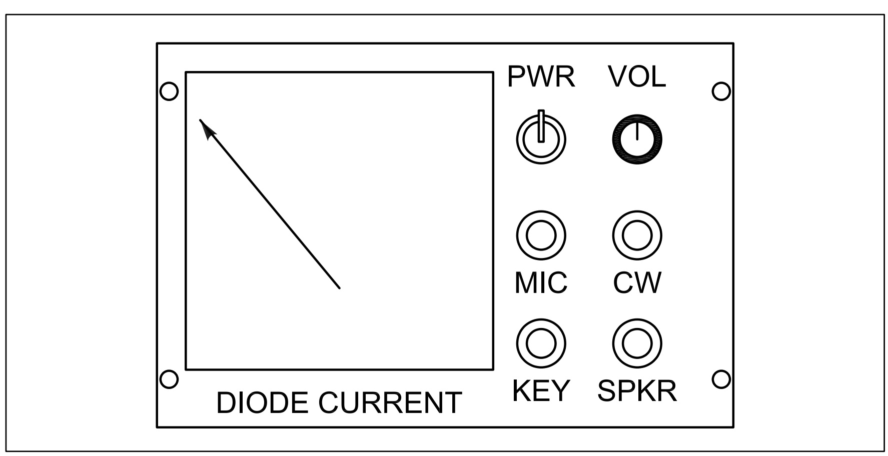


Fig 5—Rear-panel layout.

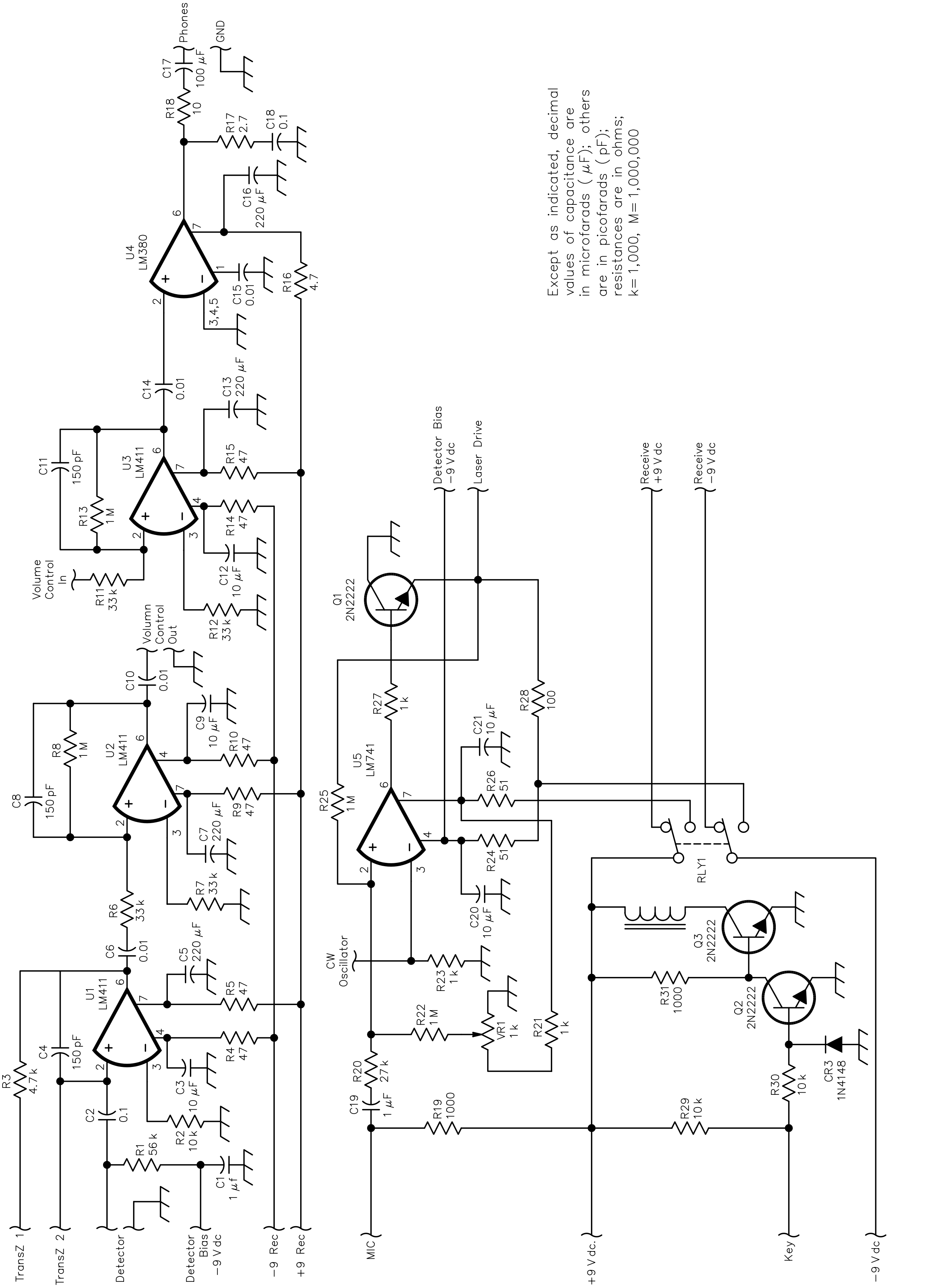


Fig 6—Schematic of PC-board circuit.

Troubleshoot as required. Temporarily connect the 10-kΩ **VOLUME** control to make the remainder of the tests. Reduce the signal level with the **VOLUME** control until the output stage is not distorted, then measure the gain. The gain from the **VOLUME** control wiper to the audio output is 66 dB. Check the noise floor of the amplifier. It should be better than –100 dBV. Without the laser connected, key the relay and set VR3 for exactly –3 V on the laser drive pad. Temporarily connect the microphone. The laser drive voltage should vary between 0 and –6 V when you key the mike and whistle. Check for the proper detector-bias voltage, which should be –9V dc. The testing of the PC assembly is complete.

14. Mount the parts and wire the front panel with sufficient lengths of wire to cut the wire to length in the next step.

15. Make a trial assembly with the side rails and the front and back panel assemblies, but without the top and bottom panels. Insert the PC board in its slots. Glue it in place with model-airplane cement. Cut the wires to length and wire the PC assembly. Do not cut the detector leads. They should be the proper lengths to reach the board with a gentle curve in the leads.

16. Mount the battery boxes on the top plate. The battery pack is made up of 12 AA batteries mounted in three holders of four cells each. The holders are screwed to the top plate. The batteries are arranged in +9-V and –9-V supplies. Wire all three in series, and make a tap at the center of the center box by inserting a #4-40, flat-head screw into the proper rivet. Attach the nut and ground lug. Cut the wires to length and wire the battery holders to the power switch. Ground the center tap to the PC board.

17. Assembly of the unit is somewhat tricky. Turn off the power switch. Insert the batteries. Check for ±9 V at the power switch with the switch off. Turn on the unit and make any checks you want. Turn power off. Remove the screws from the rear panel. Carefully slide the top and bottom covers into their slots. Be careful not to pinch any wires. Replace the screws.

The assembled transceiver is shown in Fig 9.

Results

Optical

Optical transmission through the lens and filter is 80%. The field of view is 7.6°. With the full-transmitted power illuminating the detector, the output of the transimpedance amplifier in the low-gain mode is 3 V. The beam divergence of the transmitter is 0.5×1 milliradians. The laser diode is modulated 100%.

Audio Amplifier

The completed audio amplifier displays a gain of 96 dB with the **VOLUME** control at maximum. The noise voltage at the input for minimum-detectable signal is 10 µV. The 6-dB frequency-response points are 500 and 1500 Hz.

Field Test

Two completed transceivers were set up 1096 meters apart. Reliable communication between W5KQJ and KD5FFW was demonstrated during the 1999 ARRL 10-GHz-and-Up Contest before members of the North Texas Microwave Society. The observed laser spot was an oval shape approximately one-half by one meter that was barely perceptible to the eye. As expected, the major problem in the test was positioning the laser spot. Two heavy-duty camera tripods with pan/tilt mechanisms

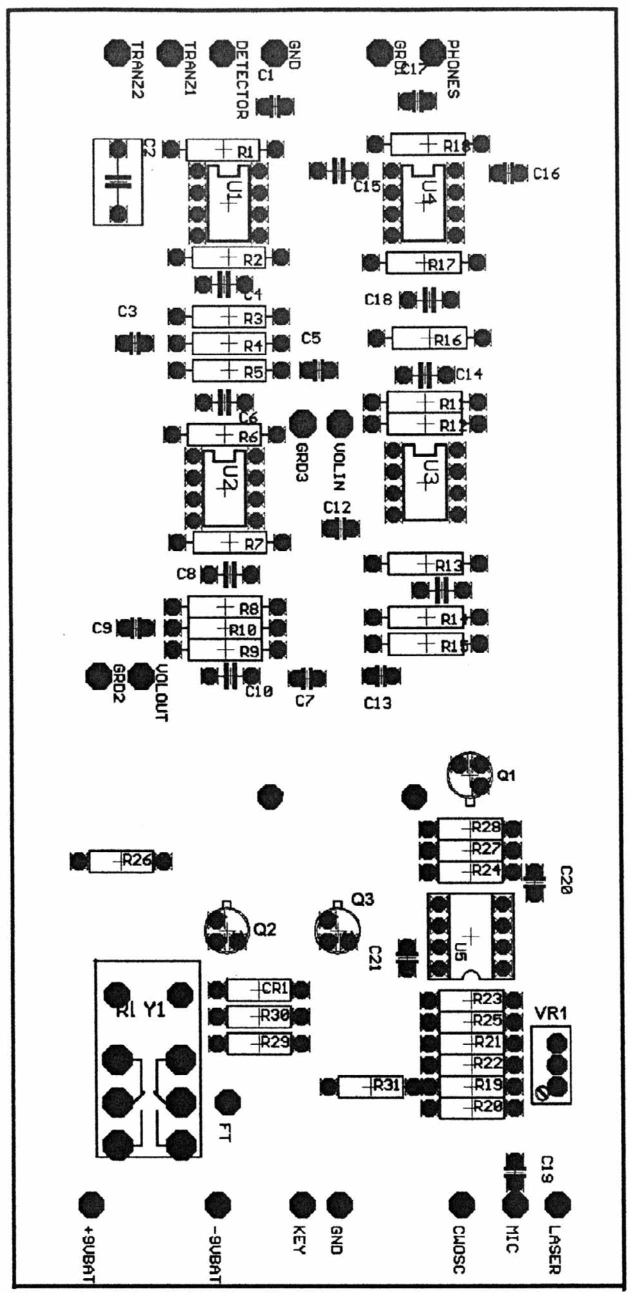


Fig 7—PC-board parts-placement diagram.

were used; but even so, positioning the spot was very difficult. The signal-to-noise ratio was excellent and appeared to bear out the mathematical model.

Other Applications

Uses of the laser transceiver are not limited to voice and CW or to contest use. By using *Hamcom* software and the usual simple interface, RTTY can be used. Slow-scan television can be transmitted using *JVFAX*. Packet is a natural. By removing the low-pass filter capacitors, video can be transmitted and received. The transceiver will work full duplex if the transmit-receive relay is replaced by appropriate jumpers. PSK-31 transmission and reception was demonstrated to the North Texas Microwave Society during the 2001 10-GHz-and-Up contest.

The PC Board

A PC board for the transceiver is available from FAR circuits. Send e-mail to **w5kqj@arrl.net** for information. Unfortunately, the tooling charges for a board with plated-through holes are too expensive, so the board must be soldered on both sides. Al-

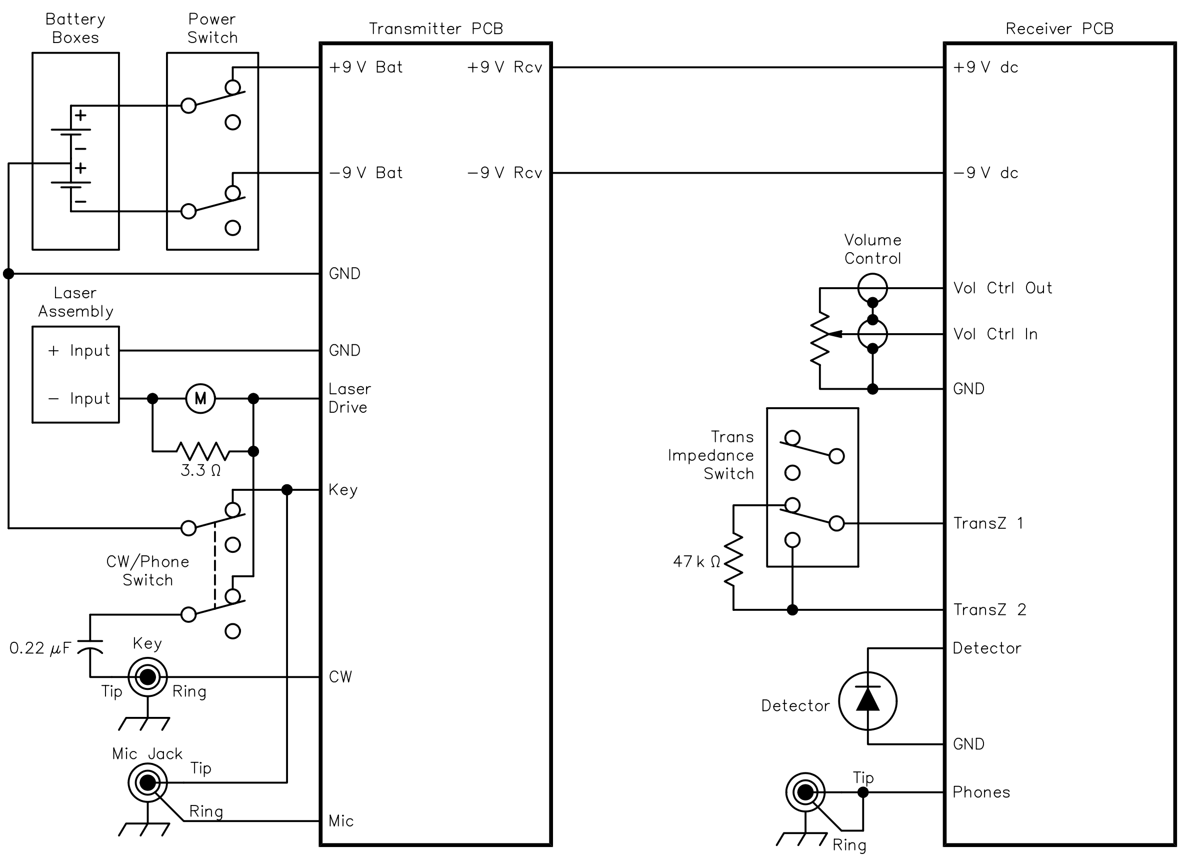


Fig 8—Chassis wiring.

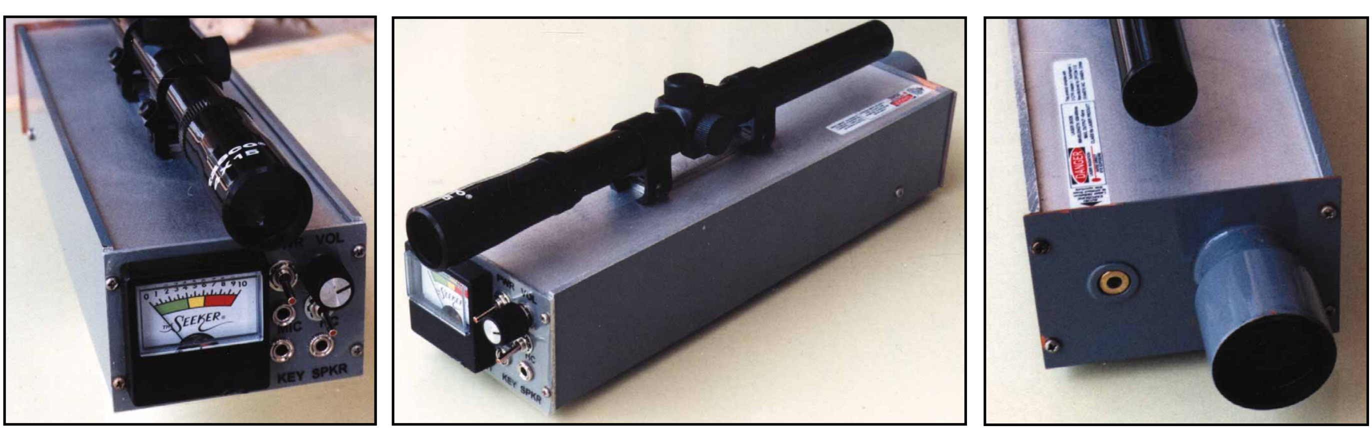


Fig 9—Several views of the completed transceiver.

Parts List

All resistors are 1/4-W carbon-film parts unless otherwise specified.

PC Board

C1, C19—1 µF tantalum (RadioShack 900-1662)
C2, C18—0.1 µF film (RadioShack 900-2265)
C3, C9, C12, C20, C21—10 µF tantalum (RadioShack 900-1605)
C4, C8, C11—150 pF ceramic (RadioShack 900-2203)
C5, C7, C13, C16—220 µF tantalum (RadioShack 900-1612)
C6, C10, C14, C15—0.01 µF film (RadioShack 900-2253)
C17—100 µF tantalum (RadioShack 900-1610)
Q1-Q3—2N2222 transistor (Digi-Key 2N2222AMS-ND)
R1—56 kΩ (Digi-Key 56KQBK-ND)
R2, R29, R30—10 kΩ (Digi-Key 10KQBK-ND)
R3—4.7 kΩ (Digi-Key 4.7KQBK-ND)
R4, R5 ,R9, R10, R14, 15—47 Ω (Digi-Key 47QBK-ND)
R6, R7, R11, R12—33 kΩ (Digi-Key 33KQBK-ND)
R8, R13, R22, R25—1 MΩ (Digi-Key 1.0MQBK-ND)
R16—4.7 Ω (Digi-Key 4.7QBK-ND)
R17—2.7 Ω (Digi-Key 2.7QBK-ND)
R18—10 Ω (Digi-Key 10QBK-ND)
R19, R21, R23, R27, R31—1 kΩ (Digi-Key 1.0KQBK-ND)
R20—27 kΩ (Digi-Key 27KQBK-ND)
R24, R26—51 Ω (Digi-Key 51QBK-ND)
R28—100 Ω (Digi-Key 100QBK-ND)
RLY1 —DPDT relay (preferred: Aromat DS2E-M-DC9V, Allied 788-1341; alternates: Aromat DS2E-M-DC12V, Digi-Key 255-1071-ND)
U1, U2, U3—LF411CN IC (Digi-Key LF411CN-ND)
U4—LM380N IC (Digi-Key LM380N-8-ND)
U5, U6—LM741N IC (Digi-Key LM741CN-ND)
VR1—1 kΩ variable resistor (Digi-Key 3299W-102-ND)

Chassis

Battery box, 4- AA
Transimpedance resistor—47 kΩ
Meter shunt—3.3 Ω
0.22 µF capacitor, film
Detector—See text
Wratten 29 optical filter
Key, mic, phones jack, two-circuit 3.5 mm
Laser assembly See text
Detector mount—3/4-inch pipe cap
Focus adjustmen/—3/4 inch pipe, 1 inch long
Filter Holder—1-inch pipe, 1/8 inch long
Lens holder—3/4-1 inch pipe reducer
Power, CW switches—toggle DPDT
Transimpedance switch—slide DPDT
Meter—250 µA from All Electronics, #MET-51
Volume—10 kΩ variable resistor #RV6
Note: All 900-series RadioShack part numbers are from their on-line catalog.
In-stores, the part numbers are different and much higher priced.

though the lack of plated-through holes is an annoyance in assembly, the board is of good quality and the result is good.

Conclusion

The laser transceiver described in this paper is capable of communication at a range of 1 km as required by ARRL rules. It is reproducible at a moderate cost without previous microwave or optical-communication experience.

Notes

[1]Edmund Scientific, 60 Pearce Ave, Tonawanda, NY 14150-6711; tel 800-728-6999, fax 800-828-3299; scientifics@edsci.com; www.edmundscientific.com/.

[2]Surplusshed.com, 407 US Rte 222, Blandon, PA 19510; tel 610-926-9226, fax toll-free in US and Canada 1 877-7SUPLUS (78-7758); surplushed@aol.com; www.surplusshed.com.

[3]Harbor Freight Tools, 3491 Mission Oaks Blvd, Camarillo, CA 93011-6010; tel 800-423-2567; www.harborfreight.com.

Lilburn R. Smith was first licensed in 1956 as WN5KQJ. He upgraded to W5KQJ in the same year. In 1977-1979, he lived in Orlando, Florida, where he was W4PMX. He holds an Extra license. He has a BSEE degree from Texas Tech University. He holds one patent for a Laser Guided Projectile and has one patent pending for an advanced display for IR images. Lilburn has been involved in microwave, VHF and laser design and development since 1959. He recently retired after a 41-year career in aerospace.

VARI-NOTCH® DUPLEXERS
FOR 2 METERS

The TX RX Systems Inc. patented Vari-Notch filter circuit, a pseudo-bandpass design, provides low loss, high TX to RX, and between-channel isolation, excellent for amateur band applications. TX RX Systems Inc. has been manufacturing multicoupling systems since 1976. Other models available for 220 and 440 MHz, UHF ATV and 1.2 GHz.

19″ RACK MOUNT

MODEL 28-37-02A
144-174 MHz
92 dB ISOLATION AT 0.6 MHz SEPARATION
400 WATT POWER RATING

TX RX SYSTEMS INC.
8625 INDUSTRIAL PARKWAY, ANGOLA, NY 14006
TELEPHONE 716-549-4700 FAX 716-549-4772 (24 HRS.) e-mail: sales@txrx.com
A MEMBER OF THE BIRD TECHNOLOGIES GROUP

HF Circuits for a Homebrew Transceiver

Work the world with these front-end circuits for a high-performance radio.

By Mark Mandelkern, K5AM

HF is the backbone of ham radio. Even at a station designed primarily for 6-meter DX work and VHF contesting, an HF radio can provide countless hours of fun while DXing, contesting and rag-chewing. Circuits for the IF stages in a high-performance homebrew transceiver have been described in previous articles.[1, 2, 3] The main transceiver panel operates at the first IF, tuning 40-39 MHz. Three front-end sections effect the conversions to HF, 50 MHz and 144 MHz. The present article will describe the HF section, shown in Fig 1.

[1]Notes appear on page 41.

5259 Singer Rd
Las Cruces, NM 88005
k5am@zianet.com, **k5am@arrl.net**

General Plan

The circuitry in the HF section includes mixers, amplifiers, and filters for converting the transceiver range of 40-39 MHz to the HF amateur bands. A 200-W tetrode amplifier is included. Use of a tube at this point provides low-distortion performance rarely achieved with solid-state amplifiers at this power level. For receiving, an RF amplifier, a mixer, and a post-mixer amplifier are employed. For protection against out-of-band signals, a high-Q front-end preselector is used. The front panel, which allows control of all essential parameters, is shown in Fig 2.

The main transceiver panel tunes the 39-40 MHz range in reverse, enabling high-side injection in the front-end sections. To cover the ham bands between 1 and 30 MHz, LO injection is required on 10 frequencies in the 41 to 69-MHz range. To avoid instability caused by a crystal switch, ten separate crystal oscillators are used, with diode switching of the oscillator outputs.

For transmitting, a mixer and a seven-stage solid-state amplifier drive a conduction-cooled tetrode power amplifier (PA). ALC is used for limiting of output power, reverse power due to high SWR, screen current, and grid current. External ALC input is also accepted from a kilowatt-level amplifier. Three of the driver stages are used primarily as buffers to ensure smooth ALC functioning without distortion.

The PA uses an 8072 tetrode; it is electrically identical to the more familiar air-cooled 8122. The 8072 has no cooling fins; its anode is clamped to a heat sink. Alternatively, an 8122 may be directly substituted, if forced-air cooling is provided. Similarly, the popular 4CX250B may be used. The control and protection circuits shown here may also be used for other small tetrode amplifiers at any frequency.

Receiver Circuits

Fig 3 is the circuit diagram of the receiving section. Signals arrive from the antenna relay in the PA section and proceed to a switched attenuator, a high-pass filter to reject broadcast-band signals, a low-pass filter to reject images at VHF and signals at the 40 MHz IF, a preselector and an RF amplifier. Included at the input to the receiver circuits is a reed relay, serving to ground any stray RF energy from the transmitter or from the auxiliary receive antenna.

No AGC is applied to the front-end section. AGC is applied only to the IF strip in the main transceiver panel at 9 MHz, after the crystal filters. This arrangement allows the IF gain to be controlled with no loss of sensitivity in the front-end. The mixers are built to handle strong signals, and the sharp 9-MHz crystal filter following the second mixer effectively keeps off-channel signals out of the AGC circuits. The front-end runs wide-open at full sensitivity– the best arrangement for weak signals. This was discussed in more detail in Part 1, page 18 (see Note 3). The main transceiver panel has an IF-gain control; it can be used for AGC-threshold operation. This provides maximum readability for very weak signals, while at the same time keeping the AGC circuits active when ear-shattering strong signals come on the same frequency—a very common occurrence in DX operation. AGC threshold operation is more effective and much safer than the all-too-common practice of operating with the AGC off. This receiving method is discussed more fully in Part 1, pages 21-22.

Auxiliary Receiving Antenna

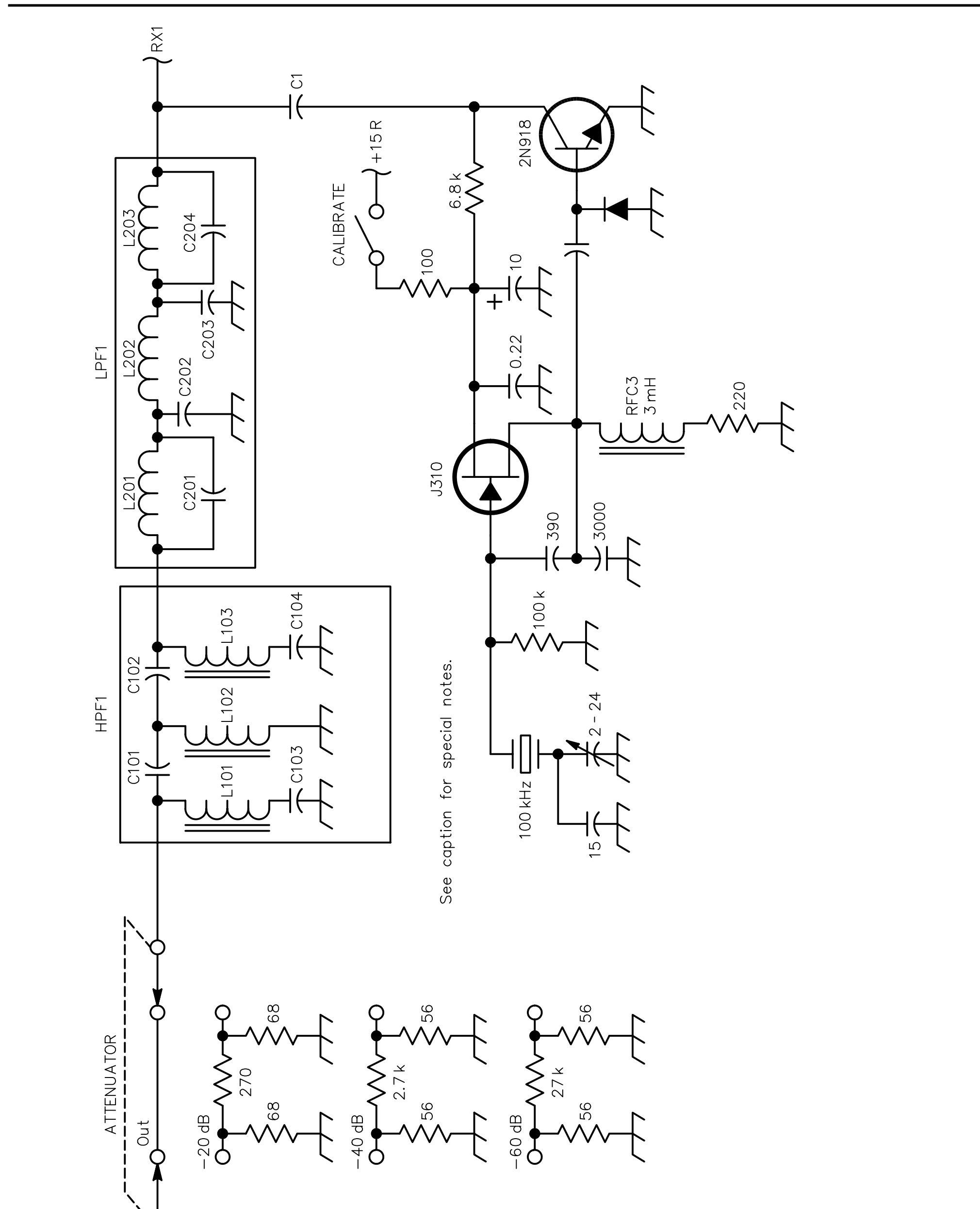
For DX operation on the 160-meter band, panel control of an auxiliary receive antenna is mandatory. In this radio, there is also provision for control of the **RX IN** jack by a switch on the operating bench or a foot switch. This external control operates not merely in parallel with the panel switch, but serves to reverse the selection on the panel. This is required because the usual set-up on 160 meters is to receive on the auxiliary antenna, with some arrangement for occasionally trying the transmitting antenna. Thus the panel switch can select the auxiliary antenna and the foot switch will then momentarily select the transmitting antenna. With a simple parallel switch, the operator would need to keep a foot on the foot switch all night. There is also an RX OUT jack; together with the RX IN jack, it can be used for an external filter or for a second receiver.

Receiver Front-End Attenuator

The front-end attenuator is necessary for optimum performance on the lower bands, since no attempt has been made to adjust the band-by-band front-end gain to fit typical ambient-noise and signal-level conditions. One advantage of this method is that high gain and good noise figure are available for use with Beverage, loop or other low-output receiving antennas. The attenuator has positions for 0, 20, 40 or 60 dB of attenuation. Smaller steps were considered unnecessary. On 40 and 80 meters, 20 or 40 dB attenuation is commonly used. The reason for the 60-dB position is simple: There was one more position on the switch I happened to find in the junk box. Unexpectedly, the last position did turn out to be useful. On the higher bands, the 60-dB attenuator is nearly equivalent to a dummy load; it provides a very convenient way to compare antenna noise with internal receiver noise. This is not only easier, but also better than removing the antenna, because a proper comparison requires a 50-Ω load at the antenna terminal.

Fig 1—HF front-end section for the K5AM homebrew transceiver. The PA tank compartment is at the left rear, with its shield cover and exhaust fan removed. The 8072 conduction-cooled tetrode is clamped inside the large block. The 160 and 80-meter toroids are at the left of the PA compartment. The long "bread-slicer" capacitor is for tuning the 24-30 MHz range. (Photos by Lisa Mandelkern)

Fig 2—Front panel. Rather than trying to achieve uniformity in knob style, several different styles are used deliberately; this helps the operator find the correct knob very quickly. For consistency and additional convenience, the same style choices are also used on the other front-end sections for corresponding controls.



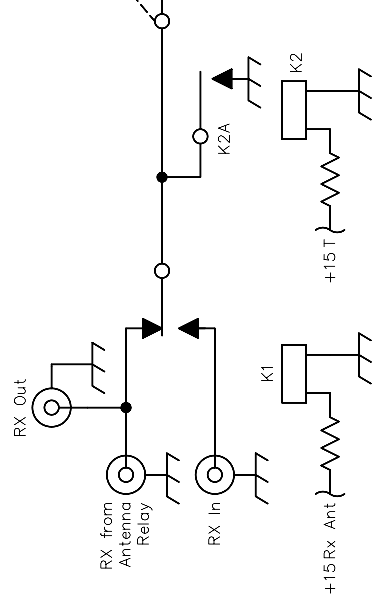


Fig 3—Receiver circuits. Except where otherwise indicated, the schematics in this article use the following conventions. Each resistor is a $^1/_4$ W carbon-film type. All trimpots are one-turn miniature units, such as Bourns 3386F; DK#3386F-nnn, MO#652-3386F-1-nnn. Stock numbers for Digi-Key and Mouser are given as DK# and MO#. Suppliers' contact information appears in Table 4. All unmarked coupling and bypass capacitors are 10-nF, 100-V, disc-ceramic units. Those marked 0.1 or 0.22 are monolithic ceramic types. Each control and power terminal has a bypass capacitor or feedthrough capacitor, as appropriate, often not shown. Also not shown are the bypass capacitors across each rectifier and switching diode and each switching transistor base-emitter junction. Electrolytic capacitors have 25 V ratings, with values given in µF; those less than 100 µF are tantalum types. Capacitors labeled "s.m." are silver-mica units, with values given in pF. Feed-through capacitors are 1 nF, 1 kV, ceramic components. Values of RF chokes (RFC) are given in µH. The MOSFETs are small-signal dual-gate VHF units. Type 3N140 is used here, but any similar type may be substituted. Replacement NTE 221 is available from Hosfelt. The JFETs are J310s. The diodes are small-signal silicon, such as 1N4148. The bipolar transistors are 2N4401 (NPN) and 2N4403 (PNP), or similar. The op amps are all sections of quad LM324N. Each op amp is powered by the +15 V and –15 V rails. Not shown is the bypassing at each op-amp power terminal, a 100-nF monolithic ceramic capacitor. The toroidal coils are wound with #26 enameled wire; the cores are available from Amidon. The various control lines are provided by the control board. Potentiometers labeled in all capital letters are front panel controls; others are circuit-board trimpots for internal adjustment. Also not shown are the usual diodes and bypass capacitors across each relay coil. The unmarked resistors at certain relay coils are selected to accommodate the particular relay installed. In the receiver circuits, the capacitors in the preselector and the filters are silver-mica types.

BPF1-BPF2—Band-pass filter, center frequency 39.5 MHz, bandwidth 3 MHz. The inductors are 480 nH, 11 turns on T37-10 iron-powder toroidal cores. The end coupling capacitors are 12 pF. The resonating capacitors are 22 pF. The top coupling capacitor is 2 pF.

C1—Gimmick capacitor; see text.

HPF1—High-pass filter, for rejecting AM broadcast signals. Measured cut-off frequency is 2 MHz; insertion loss over the HF range is less than 0.5 dB and 5 dB at 1.8 MHz. Attenuation is 20 dB at 1.4 MHz and 50 dB or better below 1 MHz. L101 and L103 are 10 µH, L102 is 2.2 µH; these are small molded inductors. C101 and C102 are 1000 pF, C103 and C104 are 3000 pF. This is a CX7 design.

K1—Auxiliary receive-antenna relay. High-frequency, low-loss, high-isolation relay, SPDT, 12 V dc, 25 mA coil. Omron #G5Y-1-DC12, DK#Z724.

K2-K14—Miniature reed relay, SPST normally open, 12 V dc coil, 1450 Ω, 8 mA. Gordos #0490-1478DZ, or similar relay. Hosfelt #45-191.

L1—53 turns on type T68-2 iron-powder toroidal core, tap 2 turns from low end.

L2—33 turns on type T50-6 iron-powder toroidal core, tap 11 turns from low end.

L3—21 turns on type T37-12 iron-powder toroidal core, tap 1 turn from low end.

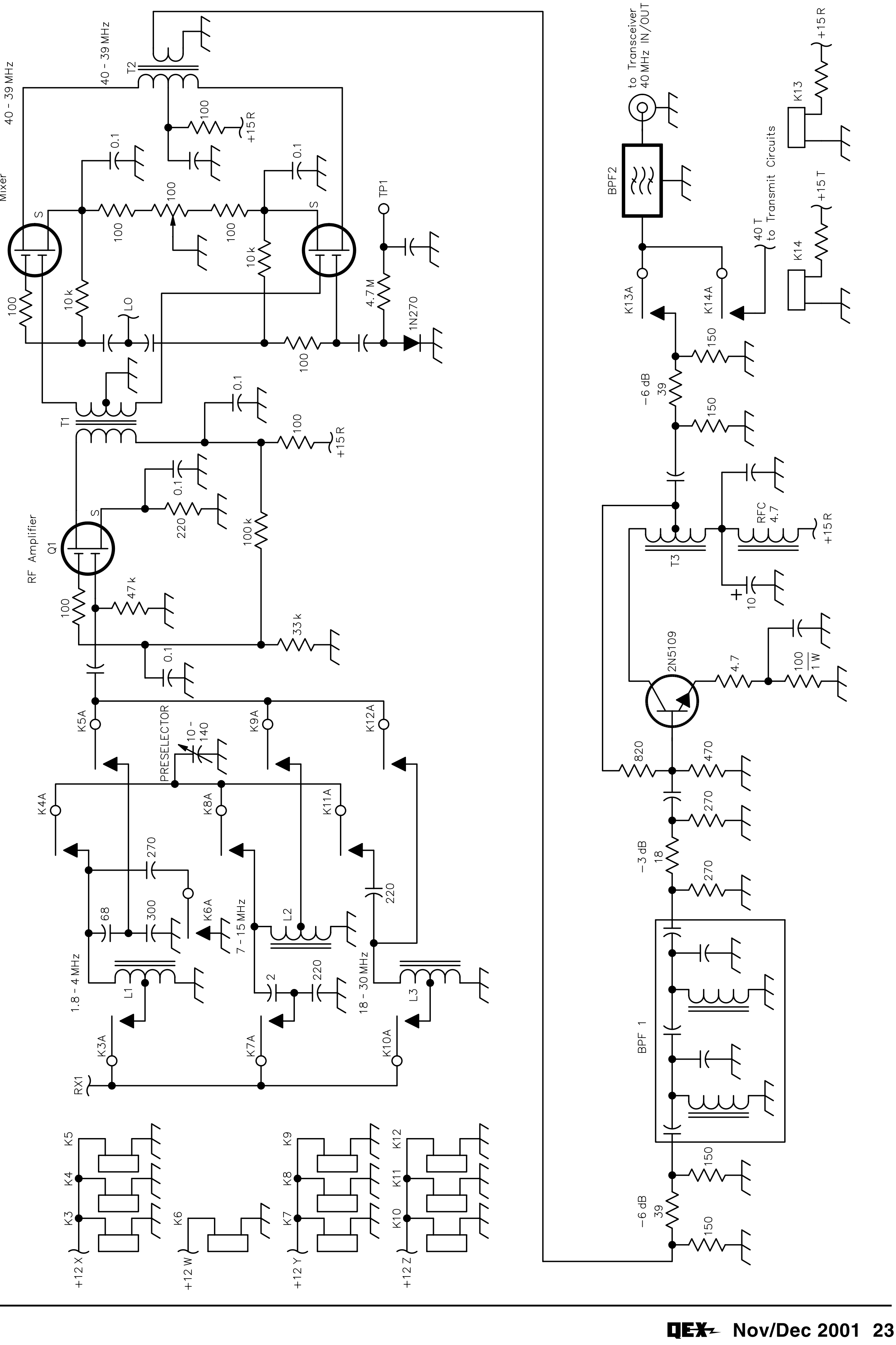
LPF1—Low-pass filter, for IF and image rejection. Measured cut-off frequency is 30.3 MHz; insertion loss 0.5 dB, attenuation 50 dB at 40 MHz. L201 and L203 are 170 nH, 8.5 turns #24 enameled wire close-wound on a phenolic form, 0.140-inch diameter. L202 is 520 nH, 17.5 turns #28 enameled wire close-wound on a phenolic form, 0.140-inch diameter. C201 and C204 are 100 pF. C202 and C203 are 150 pF. This is a CX7 design.

Q1—3N211 This is similar to the 3N140 used in other stages, except that in this stage, close to the antenna, a gate-protected version is used. Replacement NTE 222 is available from Hosfelt.

T1—Trifilar transformer, 14 turns on FT37-43 ferrite toroidal core.

T2—Primary: 16 turns center-tapped on type FT37-61 ferrite toroidal core. Secondary: 2 turns wound at center of primary. This turns-ratio converts the 50 Ω load to a 1600 Ω load at each drain.

T3—Bifilar-transformer, 7 turns on FT37-43 ferrite toroidal core.


Mixer
40 - 39 MHz
40 - 39 MHz
T2
100
+15 R
0.1
S
100
100
100
0.1
S
TP1
10 k
10 k
LO
4.7 M
1N270
100
100
T1
0.1
100
+15 R
RF Amplifier
Q1
S
0.1
220
100 k
100
47 k
33 k
0.1
K5A
PRESELECTOR
10 - 140
K9A
K12A
K4A
K8A
K11A
270
68
300
K6A
7 - 15 MHz
L2
220
1.8 - 4 MHz
L1
2
220
18 - 30 MHz
L3
K3A
K7A
K10A
RX1
K5
K4
K3
+12 X
K6
+12 W
K9
K8
K7
+12 Y
K12
K11
K10
+12 Z
to Transceiver
40 MHz IN/OUT
BPF2
K13
+15 R
to Transmit Circuits
40 T
K14
+15 T
K13A
K14A
150
−6 dB
39
150
RFC
4.7
+15 R
T3
10
2N5109
4.7
100
1 W
820
470
270
−3 dB
18
270
BPF 1
150
−6 dB
39
150

Preselector

Rejection of out-of-band signals has become one of the most prominent measures of performance for modern receivers. Most other performance factors, some even more important, have reached levels in excess of practical requirements. Modern factory-built radios have attacked the problem of out-of-band signals with a variety of front-end filter arrangements, with varying success.

Older radios from the 1940s and earlier have one or more RF amplifiers tuned along with the local oscillator by a multisection, air-variable capacitor; they often achieved front-end selectivity better than many current factory-built radios.

In the 1950s, radios appeared having tunable IF strips, requiring separately tuned RF preselectors. The Collins 75A-4 uses slug-tuned circuits at both the input and output of an RF amplifier. These are mechanically linked to the tuning dial, resulting in continuous resonating of the two front-end circuits to the signal frequency without operator intervention. The Collins 51S-1 design is perhaps the ultimate example of this method; it has a *double-tuned* circuit ahead of the RF amplifier and another single-tuned circuit ahead of the first mixer. In effect, there are 30 separate triple-tuned, turret-selected RF circuits. The front-end circuits in the Drake 2B are capacitor-tuned and require peaking by the operator with a separate panel control. The Collins KWM-2 also requires operator tuning of the receiver front-end circuits, which are shared by the transmitter circuits, by means of a separate panel control labeled EXCITER TUNING.

The Signal/One CX7 also has a separate operator-tuned preselector, with only a single tuned circuit. The preselector in my homebrew radio is derived from the CX7 design, using a three-band, high-Q tuned circuit ahead of the RF amplifier. One difference here is the use of relays to switch the three circuits; this avoids the need for more band-switch sections. The relays are driven by a diode matrix driven by the same 12-V dc bandswitch wafer that selects the local oscillator. The three bands are switched by 10 relays–three relays for each of the three tuned circuits and one extra relay to tie in a “padder” for the 160-meter band. Coils and relays for these three bands are built on three separate plug-in boards. This greatly facilitates circuit adjustment, since some tweaking of the coils and capacitors may be needed to cover the specified ranges.

The circuits shown here may be easily adapted to an outboard preselector/preamplifier for any receiver, whenever additional gain and front-end selectivity is needed. Surprisingly, the idea of an operator-tuned preselector, while thought to be already archaic and obsolete when built into the homebrew radio nine years ago, has just recently been revived as a selectable feature in a top-end radio, the Yaesu FT-1000MP-Mark-V.

RF Amplifier

This amplifier is comparable to the stage called a preamplifier in many radios, except that it cannot be switched out. The RF amplifier is needed for proper functioning of the high-impedance preselector circuits; the front-end attenuator is used when reduced gain is appropriate. Using a 3N211, dual-gate VHF MOSFET, the amplifier is fed directly from the preselector circuits. The preselector networks provide voltage-step-up impedance matching for the MOSFET, which is essential for achieving the full-gain possibilities of the transistor. The MOSFET provides higher gain than a JFET (which was also tried) and thus better system noise figure.

The MOSFET has received little press lately; this is partly due to some difficulty in locating devices with wire leads for experimenters. Nevertheless, it is still widely used (in surface-mount form) in factory-built radios. With all the talk of strong JFET front-end circuits, owners of new radios may be surprised to find a number of MOSFETs inside the box. For example, the Yaesu FT-1000MP uses a dual-gate MOSFET preamplifier for the upper bands, a MOSFET second mixer, and about 20 more small MOSFETs in other RF circuits.

Receiver Mixer

The mixer uses push-pull, dual-gate MOSFETs. Currently, JFETs are more commonly used. This 1992 decision was influenced by a *QST* article by Ulrich Rohde, KA2WEU.[4] He wrote, “... a push-pull arrangement with N-junction FETs or dual-gate MOSFETs. Both give similar performance. The dual-gate MOSFETs have slightly more conversion gain, higher intercept points and higher isolation...” The generous use of attenuator pads helps to provide proper terminations for the mixer, filters and post-amplifier.

Frequency Calibrator

The 100-kHz marking oscillator, while not strictly required, is quite useful for simple checks of the preselector, crystal filters and passband settings. The JFET oscillator is followed by a squaring amplifier to produce harmonics. Output coupling is very light, simply by means of a wire placed close to the receiver input. This capacitive coupling favors the higher frequencies and helps to produce a signal level more uniform through the bands.

Local Oscillator

Fig 4 shows the local oscillator and injection-amplifier circuits. The oscillators use series-resonant, third-overtone crystals in a common-collector circuit. The often-seen inductor shunting of the crystal is not employed; frequency adjustment is readily possible without it.

Ten separate oscillators ensure freedom from instabilities induced by mechanical switches. The oscillator section of the bandswitch carries only +12 V from a secondary regulator powered off the regulated +15-V line. The selected oscillator is powered by +12 V from the switch. The diode at the output of the selected oscillator circuit is switched on by the transistor emitter current. Both the diode and the base-emitter junction serve to isolate the oscillator when it is not selected. A single resistor, R2, serves as a common-emitter resistor for each of the 10 oscillators; it draws current only from the oscillator in use. The same oscillator bandswitch wafer is used to generate a level-sensitive control voltage that may be used to select external amplifiers and antennas. Two stages are used in the local-oscillator injection amplifier.

The crystals chosen are of the highest grade obtainable; the ICM part numbers are given in the parts list. Each crystal frequency is 40 MHz above the low edge of the corresponding 1-MHz-wide band. Rather than a 41-MHz crystal for the 160-meter band, 41.8 MHz is used. There are two reasons for this. One is merely operator convenience: The band is tuned from 000 to 200 on the counter and this saves cranking the knob over 30 turns when going from the low end of the dial to the portion that would otherwise be used for 160. The second reason is more serious. A local oscillator at 41 MHz would allow considerable LO energy to enter the 40-MHz IF circuits; shifting to 41.8 MHz reduces this problem. The counter readings for 160 meters cause no trouble; DX work is done at the low end, so the readings are 000 to 040, and

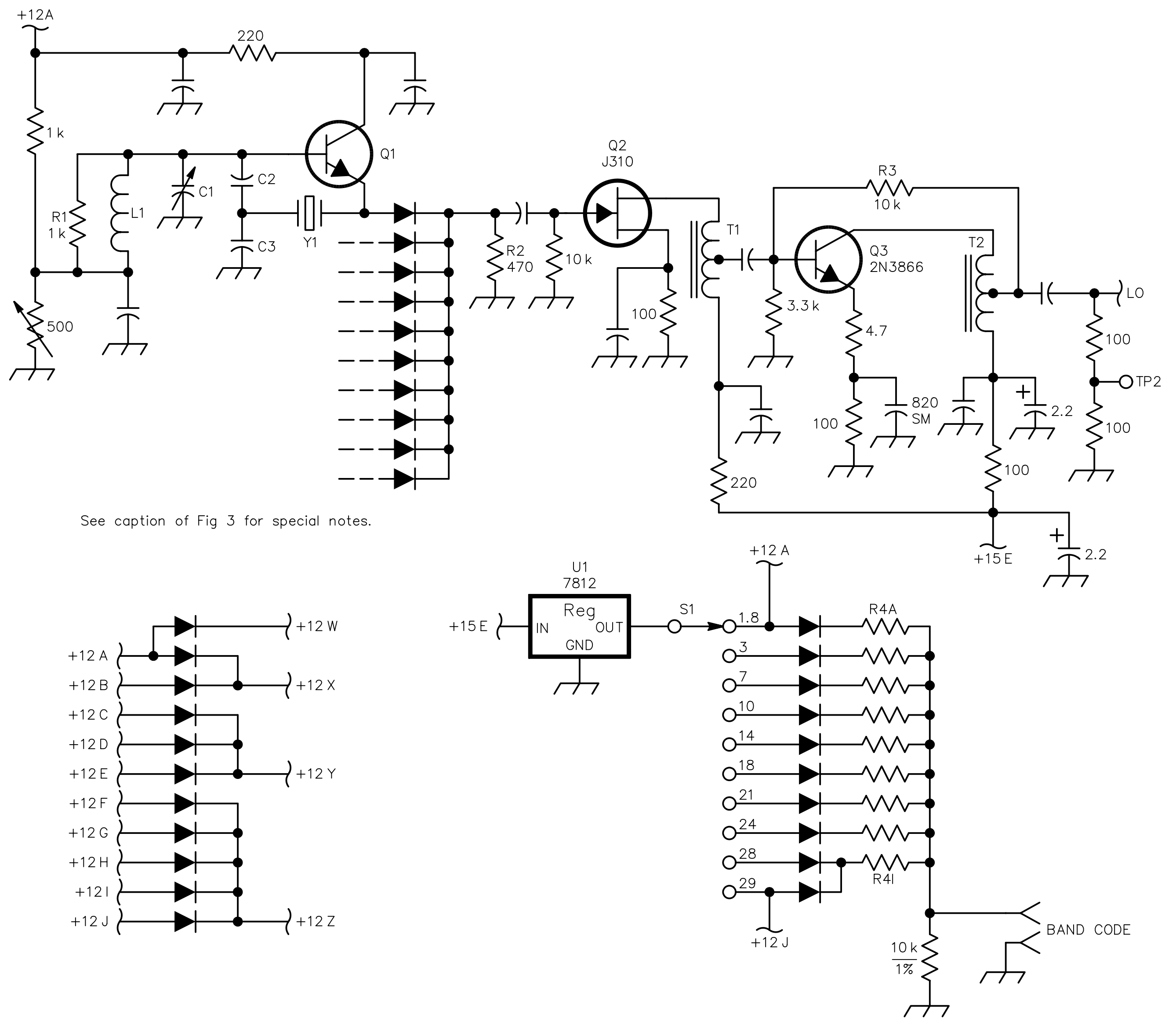


Fig 4—Local oscillator and injection amplifier. For general notes on the schematics, refer to the caption of Fig 3. Only one of the 10 separate oscillator circuits is shown. Certain component values for the 10 circuits are given in a sequence corresponding to the oscillator frequencies from 41.8 to 69 MHz.

C1—12 pF ceramic piston trimmer capacitor.
C2—Silver mica or NP0 ceramic: 39, 39, 33, 33, 33, 27, 10, 10, 10, 10 pF.
C3. Silver mica or NP0 ceramic: 150, 150, 120, 120, 120, 100, 39, 39, 39, 39 pF.
L1—Oscillator coil, 11 turns #24 enameled wire wound on R1, 0.225 inch ID, 10 required. Adjust or alter to 10 turns as required.
Q1—High-frequency bipolar transistor, f_τ rating of 600 MHz or higher, 10 required.MPS918, 2N918, 2N5179, BFY90 or similar.
R1—1 kΩ, 1 W, carbon composition, 10 required.
R4A-R4I. Nine resistors to generate the band-code output, 1% metal-film with the following values (in kΩ): 104, 47, 28, 18.5, 12.8, 9, 6.3, 4.3, 2.7. This results in band-code outputs of 1 to 9 V, for the nine ham bands.

S1—Bandswitch, 11 position phenolic wafer, 10 positions used. This wafer is mounted to the front panel, and ganged to the PA bandswitch. Ten lines, +12A to +12J, lead to the oscillators and to the diode matrix for the preselector; only two of these lines are shown.
T1—9:1 transformer, 8 trifilar turns on FT37-61 ferrite toroidal core.
T2—4:1 transformer, 4 bifilar turns on FT37-61 ferrite toroidal core.
U1—This IC voltage regulator requires bypass capacitors that are not shown; see Part 5, page 36, note 18. (Also see Note 3, here.)

Y1—Third overtone AT-cut crystal, ICM grade HA-5, ICM #471570 or #476570 above 65 MHz. The 10 crystals are cut for 41.8, 43, 47, 50, 54, 58, 61, 64, 68 and 69 MHz. For more convenient tuning, optional crystals cut for 43.5 and/or 64.5 MHz may be substituted; see text. Some savings in cost may be obtained by using grade CS-1 crystals, perhaps with nearly the same performance; ICM #471370 or #476370 above 65 MHz.

no operator confusion results. Similarly, to reduce VFO knob-cranking, a builder may choose to use crystals for 80 meters and 12 meters at 43.5 MHz and 64.5 MHz, although this will result in somewhat awkward counter readings.

Oscillator Adjustment

The procedure is the same for any oscillator using a third-overtone crystal. During adjustment, both the frequency and the oscillator's output level must be monitored. The frequency is measured at the LO amplifier output. Test point TP2 in Fig 4 provides a reduced level sufficient for a counter, without danger of overload. The counter probe is not allowed to encroach on the oscillator itself, where it may disturb the circuit, lowering the output and shifting the frequency. The bias trimpots in the oscillator circuits are individually adjusted to minimize collector current consistent with good starting and to obtain equal oscillator output levels on each band. The injection level at the receiver mixer is measured at test point TP1 in Fig 3, using the built-in RF probe. The nominal injection level is 1 V of RF. This is measured at test point TP1 as –1.0 V with a 10-MΩ dc meter, or –1.4 V measured with a 100-MΩ meter. If necessary, R3 may be altered somewhat to change the bias on Q3 and obtain the desired output.

Using the built-in RF probe is preferable to measuring injection level with a scope. Even a 10:1, low-capacity probe and a 100-MHz scope can cause a 3-dB

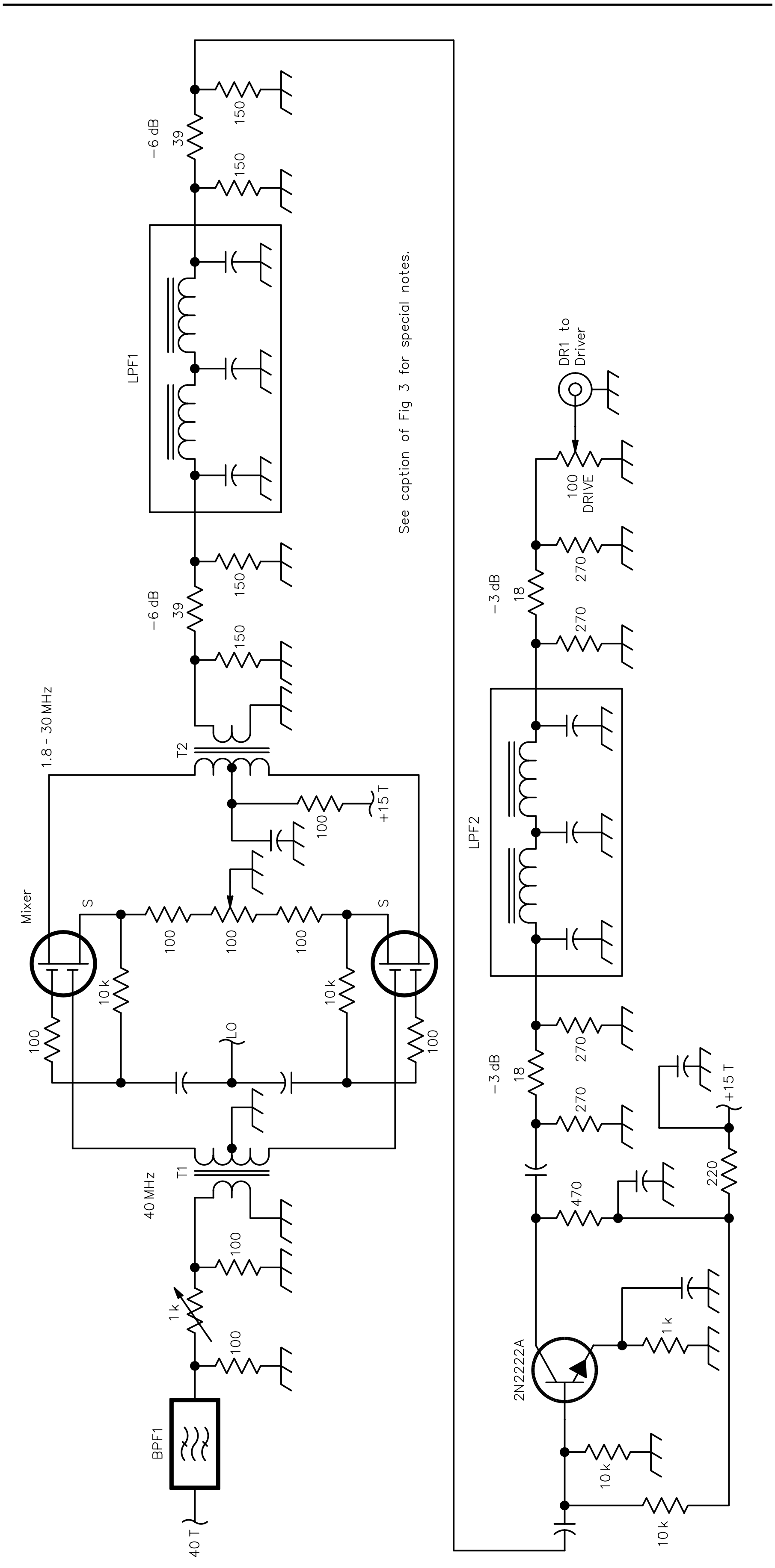


Fig 5—Transmitter circuits. For general notes on the schematics, refer to the caption of Fig 3.
BPF1—Same as BPF1 in Fig 3.
LPF1—Low-pass filter, 5-pole Butterworth, cut-off frequency 32 MHz. Each inductor is 400 nH, 15 turns on T37-12 iron-powder toroidal core. The outer capacitors are 62 pF silver mica; the central capacitor is 200 pF.
LPF2—Low-pass filter, 5-pole Chebyshev, cut-off frequency 30 MHz, ripple 0.5 dB. Each inductor is 326 nH, 11 turns on T25-10 iron-powder toroidal core. The outer capacitors are 180 pF silver mica; the central capacitor is 270 pF.
T1—Secondary: 12 turns on FT37-61 ferrite toroidal core, center-tapped. Primary: 2 turns over center of secondary. The turns-ratio is chosen not to match impedance or maximize gain, but to obtain the desired signal level at each mixer gate.
T2—Primary: 20 turns on FT37-43 ferrite toroidal core, center-tapped. Secondary: 3 turns wound over center of primary. This turns-ratio converts the 50 Ω load to 1100 Ω for each drain; this is a heavy load that contributes to good linearity.

loss in injection level when applied to gate two of a MOSFET. Any temptation to occasionally touch up the oscillator frequencies with only a counter should be resisted; the injection level at the mixers must also be watched. The built-in RF probe makes this easy. A slight change in trimmer capacity often results in a drastic change in the oscillator output level. An injection level much less than specified will degrade mixer performance. A moderately higher level is tolerable, though. The usual specification given for MOSFETs is 5 V(pk-pk); this is only needed for absolute-maximum conversion gain, which is not usually required. The level specified here is about 3 V(pk-pk); this is a compromise producing nearly maximum conversion gain, with less LO feed-through.

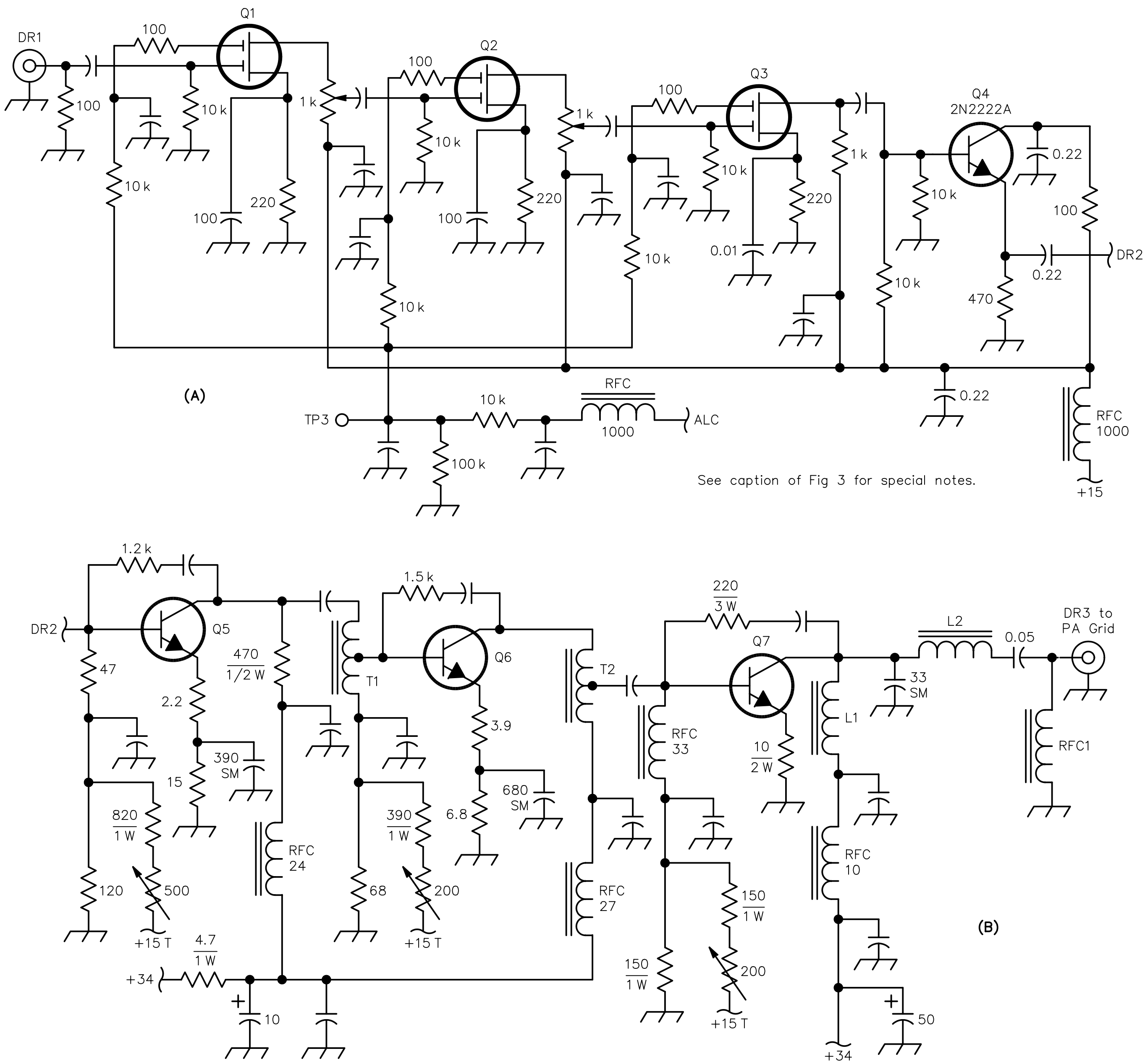


Fig 6—Driver circuit. For general notes on the schematics, refer to the caption of Fig 3. The bypass capacitors at all ALC lines are 1 nF. To ensure stability, the driver is built in two separate shielded compartments; the circuits are shown at (A) and (B). The limited bypassing of the source terminals in the first two MOSFET stages, with 100-pF disc ceramic capacitors, helps equalize the gain over the HF range. Complete bypassing of the third stage is necessary to provide adequate drive for Q4. The trimpots in the drain circuits of the first two stages are used to adjust for individual transistor gain variations and to distribute the gain equally between these two stages. The bias trimpots should be set initially to maximum resistance.

L1—100 µH, 43 turns #28 enameled wire on FT37-61 ferrite toroidal core.
L2—900 nH, 15 turns #24 enameled wire on T50-6 iron powder toroidal core.
Q5—2N3866; install a small snap-on top-hat style heatsink.
Q6-Q7—Motorola 2N5641, or similar stud-mounted, bipolar VHF transistor, dissipation rating of 15 W. (Available at Richardson)
RFC1—Output choke for protection, 82 µH, 14 turns on FT37-43 ferrite toroidal core.
T1-T2—4:1 transformer, 18 bifilar turns #28 enameled wire on FT37-43 ferrite toroidal core.

While adjusting the oscillator trimmer capacitor, the frequency will vary and the oscillator output level will have a peak. The oscillator should be set at the peak or on the *low-frequency side* of the peak—with the trimmer capacitor set to a greater capacity than at the peak. Settings on the low-capacity side of the peak may result in instabilities. A good crystal will usually have a frequency at peak output that is above the marked frequency, so adjustment at precisely the marked frequency is easily obtained. An adjustment within 100 Hz is sufficient. Although the resulting frequency-readout accuracy does not approach that of modern frequency-synthesized radios with TCXOs, experienced operators know that for serious DX work the ear is more important than the dial.

The coils should be Q-doped for stability. (Q-dope, a polystyrene coating specially formulated for RF circuits, is available from GC Electronics, #10-3702; see Table 4.) In the event that oscillation does not occur, the resonant frequency of the tuned circuit should be checked with a grid-dipper. Oscillation results when the feedback is of the proper phase. This occurs when the resonant tank frequency is *slightly above the crystal frequency*; individual coils may need to be trimmed accordingly. The bias trim-pots allow operation at the lowest feasible crystal current, dependent on individual crystal activity. This helps achieve high stability; very little frequency drift has been observed over nine years of all-day, every-day operation. Specifying crystals of the highest quality from a first-rate manufacturer has no doubt helped to produce this happy result.

Transmitter Circuits

The transmitter's mixer circuit is shown in Fig 5. After the singly balanced MOSFET mixer is a bipolar stage that delivers the HF signal to the DRIVE control on the front panel. The generous use of attenuator pads helps provide proper terminations for the mixer and the filters. From the panel DRIVE control, the signal is fed to the seven-stage driver. The resistive DRIVE control is crucial to keeping all stages operating well within their linear regions, without relying on excessive ALC voltage and without utilizing PIN diodes, which may degrade IMD performance. The low-pass filters serve mainly to reject the 40-MHz IF and the LO. Harmonics will also be generated in the PA and must be rejected there.

Driver Circuit

Fig 6 shows the driver circuit. The design of most of this section is derived from the CX7.[5] Three MOSFET stages provide moderate gain, but are used mainly for ALC. The 100-kΩ resistor on the ALC line performs no essential function; however, it is important during the construction phase. It is usual to construct a circuit such as the driver board as a separate module, and bench-test it before installation. During this time, the MOSFETs—with their unprotected gates—would be subject to damage if no return path to ground was provided for the gain-control gates. Even when the driver is installed, it is questionable what return path might be provided by the ALC circuit on the control board when the radio is powered off. The last three bipolar stages raise the level to about 500 mW to drive the grid of the PA tube. Thus, the 8072 tetrode stage has a gain of about 26 dB. The last three bipolar stages operate in class A at 34 V dc; this ensures good linearity.

For driving VHF transverters at HF, a take-off tap may easily be added at the output of Q5. ALC from the VHF units may then be applied to the HF section. Some factory-built radios do not provide an ALC function when used with transverters; this is one cause of splatter on the VHF bands. The driver must always be protected against high RF output voltages; this is done automatically by the grid ALC circuit and the PA-heater warm-up delay circuit. When the heater is off, though, the +15T line cannot be enabled. When driving a transverter at HF, provision should be made to enable only the portion of the driver that is used.

When an 8072 tetrode does fail, it is likely to do so in a rather unfriendly way. A short from screen to grid may apply a transient pulse from the +300 V line through the blocking capacitor to the driver. So, one may sadly find that in addition to the loss of the tube, there are several shorted transistors in the driver circuit. The RFC to ground at the driver output prevents damage from this transient. Two blocking capacitors are then required; one here in the driver and another at the PA grid.

The last three driver stages are biased for collector currents of about 30, 130, and 340 mA, respectively. Bias adjustments are made with the individual trimpots while measuring the emitter voltages: 0.45, 0.9 and 3.4 V. The bias-adjustment circuits include extra resistors in series with the trimpots. These extra resistors limit

Fig 7—Power amplifier circuit. For general notes on the schematics, refer to the caption of Fig 3. Only one of the seven separate fixed-tuned plate tank circuits is shown in the diagram. The bypass capacitor at the screen ring is connected as described in the text. There are no bypass capacitors connected to the screen pins on the tube base; the 560-Ω resistor serves to decouple the screen. The trimmer capacitors in the SWR bridge may be small ceramic types, but air trimmers are more likely to survive the occasional severe mismatch or atmospheric static discharge. Most of the PA components were salvaged from old CX7 radios.

C1—Tank tuning capacitors for the seven fixed-tuned ranges. 160 meters: fixed 100 pF, 7.5 kV transmitting, with 4-24 pF air trimmer in parallel. 80 through 15 meters: 4-24 pF air trimmer.

C2—Tank loading capacitors for the seven fixed-tuned ranges, 30-200 pF. Elmenco #304M mica trimmer, modified with four plates and double mica insulators.

C3—Tank tuning capacitor for the panel-tuned 24, 28 and 29 MHz ranges: 100-pF transmitting air variable, 3 kV, spacing 0.070 inch.

C4—Tank loading capacitor for the panel-tuned 12 and 10-meter ranges: Receiving air variable, 700 pF.

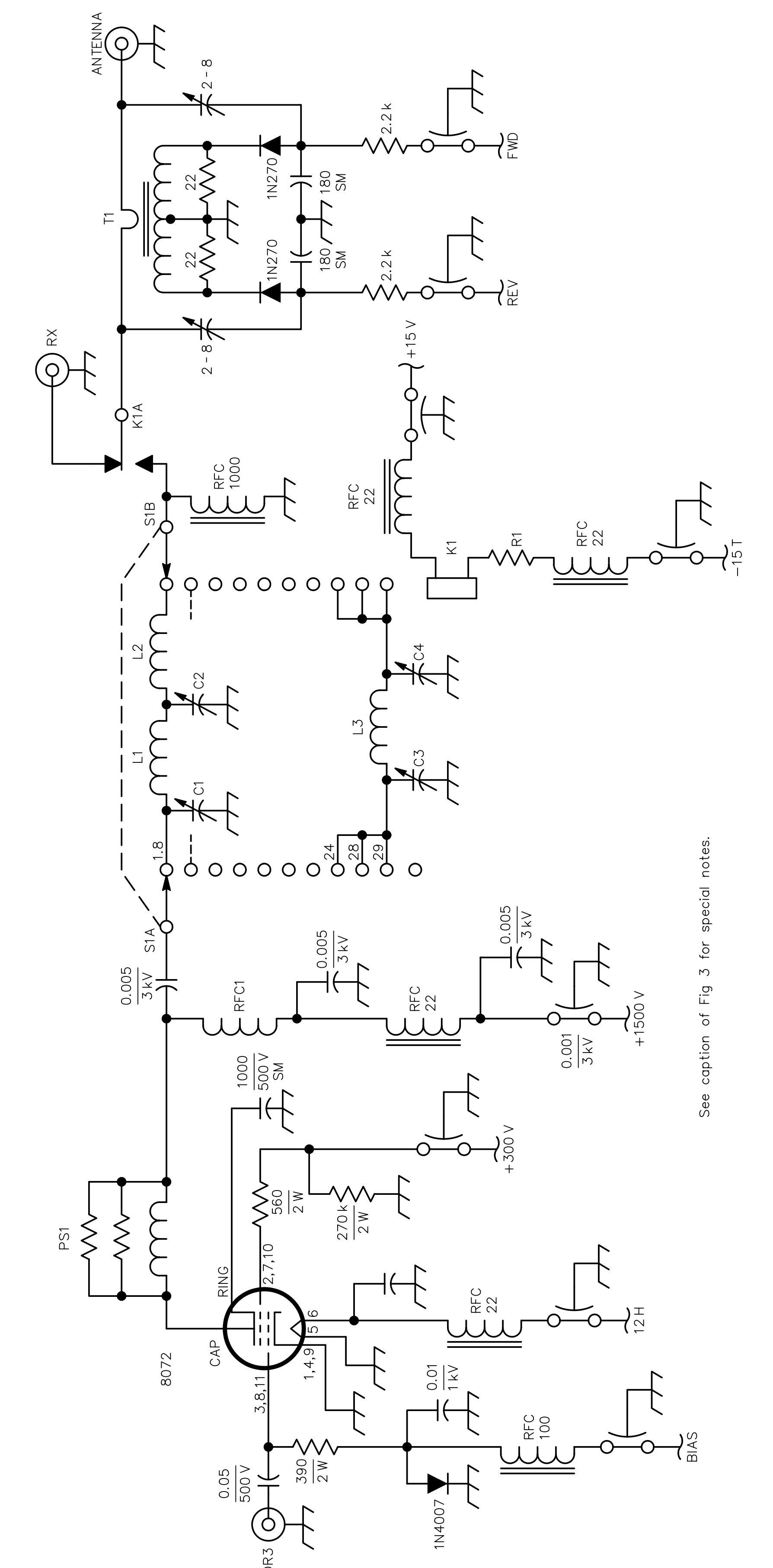
K1—Antenna TR relay. Kilovac type HC-1, 3.6 kV breakdown, 18 A current capacity, 6 ms operating time, 26.5 V dc, 80 mA coil. Jennings type RJ1A is identical; both are available from Surplus Sales of Nebraska. The reed relay from a CX7 may also be used with good results, but less reliability. The reed relay requires a higher coil voltage. The circuit allows easy modification for various coil voltages, by applying different voltages to the terminal labeled +15 in the diagram.

L1—Tank coils for the seven fixed-tuned ranges. Amidon T200A-2 iron-powder toroidal cores are used for 160 and 80 meters; fragments of type T200-2 are used for the next four bands (see text). The coils are wound with #16 enameled wire for the four lower bands, and with #14 for the other three bands. The ferrite cores are insulated with Teflon tape before the winding is applied. The 160 and 80-meter tank circuits are π-L networks; the other tank circuits are π-networks. 160 meters: 47 turns. 80 meters: 38 turns. 40 meters: 18 turns on 1/2 core. 30 meters: 16 turns on 1/2 core. 20 meters: 11 turns on 1/3 core. 17 meters: 10 turns on 1/4 core. 15 meters: 12 turns air-wound, 1.125 inch ID, 1.5 inch long. Depending on individual component characteristics, the coils may need adjustment or trimming to obtain resonance and proper loading.

L2—L-network coils for 160 and 80 meters. T200-2 iron-powder toroidal cores are used. 160 meters: 29 turns. 80 meters: 17 turns. L2 is omitted for the other bands.

L3—Tank coil for the panel-tuned 12 and 10 meter ranges: 6 turns #8 bare wire, 1.25 inch ID, 1.5 inch long. Silver plating is optional, but not likely to improve performance significantly.

PS1—Parasitic suppressor: 4 turns #14 bare copper wire, 0.25 inch ID, 0.875 inch long. Two resistors, each 68 Ω, 3-W metal-oxide, are connected to the coil with minimal lead lengths. It is best not to wind a parasitic suppressor coil over a resistor,



as the RF field induced in the resistor may cause unwanted effects.
R1—Select according to the coil requirements of the antenna relay installed. For a 12 or 48-V coil, instead of the +15 V connection, use ground or the +34-V line.
RFC1—Plate choke: 140 µH, 500 mA. The CX7 choke works very well, even on the new bands. Other chokes must be checked for series resonances.
S1—PA bandswitch, 2 ceramic wafers, 11 position, nonshorting. Only 10 positions are used. This switch must handle reasonably high RF voltages. The more common size, with 1.25-inch spacing between mounting studs, is inadequate. The surplus flea-market switch used here has 1.5-inch spacing. The "nonshorting" specification refers to the narrow width of the moving contact; this lessens the chance for arcing between contacts. The 12-position switch, which has very close spacing at the lug for the moving contact, should be avoided. The switch is driven by a shaft connected to the rear of the local-oscillator bandswitch on the front panel.
T1—FT37-61 toroidal core. Secondary: 10 bifilar turns. Primary: a single #14 wire, with Teflon sleeving, through the toroid.

the range of control and prevent inadvertent application of maximum bias voltage to a stage, with probable attendant destruction of the transistor. The resistors also permit the use of smaller-value trimpots and reduced dissipation. The last stage operates at about 10 W collector dissipation—a very sturdy class-A stage for delivering only ½ W to the PA. This helps maintain good linearity, since the PA's driving impedance changes when grid current begins to flow, and there are no tuned circuits at this point to provide a flywheel effect.

Output Power Level Control

The main transceiver panel provides a fixed transmitter output level of –7 dBm. Power-level control and ALC operation are effected separately at each of the three front-end sections. The panel adjustments at the front-end sections eliminate the need for any operator adjustments at the main transceiver panel when switching front ends. This is particularly convenient for instant switching between the VHF bands during a contest—a necessary operating feature for a top score. In this HF section, the transmit signal from the mixer circuit is fed to the panel DRIVE control at a level of about 50-200 mV (pk-pk) (depending on frequency), and then to the driver section at about 0-50 mV (pk-pk) (depending on the desired output power and the amount of ALC compression in use). For precise control of operating parameters, both a DRIVE control and an OUTPUT control are required. Each control affects the output power and the ALC functioning, but in very different ways. The OUTPUT control varies a voltage threshold in the ALC circuit, setting the level at which the ALC circuit begins to limit the transmitter output power. The power output may be varied from nearly 0 up to 200 W. On the other hand, the DRIVE control directly varies the actual RF signal level applied to the driver. The transmitter RF from 1.8 to 29.7 MHz is routed directly through the carbon DRIVE potentiometer. As the DRIVE control is advanced, the drive level first reaches the point at which the power output attains the level preset by the OUTPUT control; the ALC voltage begins to rise at this point. Further advancement of the DRIVE control does not increase the power output; the ALC circuits keep the output at the preset level.

This arrangement avoids a common problem that occurs when only an OUTPUT control is provided. In that situation, an excessively high ALC voltage is required to achieve low output power levels; this may be more than the ALC-controlled stage may be able to handle without distortion. This problem can arise in QRP operation, when driving a low-drive tetrode kilowatt amplifier, or when driving a VHF transverter. The OUTPUT control in this HF section is used for QRP operation, low-power contest operation at 100 W, when conditions allow reduced power or with a linear amplifier lacking an ALC provision. Normal operation is very simple: The OUTPUT control is left at maximum, and the DRIVE control is used to set the proper ALC compression level.

Automatic Level Control

The **DRIVE** control on the panel of the HF front-end section is adjusted for the correct amount of ALC compression, usually because of limiting of the 8072 grid current or for proper drive level to a kilowatt amplifier. An ALC control line runs from each kilowatt amplifier back to the corresponding front-end section, with ALC metering at the transceiver. The ALC control voltage is applied to dual-gate MOSFETs, as is common. An IMD problem can occur if excessive ALC voltage is applied to a MOSFET to obtain a large gain reduction. An extreme case occurs when an ALC line is improperly used to drastically reduce the power output of a radio to drive a low-input-level transverter. A proper drive-level adjustment will result in ALC compression of 2 to 3 dB—a moderate amount. A single MOSFET may be able to handle a 3-dB gain reduction without producing IMD, but to ensure the cleanest signal possible, the ALC voltage is applied to three cascaded MOSFET stages, so that each is required to operate with a gain reduction of only 1 dB.

Proper output and drive-level adjustment by the operator is easy. The OUTPUT control is set at the desired level. Then, while keying or speaking, the DRIVE control is adjusted for the optimum ALC meter reading of about –2 V (40%) on the main transceiver panel. The ALC meter reads signal level while receiving and is calibrated from 0 to 100 dB. When transmitting, the same meter indicates the ALC voltage from 0 to –5 V; the scale readings may be interpreted as percentages. The proper DRIVE control setting is not critical; any reading between 20% and 80% will produce a clean signal at full power.

PA Circuit

The PA's output tank circuit is adapted from the CX7 design. The CX7 provides a set of manual PA tuning and loading controls and an optional "broadband PA tuning" feature. This feature involves banks of preset internal tune and load-trimmer capacitors—merely a fixed-tuned arrangement. It worked well enough on the lower bands, but the poor L/C ratio on 10 meters made coverage of an entire 1-MHz segment difficult. In addition, coil-turn shorting and toroid-core losses resulted in reduced output on 15 meters.

The K5AM homebrew radio uses a variation of that scheme. There is no operating choice between manual or fixed tuning; the choice is built-in. There are 10 1-MHz bands to be covered. For the lower seven bands, seven separate fixed-tuned π- or π-L networks are used, with no provision for manual control. For the three upper bands (24, 28, 29 MHz), there is one manually tuned π network with front-panel controls, with no provision for fixed-tuned operation. Air-wound coils are used for 15 meters and up. This arrangement improves performance by avoiding the use of shorted turns on a single tank coil, by utilizing components that are more appropriate and by providing better L/C ratios on the upper bands. For contesting, the panel controls are tuned for 28 MHz; the result is equivalent to fixed-tuned, no-hands operation on all bands.

The PA circuit is shown in Fig 7. The design shown here may be easily adapted to an external amplifier for any low-power homebrew transceiver. The use of completely separate tank circuits simplifies the construction. The bandswitch requires only two wafers in the PA compartment and a single wafer at the panel to switch +12 V to the selected local oscillator and the diode matrix for the preselector. The 270-kΩ resistor installed close to the tube's screen terminal provides an important protective function. In the event of loss of screen voltage and an open, unloaded screen circuit, induced current from the plate could produce a high voltage at the screen terminal, damaging the tube and/or the screen bypass capacitor. The resistor drains off this induced current.

There are eight separate PA tank circuits; coils for the lower six fixed-tuned ranges use iron-powder toroidal coils or coils wound on portions of a toroidal core. The 15-meter, fixed-tuned range and the panel-tuned range for 12 and 10 meters use air-wound coils. Portions of a core were used because it was found that when only a few turns were wound on one side of a large toroid, nearly the same inductance could be obtained using only a piece of ferrite material slightly larger than the winding; this method also saves space. Alternatively, smaller-sized cores could be used. (The idea for using a fragment of a large core was provoked by an unscheduled event: accidentally dropping a toroid on the concrete floor of the garage shack.)

The use of an air-wound coil for 15 meters followed a few experiments. The CX7, while reaching 200 W or more on the lower bands, was notorious for dropping to about 130 W on 10 meters, and surprisingly, even lower on 15 meters, to little more than 100 W. It was usually thought that shorting turns on the large toroid with the bandswitch was the main reason. The method used here avoids that problem. Another cause of power loss was also found. When a ferrite core was tried for 15 meters, only 170 W was obtained, the same as on 17 meters, while the rig reached over 200 W from 160 to 20 meters. Further, the ferrite core became extremely hot, burning the insulating fabric. With the air-wound coil, the power on 15 meters increased to 190 W. The conclusion might be that type-2 iron-powder material (Micro-metals red) is unsuitable at this power level on 15 meters, and is not optimal on 17 meters. Type-6 (Micro-metals yellow) could be tried for these bands.

The panel-tuned circuit yields 180 W on the 12 and 10-meter bands.

SWR-Bridge Adjustment

The usual procedure is followed. First, the transmitter is set up to deliver about 100 W to a good dummy load on 10 meters. Then the trimmer capacitor associated with the reverse detector (closest to the tube) is adjusted for minimum voltage at the REV terminal. The input/output RF connections and REV/FVD connections to the bridge are then reversed by reversing the entire bridge subassembly; the adjustment is then repeated. The bridge may be left in this final position. This completes adjustment of the bridge itself; further circuit adjustments on the control board are indicated below.

PA Protection Circuits

The 8072 is one tough bottle! I use a total of eight of these tetrodes. They were all obtained on the used/surplus market at very little cost; most are 25 to 30 years old. With proper protection circuits, failures almost never occur.

Screen Current Limiting

The crucially delicate part of the 8072 is the screen; screen protection is mandatory. Unfortunately, the Signal/One CX7 had no effective screen-protection circuit. Thus, the failure rate of the 8072 in the CX7 was quite high, earning the 8072 an undeserved notoriety. The protection circuit used here monitors the screen current and uses the ALC system to limit the drive level to the PA, holding the screen current to a safe level. At 300 V, a preset limit of 20 mA keeps the screen dissipation under 6 W, well below the manufacturer's absolute-maximum rating of 8 W. High screen current occurs under improperly light loading conditions. When the loading is set too lightly with the manual controls or when a faulty antenna provides an improper load, the ALC voltage rises and the transmitter output will be reduced, but the screen current will be held at the safe limit.

Grid Current Limiting

Grid current is limited by the ALC circuit. This is not so much to protect the 8072 tube, which is also rated for class-C service and can take 50 mA of grid current, but to ensure linearity. Lack of ALC in amplifiers is a major cause of splatter on the ham bands. The 8072 requires grid current of about 2-3 mA for full linear output. This means that the tube operates slightly into the class-AB_2 region. The CX7 grid-ALC circuit provides limiting at about 4 mA of grid current and is very simple and reliable. It is used here with only slight modification and with components selected for a grid current of about 2 mA. The circuit does require initial adjustment and further re-adjustment if the tube or bias setting is changed. This grid-ALC circuit should be easily adaptable to other tetrodes and other bias voltages.

The grid-ALC circuit is shown in Fig 8; it is installed on the control board. Using only a single transistor, the circuit includes provision for bias-voltage regulation and adjustment. The transistor operates as an emitter-follower voltage regulator. The *Idle* trimpot establishes a preset voltage at the base, and thus also at the emitter. This voltage, usually about –20 V dc, is the operating bias for the PA grid. After the idle adjustment is made for 100-mA plate current, the *Balance* trimpot is adjusted to obtain 0 V at the collector, test point TP4. When the grid draws current under peak RF drive conditions, nearly all this current appears at the transistor collector. This current increases the voltage drop in R1 and drives the collector negative. For example, a grid current of about 2.2 mA will cause an increase of 7.2 V in the voltage drop across R1. The collector is thus driven from 0 to –7.2 V. The Zener diode and the three diodes in the remainder of the ALC circuits have a total drop of about 6.2 V. The resulting –1.0 V on the main ALC line will begin to reduce the gain of the driver and to indicate on the ALC meter. The grid-current threshold at which ALC action begins can be changed by changing the Zener diode. When not transmitting, or between code elements, the **KEY1** line is open, and the full –60 V from the bias supply is applied to the grid as cut-off bias.

Heater Warm-Up Delay

The 8072 heater requires a 60-second warm-up period before plate current may be safely drawn. The classic glass vacuum time-delay tube is still available at rather inflated prices and is only infrequently found at flea markets at bargain prices. The easiest solution, as for most timing problems, is an op-amp timer. The warm-up circuit is on the control board. Transmitting is inhibited until about a minute after the 8072 heater is switched on.

Heat-Sink Fan Timer

The 8072 PA heat sink does not require a fan; but in the interest of long service life, it is best to keep equipment as cool as possible. In high-duty-cycle contest operation, or on AM or RTTY, the heat sink can become quite hot if a fan is not used. A small, inaudible 120-V ac "muffin" fan keeps the heat sink fairly cool. A timer turns on the fan at the start of each transmission, and keeps it running for several minutes after the end of the transmission. The result is that during a contest, the fan runs continuously. The fan timer circuit is on the control board. Another muffin fan, also inaudible, is installed directly above the tube; it runs continuously whenever the heater is lit. In receive mode, this exhaust fan re-

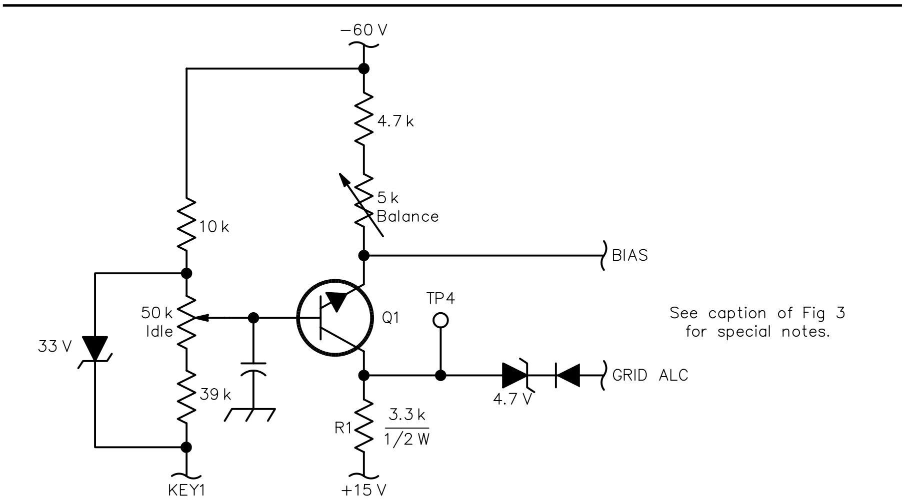


Fig 8—Grid-ALC circuit. For general notes on the schematics, refer to the caption of Fig 3. This circuit allows a certain amount of grid current to flow before developing ALC voltage to limit the drive level; see text. The circuit may be adapted to any small tetrode. Q1—MPS-A42, or similar small NPN transistor with a voltage rating of at least 150 V.

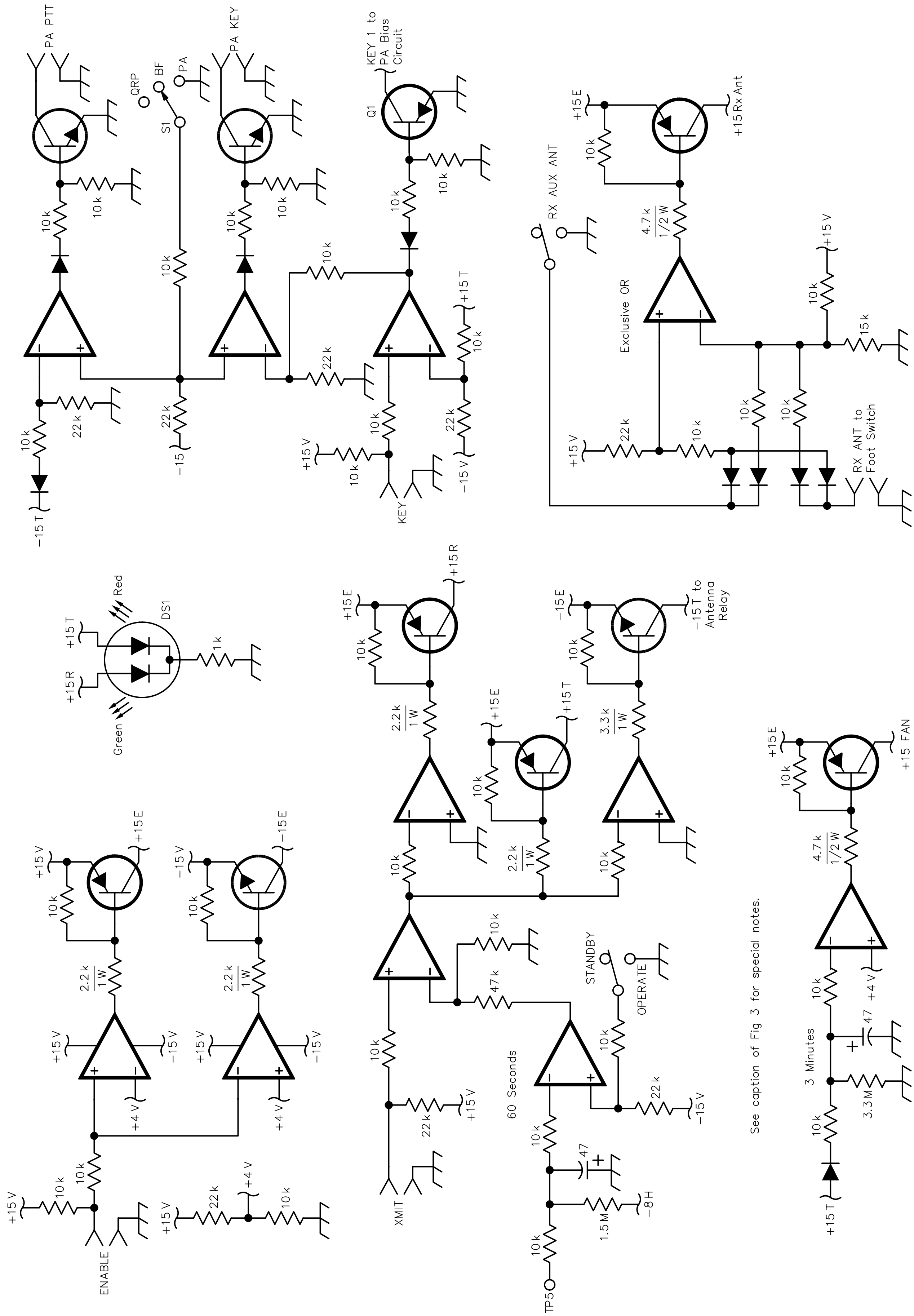

PA PTT
PA KEY
QRP
BF
PA
S1
Q1
KEY 1 to PA Bias Circuit
KEY
RX AUX ANT
Exclusive OR
+15 Rx Ant
RX ANT to Foot Switch
DS1
Red
Green
ENABLE
XMIT
60 Seconds
STANDBY
OPERATE
TP5
−15 T to Antenna Relay
+15 FAN
3 Minutes
See caption of Fig 3 for special notes.

Fig 9

Fig 9—Control circuits. For general notes on the schematics, refer to the caption of Fig 3. Except for the two op amps in the *Enable* circuit, the op amps on this board are all powered from the +15E and –15E lines.
DS1—Two-color LED, 3 leads.
S1—Rotary switch; 2 pole, 3 position. The other section, for QRP operation, is shown in Fig 11.
Q1—MPS-A92, or similar small PNP transistor with a voltage rating of at least 150 V.

moves heat generated by the tube heater. In transmit, it helps cool the anode.

Control Circuits

Many functions are carried out by the control board. One is special to the arrangement of this homebrew station, with a 40-MHz transceiver and three front-end sections. The front ends must be capable of quick change, with just one button on the operating bench and no cable changes. This is managed by the front-end relay box described in Part 5 (see Note 3). Aside from the coax relays involved, the key to the system is the *enabling* feature built into each front-end. The single control line for this purpose is termed the ENABLE line. When grounded, it enables the selected front-end to function; otherwise all circuits are disabled. When the HF section is enabled, the LED above the STBY/OPER switch lights up green. Then, when transmitting, the same LED turns red.

Other control lines from the front-end switch box are for XMIT and KEY to control the front-end section, and for ALC metering on the main transceiver panel. Although keying and timing of the keying waveform are done at 40 MHz on the main transceiver panel, the 8072 tetrode in the HF section is also keyed. The HF section serves also to control the kilowatt amplifier. The panel switch labeled QRP/BF/PA, selects QRP operation, barefoot operation at up to 200 W from the HF section or kilowatt operation using the external amplifier. In each of these modes of operation, the OUTPUT control on the panel sets the desired power output level. The control circuits are shown in Fig 9.

Enabling and TR circuits

In addition to enabling by the front-end relay box, the HF section has an OPERATE switch that enables the transmit circuits. The STANDBY position of this switch is very useful for testing memory keyers and computer CW or voice features, without transmitting QRM. The STANDBY mode allows the main transceiver panel to remain in *Operate*, so the computer interface, voice levels, clipping, sidetone and so forth, can all be checked out and adjusted without transmitting a signal.

When the HF section is enabled, the +15E, –15E, and +15R lines are energized. If the OPERATE switch is engaged, and the XMIT line is closed by circuits on the main transceiver panel, then the +15R line drops out, and the +15T line is energized. The control signals between the various units are all carried by two shielded cables: one from the front-end relay box to the HF section (or, alternatively, directly from the main transceiver panel), and one from the HF section to the kilowatt amplifier. In addition, phono jacks permit separate connection to each control line; these phono jacks are normally used only during tests on the workbench.

Auxiliary Receiving Antenna Control

To enable the reverse-selection, foot-switch function described above, an *exclusive-or* circuit is used, with diodes and an op amp. The auxiliary receive antenna is connected whenever one, but only one, of the two switches, panel switch or foot-switch, is closed.

Control of External Kilowatt Amplifiers

Although the kilowatt amplifiers in my shack have the usual STANDBY /OPERATE switches, each front-end section includes a switch to select either barefoot or amplifier operation. This is easily accomplished because control of each amplifier is provided by its associated front-end section, rather than by the main transceiver panel. If the QRP/BF/PA switch on the HF panel is set to PA then the external PA PTT and PA KEY lines are closed whenever the XMIT and KEY control signals are received from the main transceiver panel. The control board in each front-end section includes circuits connecting to the PTT, KEY and ALC circuits in the HF amplifier. The external kilowatt amplifiers are keyed along with the transceiver; thus, plate current is reduced to zero between CW code elements. This reduces average plate dissipation considerably.[6]

The main transceiver panel provides an XMIT control signal only for the selected front end. This means that selection of the appropriate amplifier is automatic. In more typical situations, this arrangement would mean that the radio controls the transverter and the transverter controls its associated amplifier. One button on the operating bench instantaneously switches the entire station from HF to 50 or 144 MHz. Further, the main panel BAND switch option can be used so that merely switching PTOs will automatically switch front-end sections and amplifiers. A typical VHF contest set-up with the two PTOs is for 50 MHz and 144.200 MHz. While running stations on six meters, one light touch on the MON button above the main tuning knob is enough to check for propagation on two meters. During sunspot-maximum periods, PTO A can be used with the HF section for DX chasing, with PTO B set for 50.110 MHz. One touch on the MON button is enough to check for that elusive six-meter, F_2-layer propagation.

ALC Circuit

Fig 10 shows the main ALC circuit; it is installed on a portion of the control board. This ALC method could be adapted to a variety of amplifiers. The ALC-combining op amp U1 is driven by five ALC signals: the output-power limiting circuit, the reverse-power (SWR) limiting circuit, the screen-current protection circuit, the grid-ALC circuit and the external kilowatt amplifier. The 10-µF capacitor near the combining op amp holds the ALC voltage between syllables; the decay time constant is 200 ms. Each ALC input is derived from a low-impedance source, so the attack time is very short. After the five ALC signals are combined, the result is applied to the three ALC-controlled stages on the driver board.

Op amp U1 is a voltage follower whose output corresponds to the highest of the five applied inputs. Usually only one of the ALC inputs hits its limit first and assumes control of the drive level. For example, my tetrode kilowatt amplifier requires a drive power of only 30 W; thus, the amplifier ALC controls the HF section, and the other four ALC circuits never get near their limits. When running flat-out barefoot with a good antenna, grid ALC is the controlling factor.

Adjustment is easy. The *Bias* trimpot is first set to provide a no-drive level of +2 V at the MOSFET gain-control gates (G2), test point TP3, in the driver section of Fig 6. This puts the operating point for the MOSFETs at the knee of the gain curve; see Fig 10 in the AGC article (Note 2). The *ALC Adjust* trimpot is then set so that 10 dB of

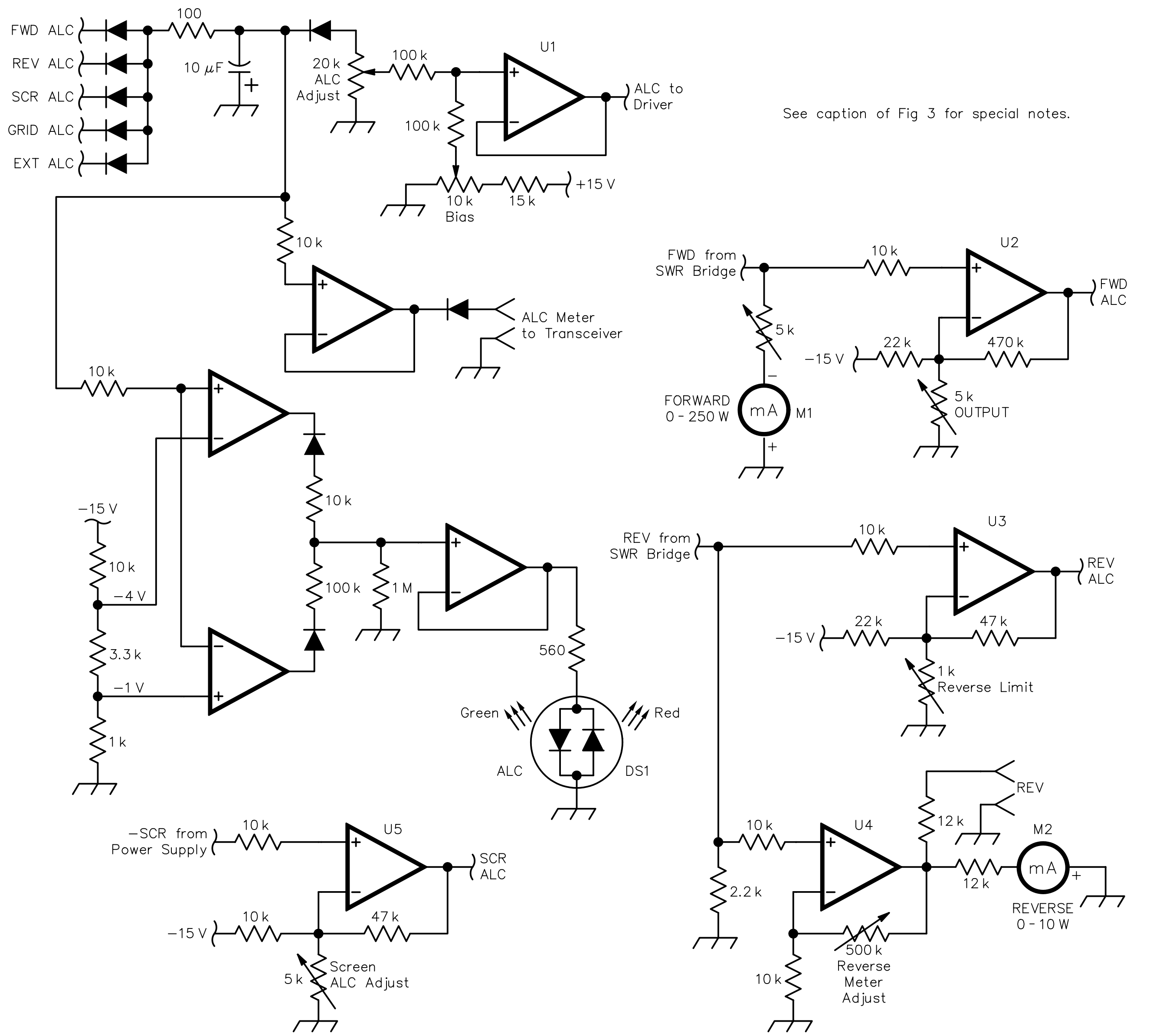
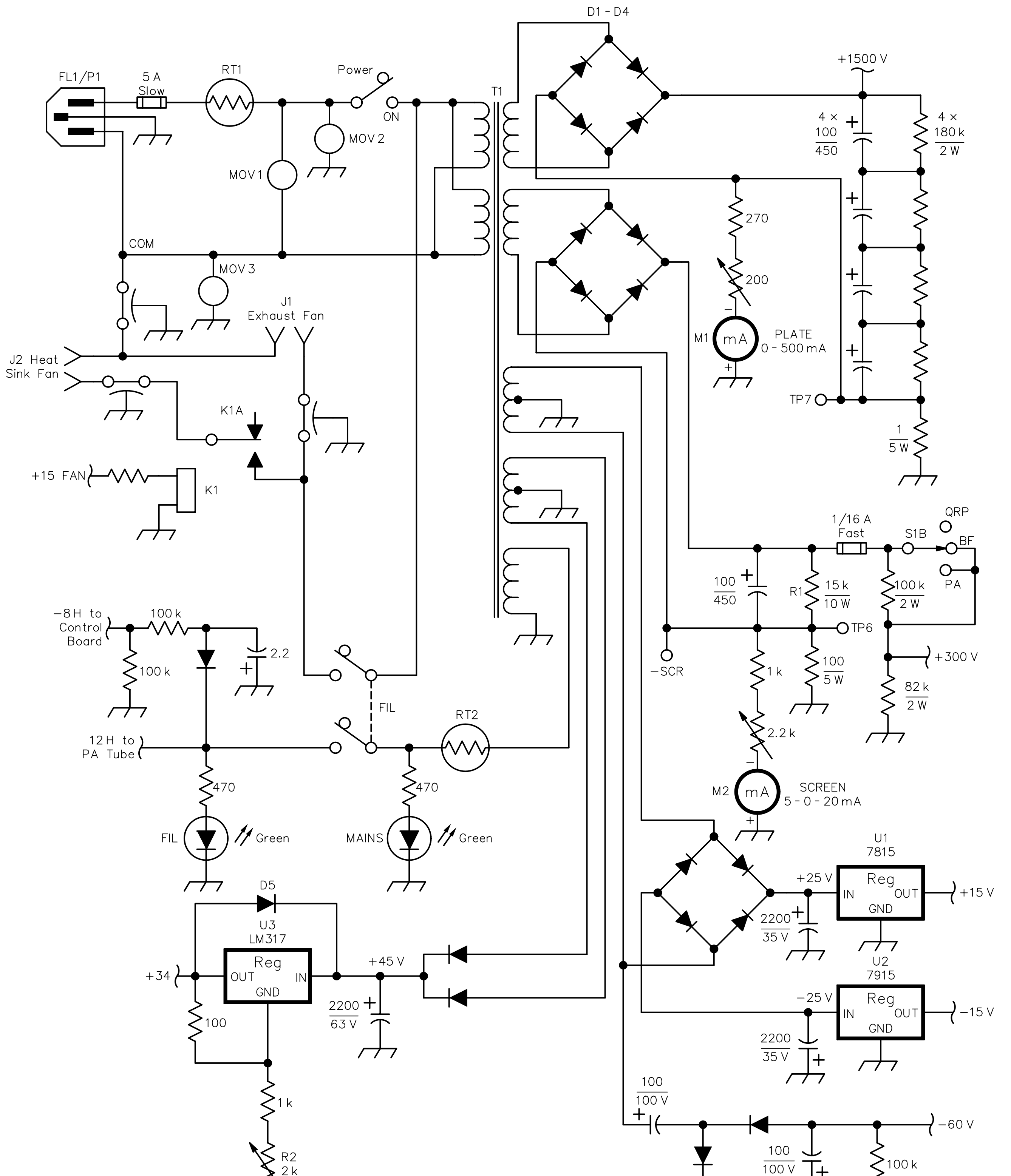


Fig 10—Main ALC circuit. For general notes on the schematics, refer to the caption of Fig 3.
DS1—Two-color LED, two leads; glows green or red, corresponding to a positive or a negative current.
M1-M2—Milliammeter, 1 mA, 100 Ω

Fig 11—(right) Power-supply circuit. For general notes on the schematics, refer to the caption of Fig 3. Diodes are type 1N4007, except as noted. The IC regulators are style T0-220, bolted to the rear panel as a heatsink.

D1-D4—1 A, 3 kV PIV
D5—This diode protects the IC regulator from damage by preventing the electrolytic capacitor on the driver board from discharging into the IC.
FL1/P1—Line filter with IEC ac chassis connector, 6 A.
J1, J2—Standard ac fan connector, with fan cord .
M1-M2—Milliammeter, 1 mA, 100 Ω.
MOV1-MOV3—Metal oxide varistor (MOV) transient protectors, GE type V130LA10A, MO#570-V130LA10A.
R1—This resistor draws more current than required for a bleeder; its purpose is to help stabilize the screen supply.
R2—Preset to minimum resistance to prevent damage to the driver transistors before adjustment.
RT1—In-rush limiter for ac line, Keystone Thermometrics CL-60, 10 Ω cold, 5-A operating current, MO#527-CL60.
RT2—In-rush limiter for PA heater, Keystone Thermometrics CL-180, 16 Ω cold, 1.7-A operating current, MO#527-CL180.
S1—Rotary switch; 2 pole, 3 position. The other section, for control of the external kilowatt amplifier, is shown in Fig 9.
T1—Power transformer; see text and Note 8.
U1-U3—These voltage regulators require bypass capacitors that are not shown; see Part 5, page 36, Note 18. (Also see Note 3, here.)



See caption of Fig 3 for special notes.

compression is obtained when the ALC meter reads –4 V (80%). One way to measure the compression is with a dual-trace scope at the input and output of the first section of the driver. There is a bit of interaction between the two trimpots; but it only takes a few minutes for the one-time adjustment of this circuit. The ALC begins to activate when the meter indicates –1 V and the ALC LED lights up green. At –2 V, the optimum operating point, the ALC compression is only 2 dB. A test can be run and an ALC-meter calibration chart drawn up; the exact correspondence between the ALC voltage and the compression level will depend on the individual MOSFET characteristics. For the ALC meter circuit in the transceiver, see Fig 9 in the AGC article (Note 2).

Power Output Control

The OUTPUT control on the panel sets the power output. When the set limit is reached, a control voltage is fed to the main ALC circuit. This ensures a definite, fixed output for QRP contest operation, or for driving a linear amplifier that does not provide ALC. Amplifiers without ALC are a major cause of splatter on the ham bands. The output level sampling is done by the directional coupler. From the same coupler, a sample of reverse power level is fed to the control board and used to cut back the drive when high SWR is sensed.

A jack labeled REV on the rear panel provides a reverse power sample for a remote meter; this is used for adjusting a remote antenna tuner. The radio is set for CW QRP operation at 5 W on a clear frequency, usually on a lower band in the daytime with little propagation. A shielded cable for the keying line and the REV line runs to the remote tuner at the base of the vertical antenna. The remote tuner includes a switch for the key line and a 1-mA meter for SWR indication. After initial tuning of a new vertical antenna at the center of each of the 160, 80 and 40-meter bands, a homebrew automatic tuner keeps the antenna tuned for 1:1 SWR throughout the band.[7]

The *Forward Power ALC* circuit uses op amp U2 in Fig 10. The OUTPUT panel control sets the threshold at the inverting input. When the voltage from the forward detector in the SWR bridge —as applied to the non-inverting input—reaches this level, the op amp's output begins to go negative. The feedback is chosen to provide enough gain to keep the transmitter's output within a few percent of the selected level. The *Forward Power* meter is driven directly from the detector in the SWR bridge; adjustment of the trimpot is done using an accurate external meter.

The *Reverse Power ALC* circuit uses op amp U3 in Fig 10. The circuit is similar to the one for forward power, except that the threshold is set internally by a trimpot on the control board. Op amp U4 is used to amplify the reverse detector's output to drive the meter to a full-scale reading of 10 W. To adjust the circuit, an external meter is used with a not-quite-perfect antenna to establish 10 W of reverse power, and the *Reverse Meter Adjust* trimpot is adjusted for a full-scale reading. Then the *Reverse Limit* trimpot is adjusted so that the ALC meter begins to indicate.

The reverse-meter circuit exemplifies the advantages of op-amp drive for meters. Not only can almost any meter be adapted to almost any requirement, but also op-amp drive can provide foolproof meter protection. In this circuit, the op amp's output is at most about –14 V. With the 12-kΩ meter resistor, the maximum meter current in any situation will be 1.16 mA; the meter can never be damaged (see Note 6).

The *Screen ALC* circuit utilizes the screen-current metering shunt in the power supply. A current of 25 mA through the 100-Ω shunt produces –2.5 V at test point TP6 in Fig 11 and results in a full-scale meter reading. The sense voltage from the shunt is also applied to op amp U5 in Fig 10. A –2.5-V threshold at the inverting input is set by the *Screen ALC Adjust* trimpot. Allowing for a 5-mA standing current on the +300-V line, this corresponds to a screen current limit of 20 mA. The *Grid ALC* circuit, shown in Fig 8, was discussed above.

The heater-warm-up-delay timer inhibits the PTT circuit until the tube is sufficiently warm, about 60 seconds after the heater is switched on. The fan timer circuit keeps the heatsink fan running about 3 minutes after each transmission. Both the fan timer and the warm-up delay timer use op amps, only a quarter of a quad package for each. Op amps as timers often result in more compact circuits than those using the ubiquitous 555, at least when there is no high-output-current requirement. Also, op-amp circuits tend to be simpler and more straightforward; the two inputs permit convenient setting of the reference voltages.

Metering

The motto of my shack is: "You can never have too many meters!" There are 37 meters on the homebrew gear at my contest operating bench, not including factory-built gadgets or the boat-anchor bench. This HF section has four meters: plate current, screen current, forward power, reverse power. With the two meters on the main transceiver panel, these two units together constitute an HF radio with only six meters.

No meter switch is used; this allows quick, confusion-free readings without reference to a switch position or multiple scales. The plate current reads 500 mA full-scale. The operating bias is adjusted to obtain 100 mA idling plate current; this usually requires about –20 V at the grid. In receive mode or between CW elements, the tube is cut off by the full –60 V from the bias supply. At full drive, the plate current runs somewhere between 200 and 300 mA, depending on the tuning and loading adjustments and the condition of the tube. The 8072 screen current at idle usually runs slightly negative. To allow measurement of this, the meter is configured with a total range of 25 mA, indicating from –5 mA to +20 mA. This is done simply with an 82-kΩ resistor on the +300 V screen-voltage line in the power supply. Together with the 270-kΩ resistor at the tube, these draw a total standing current of 5 mA. In standby or idle, the screen-current meter reads 20% of full-scale, indicating 0 mA.

The forward-power meter is driven directly by the directional coupler; the reverse-power meter uses an op amp to read lower levels. Although the forward-power meter reads 250 W full-scale, the reverse-power meter reads only 10 W full-scale. This provides a very quick and convenient rough check of the SWR, a variation on the crossed-needle idea. When the two meters read the same—and the needles are observed to be parallel—the SWR is 1.5; this is usually considered a very good match. If the reverse indication is less than the forward, that's even better; if much higher, we may need to work on the antenna. This method does not give the exact SWR reading that a crossed-needle meter would give, but it does provide a quicker indication of satisfactory operation. The high sensitivity of the reverse-power meter is also useful for adjusting antenna tuners at an interference-minimizing level of 5 W.

ALC Indicator

A special feature, which might be considered a frill, is a panel ALC indicator using the two-color LED shown in Fig 10. The main transceiver panel has an ALC meter, always on line when transmitting, so an indicator on the HF panel is not strictly necessary; however, it is very useful! Its main use is for rapidly setting the proper drive level, especially when driving a kilowatt amplifier. For very rapid band changes, it is much easier to observe the LED than to read the meter. The LED glows either green or red. The circuit is arranged so that the LED lights up green when the ALC line reaches –1 V, and changes to red when it reaches –4 V. Anything between these levels is okay, so the interpretation of these colors will be obvious to any licensed driver. After the initial hurried drive-level setting, the meter can be used to set the optimum ALC level of –2 V; but by that time, you have probably already worked a new country!

The ALC LED is also useful when tuning the PA on the 12 and 10-meter bands. Tuning must be done with full drive, as for any linear amplifier. It happens sometimes that a linear amplifier is tuned for maximum output with low drive, and then the drive is brought up fully. The intention may be to reduce dissipation while tuning, to prevent damage to the tubes; but the result is not only reduced output, but also a severely inadequate loading adjustment—and splatter! An easy way to reduce dissipation at full drive is to tune up with CW dits at 50 wpm. Even better is pulse tuning at 33% duty cycle (see Note 6). With the ALC LED in a prominent position near the tuning controls and meters, it is easy to ensure that full drive is applied and maintained while tuning the PA for the higher bands.

Power Supply

Although the power supply section is routine, the number of various voltages required could present a problem in locating a suitable transformer. In addition to the usual +15 V and –15 V regulated lines, a +34 V regulated line is needed for the transmitter driver. In addition, the PA tube requires 12 V ac for the heater, –60 V for the bias/standby circuit, +300 V for the screen, and +1500 V for the plate. This problem was solved in the same way as the PA anode clamp and heat sink problem, with a used Signal/One CX7 power transformer. An exact replacement transformer is currently available.[8] The HF section circuits draw considerably less power than a complete CX7, so the transformer runs cool and should have a very long service life. The same transformer could be used to simplify power supply construction for any similar small tetrode amplifier.

The power supply circuit is shown in Fig 11. A 1500-V transformer secondary winding and a bridge rectifier provide +1500 V for the plate. A 300-V winding and a bridge rectifier provide +300 V for the screen. A 50-V center-tapped winding serves to simultaneously supply two rectifiers, for +25 Va and –25 V. Each of these rectifiers is a full-wave, center-tapped (FWCT) circuit using two diodes. The similarity in appearance to a bridge rectifier is deceptive; the clue is the grounded center-tap. These two lines supply the regulators for +15 V and –15 V. The bias supply, a voltage doubler providing about –60 V, is driven from one side of the 50-V winding. Finally, a FWCT rectifier supplies a +45-V line that powers the +34-V regulator for the transmitter driver. A separate switch controls the 12-V ac to the PA tube heater. This allows all-day-long receiver operation without keeping the tube warm, while waiting for the brave operators in the DXpedition boat to land on the island.

Plate and screen currents are measured with shunts in the negative leads. The meters may be safely calibrated with the radio power off. A small test power supply on the workbench is used to establish a current through a shunt: at test point TP6 or TP7 in Fig 11. The current is measured with an accurate, external meter. The trimpot is then adjusted for the correct reading. The safest method for working on the high-voltage supply is to connect an analog voltmeter, turn off the radio, unplug the line cord and watch the meter needle drop to zero.

The ac primary circuit has a full complement of protection features: an in-rush current limiter RT1 on the ac line, a slow-blow fuse and three metal-oxide varistor (MOV) line-surge protectors. In addition, in-rush current limiter RT2 on the 12-V ac line prevents thermal strain and tube damage during the heater warm-up period, even when the heater is switched on sometime after the main ac switch is turned on. Because of the way the panel lights are wired, the effect of RT2 is visually rather amusing. If the main ac switch has already been closed, then when the heater switch is closed, the ac line indi-

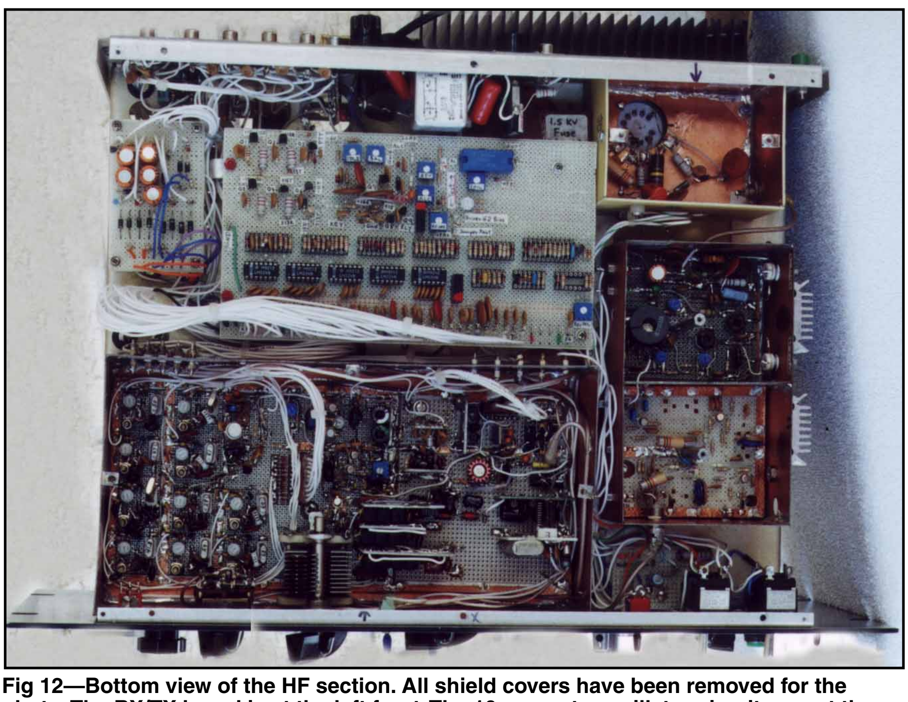


Fig 12—Bottom view of the HF section. All shield covers have been removed for the photo. The RX/TX board is at the left front. The 10 separate oscillator circuits are at the left side of this board. The variable capacitor at the front is for tuning the preselector; the three plug-in relay and coil boards are to the right. The two driver sections are at the right front; the stud-mounted transistors are bolted through the side wall into small heatsinks fastened to the outside of the wall. At the right rear is the PA grid compartment, completely shielded from the remainder of the circuit.

The control board is at the center rear; it is hinged at the left side for ease in servicing and modification. The part of this board using op amps is wire-wrapped. The part with discrete components is wired point-to-point. Each lead to the control board passes through two holes in the board, with its insulation, before connecting to its terminal; this provides strain relief and prevents broken wires.

cator light *goes out* for a few seconds! No need to worry. This is caused by the current drain of the heater and the voltage drop through the in-rush limiter before it warms up to operating temperature. This demonstrates the protection provided by the in-rush limiter. A similar effect is seen if the heater switch is already in the **ON** position when the main line switch is closed. When the PA-tube heater is switched on, a rectifier and filter on the heater line 12H produce the control line –8H to start the heater warm-up delay timer on the control board.

The schematic diagram for the screen supply includes the QRP section of the **QRP/BF/PA** panel switch described in connection with the control board. Tetrodes are real power hogs in idle condition. Even in normal operation, 150 W is a lot of idle power for a 200-W amplifier. For a QRP contest at 5 W output power, it would be bizarre to simply reduce the drive while running 150 W input power. The problem is solved by lowering the screen voltage for QRP operation. The idle current is reduced to zero, and 5 W output is obtained with less than 10 W of dc plate input power. This QRP switch is only for CW; linearity is destroyed without adequate idle current.

Construction

A major goal in construction is to ensure ease of access to all circuits. This will facilitate repairs, but it is mainly to allow convenient experimentation and modification. Fig 12 shows the bottom of the HF section. The main RX/TX board is built using a variation of dead-bug ugly construction. Two boards are used: Above is a pad-per-hole "perf" board, underneath is a solid-copper board forming a ground plane. The lower board is single-sided, copper-plated on only the lower side; the non-plated side insulates the ground plane from the upper board. The boards are bonded with copper foil grounds wrapped and soldered around the edge, wide copper foil ground strips running through the interior of the board, and wires through drilled holes. This provides a better ground plane than pad-per-hole boards with a perforated ground plane on the underside, and is more convenient than true dead-bug style. The pads provide very rigid mounting points and the resulting circuit is much more solid than with true dead-bug construction; this is important for stability in the oscillator section. Several portions of the circuit are built on small pieces of pad-per-hole board and soldered vertically onto the main board; this method conserves space, and is especially useful for added circuits and modifications.

The 8072 anode requires a solid brass clamp, a voltage-insulating, thermal-conducting beryllium block and a heatsink. These were all salvaged from an old CX7. The insulating block, which protrudes through the rear wall, is seen in Figs 1 and 13. The heat sink, mounted outside the enclosure, is seen also in Fig 14. The small cooling fan is mounted to the heatsink; a slight upward tilt works best. *Beryllium dust is dangerous*; *the block should not be drilled or machined in any way.* To avoid the need for locating the special conduction-cooling components, a type-8122 tetrode may be directly substituted; forced-air cooling must then be provided.

A tube socket is not used. Better air circulation and heat transfer by convection is achieved by using separate clip-on contacts for each required pin. While anode cooling is always taken seriously, adequate cooling for the seals at the base of a tube is sometimes neglected. The Eimac data sheet for the 8072 states: "Mounting should always be such that free movement of air past the base by convection is possible." The exhaust fan above the tube also helps to produce adequate air circulation for the base.

An equally valid reason for avoiding a chassis-mounted socket is that this prevents mechanical and thermal strain on the tube. This consideration is unique to a conduction-cooled tube, which has its anode rigidly clamped to a heatsink. The contacts are taken from a seven or nine-pin miniature tube socket. Heavy leads connect the contacts to pads cut into the edge of the opening in the copper circuit-board material. This allows shorter RF paths than a socket would allow. Only a single pin need be connected for tube elements provided with multiple pins. The screen is bypassed by a capacitor connected directly to the screen ring; this was not done in the CX7. One capacitor lead is simply soldered to the screen ring; the other lead is soldered to the ground plane. This solder-to-the-tube method may seem a bit less shocking when one considers that this top-quality tube may not need replacement for up to 20 years.

The transmitter driver is built in two separate copper compartments. The driver and the PA are located as far as possible from the local-oscillator section to minimize heat-related drift. The PA has separate compartments for the plate and grid circuits; this contributes greatly to stability. For the chassis, a surplus CX7 aluminum chassis was recycled. This may have been an economical choice, but it did not work out very well. It would be easier to construct a somewhat larger enclosure with LMB components, using the methods described for the main transceiver panel in Part 5 (see Note 3). A 7×17×14-inch enclosure with a seven-inch black rack

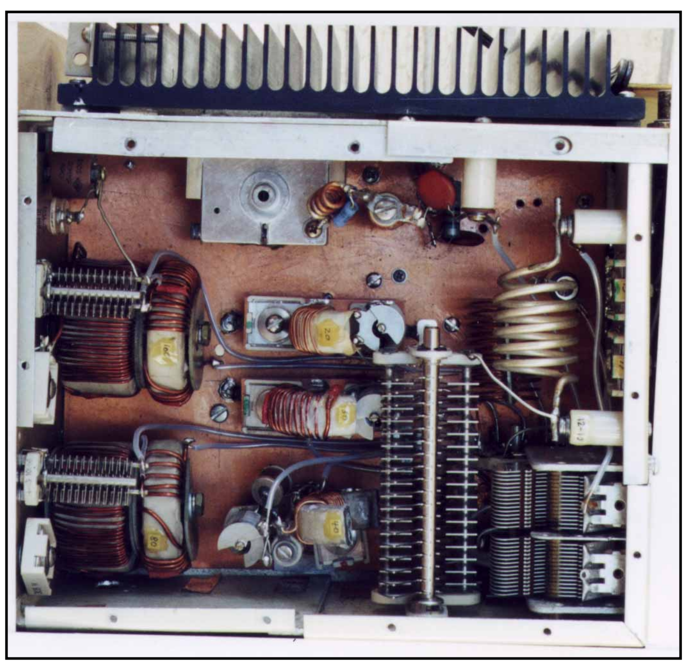


Fig 13—PA tank compartment. The large six-turn, silver-plated coil, the long, wide-spaced air variable tuning capacitor and the small two-section loading capacitor are used only on the 12 and 10-meter bands. The other seven bands utilize seven separate fixed-tuned tank circuits. The 160 and 80-meter toroidal coils are each mounted to the side wall with a single long brass threaded rod. A mounting bracket at the inside end (if not insulated) would create an effective *shorted-turn*, and seriously disrupt the functioning of the tank circuit.

panel should work well. All wiring is done with Teflon-insulated wire; RF is routed between sections with Teflon-insulated miniature coax, type RG-178B.

Alignment

It would be impractical to include complete step-by-step alignment details, especially since builders are likely to introduce modifications and improvements. Some special alignment notes have been included above; a few general remarks will be added here. At first, trimpots should be set to midrange or for minimum circuit current. To test the HF section on the bench without the transceiver, a shorted plug at the ENABLE jack is required. Another plug at the XMIT jack will allow the STBY/OPER switch to be conveniently used as a TR switch.

Most of the alignment, testing and adjustment can be done with the PA heater off. This avoids the bother of connecting a dummy load and allows the low-level transmitter sections to be tested and adjusted for hours on end without overheating the tube. Some precautions are still required, though. The heater-warm-up-delay circuit will prevent testing the transmitter sections. To defeat the warm-up feature when the heater is off, temporarily jumper test point TP5 in Fig 9 to –15 V. Now there is danger of damaging the output stage of the driver with excessively high RF voltage, since the grid-ALC circuit will not function when the tube is cold. To limit the driver output and test the grid-ALC circuit, temporarily connect a small diode from grid to cathode at the tube socket; this will simulate a warm tube.

For testing the transmit circuit only up to the second section of the driver without ALC, simply connect the base of Q5 to ground. To test the driver and grid-ALC circuit with the tube warm, but with no plate dissipation, output power or need for a dummy load, the screen fuse can be removed. For receiver tests, it is safer for the signal generator to use the auxiliary receive antenna jack RX IN.

The toroidal coils in the various filters should be measured and adjusted for proper inductance, since the permeability of individual cores may vary somewhat from published data. Some of the coils are quite sensitive to small changes in the position of the windings; they should be Q-doped to hold

Table 1—Receiver Sensitivity Measurements

Test results for minimum discernible signal (MDS) are given in dBm. To put the results in perspective, two other radios were also tested. The homebrew radio was tested at a bandwidth of 200 Hz. The Yaesu FT-1000MP (#9K470018) was tested at a bandwidth of 250 Hz using INRAD filters in both the second and third IF amplifiers, an INRAD first IF amplifier and an IF gain setting of 12. Measurements for the Collins 75A-4 (#2484) were taken using an 800-Hz mechanical filter.

	Receiver		
Frequency (MHz)	*K5AM*	*FT-1000MP*	*75A-4*
1.82	–142	–138	–143
3.52	–140	–138	–142
7.02	–133	–137	–141
10.12	–136	–138	–
14.02	–138	–139	–141
18.1	–144	–139	–
21.02	–142	–140	–143
24.91	–142	–143	–
28.02	–142	–144	–143

Table 2—Receiver Third-Order Dynamic Range Measurements

These tests were made at 14 MHz. The table shows dynamic range, DR3, at two-tone spacings of both 20 kHz and 2 kHz. The third-order intercept point, IP3, is also shown for both spacings. To put the results in perspective, two other radios were also tested. Tests were conducted for the homebrew radio at a bandwidth of 200 Hz. Tests for the Yaesu FT-1000MP were conducted at a bandwidth of 500 Hz using the stock filter in the second IF amplifier and an INRAD filter in the third IF amplifier; tests were also run at a bandwidth of 250 Hz. Tests for the Collins 75A-4 were conducted with the 800-Hz filter.

The measurement for the Yaesu at a bandwidth of 500 Hz and a two-tone spacing of 20 kHz is close to that reported in the ARRL product review.[10] For a two-tone spacing of 2 kHz, the measurement corresponds closely to that shown by the graph in the ARRL lab expanded report, provided the graph is adjusted so that its level at wide two-tone spacing conforms to the direct measurement.[10, 11] The tests show that the dynamic range of the Yaesu drops considerably when the two-tone spacing is reduced from 20 kHz to 2 kHz. This is because the roofing filter rejects interfering signals only at the wider spacing.

The homebrew radio uses no roofing filter; the dynamic range does not degrade drastically in the presence of nearby signals. During crowded band conditions, significantly better third-order IMD performance is obtained with the K5AM homebrew radio.

	20-kHz Spacing		*2-kHz Spacing*	
Receiver	*DR3 (dB)*	*IP3 (dBm)*	*DR3 (dB)*	*IP3 (dBm)*
FT-1000MP, 250 Hz	95	+4	75	–26
FT-1000MP, 500 Hz	92	+1	74	–27
K5AM	91	–5	88	–9
Collins 75A-4	72	–33	65	–43

the settings and retested. The completed filters should be tested before installation. The *Balance* trimpot in the receiver mixer-source circuit is adjusted for minimum LO feedthrough at 41.8 MHz; a setting near midrange indicates that the MOSFET pair is reasonably well matched.

Transmitter Alignment

The input to the transmitter-mixer circuit from the transceiver at 40 MHz is at a fixed level of –7 dBm, or about 280 mV (pk-pk). The trimpot should be adjusted to obtain a level of 100 mV (pk-pk) at each mixer signal gate; too much drive will increase spurious outputs. Even a 10:1 low-capacity scope probe at the gate will severely load the circuit. The required correction factor can be found by watching the signal level at the driver when the scope probe is touched to the gate. My probe caused a 3-dB drop, so the correction factor is 1.4 and a scope reading at the gate of 70 mV (pk-pk) is appropriate. The input level is easier to measure at the input transformer primary. The transformer has a voltage step-up of three times to each gate, so about 35 mV (pk-pk) at the primary is suitable. The *Balance* trimpot in the mixer source circuit is adjusted for minimum LO feedthrough at 41.8 MHz. The trimpots in the first section of the driver compensate for individual MOSFET characteristics; they are set to keep the gain in each of the first two stages roughly equal.

PA Alignment

The same procedure is followed for presetting the fixed-tuned circuits as for the panel-tuned circuit while operating. The best indication of proper loading conditions in a tetrode amplifier is screen current. The tuning control is always adjusted for peak screen current. In a stable amplifier, this should correspond exactly to maximum output and minimum plate current. The degree of amplifier loading is indicated by the level of screen current at this peak.

Different samples of the 8072 will develop maximum power at different peak screen-current levels. For most tubes, this will be between +5 and +10 mA. A tuning peak at 0 mA indicates excessively heavy loading and reduced output. A peak at +15 mA indicates excessively light loading. This again results in reduced output; but in this case, also the likelihood of distortion and splatter. For best linearity, loading should be adjusted slightly on the heavy side of the setting for maximum output. For example: If maximum output occurs at +9 mA, a loading adjustment that results in a tuning peak at +7 mA is best.

Performance

Complete performance measurements for the main transceiver panel and the three front-end sections must be deferred to a later article. For now, we report only a few measurements that relate specifically to the receiver and to the HF section. To put the data in perspective, measurements for a few other receivers are included. One of these, the Yaesu FT-1000MP, is used at my vacation cabin on Horse Moun-

Table 3—Receiver Second-Order IMD Measurements

The test signals are at 6 MHz and 8 MHz; the receiver is tuned to 14 MHz. The table lists the second-order intercept point IP2. To put the results in perspective, measurements for several factory-built radios are also listed. A number of older radios are included; they utilize a variety of elaborate and expensive RF tuning mechanisms, and it was of interest to see how effective they are with respect to second-order IMD. The table includes production dates when available. Measurements were made for the homebrew radio, the FT-1000MP, and the "boat anchors." Other data were taken from *QST* product reviews.[10]

Receiver	*IP2 (dBm)*	*Year*
Yaesu FT-1000MP-Mark-V, VRF on	112	2000-
Collins 51S-1	98	1959-72
Yaesu FT-1000MP	89	1995-2000
K5AM	83	1992
Drake 2-B	81	1961-65
Collins 51J-4	78	1952-62
Hammarlund HQ-129-X	77	1946-53
Collins 75A-4	74	1955-58
National HRO-5TA1	73	1944-47
Yaesu FT-1000MP, at RX IN jack	72	1995-2000
Hammarlund SP-600	70	1950-72
Yaesu FT-1000MP-Mark-V	69	2000-
Collins KWM-2	66	1959-75
Icom IC-756PRO	63	
Kenwood TS-570S(G)	59	
Ten-Tec Omni-VI-Plus	58	
Icom IC-718	55	
Yaesu FT-100	53	
Ten-Tec Pegasus	44	
Icom IC-706MKIIG	39	
Yaesu FT-847	15	

Table 4—Parts Suppliers

Amidon Associates, 7714 Trent St, Orlando, FL 32807; tel 800-679-3184, fax 407-673-2083; **tracy@bytemark.com**, **www.bytemark.com/amidon**.

Digi-Key Corporation, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677; tel 800-344-4539 (800-DIGI-KEY), fax 218-681-3880; **www.digikey.com**.

GC Electronics, GC/Waldom Inc, 1801 Morgan St, Rockford, IL 61102-2690, distributors nationwide; tel 800 435 2931, fax 800 527 3436; **www.gcwaldom.com.**

Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952-1158; tel 800-524-6464, fax 800-524-5414; **hosfelt@clover.net**; **www.hosfelt.com**.

International Crystal Manufacturing Company, Box 26330, Oklahoma City, OK 73126-0330; tel 800-426-9825, 405-236-3741, fax 800-322-9426, 405-235-1904; **www.icmfg.com/**.

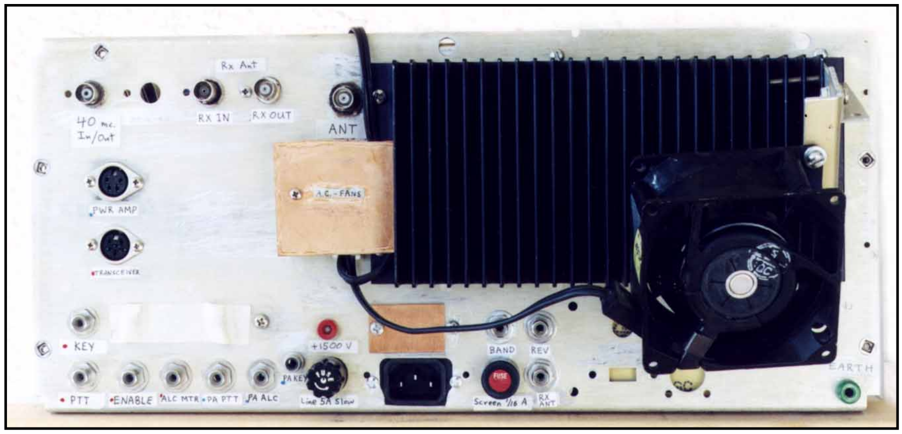
International Radio, 13620 Tyee Rd, Umpqua, OR 97486; tel 541-459-5623, fax 541-459-5632; **INRAD@rosenet.net**, **www.qth.com/inrad**.

Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX; tel 800-346-6873, fax 817-483-0931; **sales@mouser.com**; **www.mouser.com**.

Richardson Electronics, 40W267 Keslinger Rd, LaFox, IL, 60147-0393; tel 800-348-5580; **gloria@rell.com**; **www.rell.com**.

Surplus Sales of Nebraska, 1502 Jones St, Omaha, NE 68102; tel 402-346-4750, fax 402-346-2939; **www.surplusales.com**.

Fig 14—(right) Rear view of the HF section. The small copper box in the center covers feedthrough capacitors for the ac leads to the two small muffin fans. The two DIN connectors are for control cables to the main transceiver panel and to the kilowatt amplifier. This makes connections quick and easy. All control lines are also available at the phono jacks.



tain in New Mexico. Another, the Collins 75A-4, holds a pre-eminent position on the boat-anchor bench at home. Thanks to Emil Pocock, W3EP, for the suggestion to include the 75A-4, often declared by old-time operators to be the finest receiver ever built. This suggestion also led to the idea that I perform certain tests on several other old radios. The testing methods followed, as closely as possible, the procedures specified by the ARRL lab.[9] Results of the sensitivity tests for the various bands are given in Table 1.

Dynamic Range

Table 2 shows the results of third-order IMD dynamic range tests. The results for the Yaesu in the two-tone test at a spacing of 20 kHz demonstrate the effect of the roofing filter in the first-IF section. For the closer spacing of 2 kHz, the roofing filter provides no protection; the measurement then relates more directly to mixer performance and other factors such as diode switching. This problem was discussed in more detail in Part 1, page 18 (Note 3).

The K5AM homebrew radio uses a tunable first-IF section with no roofing filter. The dynamic range does not degrade drastically when the interfering signals are closer to the operating frequency. During crowded band conditions, this homebrew radio delivers significantly better third-order dynamic range performance.

Second-Order IMD

Second-order IMD test results are given in Table 3. There are wide differences between receiver designs that affect second-order IMD performance. For example, the Yaesu FT-1000MP uses eleven fifth-degree apolar Chebychev band-pass filters. A band-pass filter for the 12 to 15 MHz range is in use on the 20-meter band, where these measurements were made. In addition, the Yaesu uses three high-pass filters to reject signals in the f/2 region; these filters are included specifically to improve the second-order dynamic range. The results are excellent. The table also shows the results for the same radio using the RXIN jack. Unfortunately, this configuration bypasses the high-pass filters, and the performance is significantly degraded. Although measurements were not made on the 160-meter band, second-order IMD performance is also likely to be degraded there. This may be a serious consideration if a beverage antenna is used at the RXIN jack in areas with strong broadcast band signals.

The Yaesu FT-1000MP-Mark-V uses a selectable panel-tuned preselector called VRF; the results are exceptional, although there is a loss in sensitivity when the preselector is enabled. Another important design factor likely to affect second-order IMD performance is the use of diode switching in the front end of the factory-built radios. The high-pass filters in the Yaesu FT-1000MP are relay-switched and placed ahead of the diodes that switch the band-pass filters. No diode switching of signals is used in the K5AM homebrew radio.

My homebrew radio uses only a single, operator-tuned, high-Q LC circuit for preselection; the second-order IMD performance is adequate. High-pass filters could be easily added. Simple filters with only 20-dB rejection of signals in the f/2 region would raise the second-order intercept point by 40 dB.

Summary

The K5AM homebrew transceiver with the HF section has served faithfully for nine years, through many CW, RTTY and SSB contests and during countless hours of DXing. In a period of a little over a year, the radio worked over 100 countries on the 160-meter band (with help from an amplifier and a balloon). Thanks to the built-in protection circuits, there have been no breakdowns. The straightforward operating features, multiple meters and large knobs have contributed to highly enjoyable operating. The performance of the radio leaves little to be desired.

Readers are certain to notice points in the circuits where improvements are possible; please send in your ideas. This article completes the description of the homebrew transceiver up to 29.7 MHz. A subsequent article will describe the VHF sections.

Notes

1. M. Mandelkern, K5AM, "Evasive Noise Blanking," *QEX*, Aug 1993, pp 3-6.
2. M. Mandelkern, K5AM, "A High-Performance AGC System for Homebrew Transceivers," *QEX*, Oct 1995, pp 12-22. Corrections in *QEX*, Jul/Aug 2000, p 59.
3. M. Mandelkern, K5AM, "A High-Performance Homebrew Transceiver." Part 1, *QEX*, Mar/Apr 1999, pp 16-24 (General plan). Part 2, *QEX*, Sept/Oct 1999, pp 3-8 (IF board). Part 3, *QEX*, Nov/Dec 1999, pp 41-51 (RF board). Part 4, *QEX*, Jan/Feb 2000, pp 47-56 (AF board). Part 5, *QEX*, Mar/Apr 2000, pp 23-37 (Logic board, etc). Corrections in *QEX*, Jul/Aug 2000, p 59, and *QEX*, Nov/Dec 2000, p 60.
4. U. L. Rohde, KA2WEU, "Recent Advances in Shortwave Receiver Design," *QST*, Nov 1992, pp 45-55.
5. The main differences are these: In lieu of an RCA type CA-3028A differential amplifier IC for (sometimes erratic) ALC control, three MOSFETs are used. The keyed stage is not used; keying is accomplished at 40 MHz on the main transceiver panel. In lieu of a low-pass filter here, a second filter is used in the mixer section on the RX/TX board. The bipolar emitter-follower stage using Q4 is added. Separate bias circuits are used for each of the last three bipolar amplifiers, and each stage has a trimpot for bias adjustment. TR switching is applied to each of the last three stages. The choke protection circuit at the output is added. Instead of heatsinking the last two stages to the chassis, these stud-mounted transistors are bolted through the wall of the driver compartment to small exterior heat sinks on the side of the enclosure. This helps keep heat away from the local oscillator. The control labeled OUTPUT on the CX7 is a *Drive* control in the sense used here; the CX7 has no *Output* control.

6This method is discussed further in M. Mandelkern, "A Luxury Linear," *QEX*, May, 1996, pp 3-12, and "Design Notes for 'A Luxury Linear' Amplifier," *QEX*, Nov 1996, pp 13-20.

7M. Mandelkern, K5AM, "An Automatic, Remote Antenna-Tuning Controller," *QST*, Sep 1995, pp 46-49.

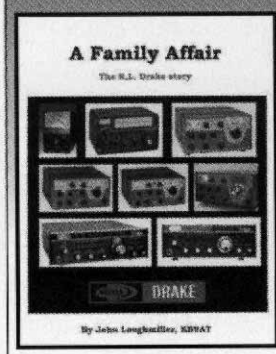
8The transformer is available as model #CX7 from Peter W. Dahl Co, 5869 Waycross Ave, El Paso, TX 79924; tel 915-751-2300, fax 915-751-0768; **pwdco@pwdahl.com**; **www.pwdahl.com**.

9M. Tracy, KC1SX, and M. Gruber, W1MG, *ARRL Test Procedures Manual*, Revision F, June 2000. This document is available as **www.arrl.org/members-only/prodrev /testproc.pdf** on the ARRL members' Web site.

10Product reviews and expanded lab reports are available at **www.arrl.org/members-only/prodrev** on the ARRL members' Web site.

11For a discussion of dynamic range versus two-tone frequency spacing, see E. Hare, W1RFI, "Swept Receiver Dynamic Range Testing in the ARRL Laboratory," *QEX*, June 1996, pp 3-12, 29.

A Family Affair
The R.L. Drake Story



• Brand new!
• Printed October 2000
• 23 Chapters
• 300 Pages
• 150 Photos
• Glossy four color cover
• Over 150 pages of radio mods.
• $29.95 (+$4.95 ship)

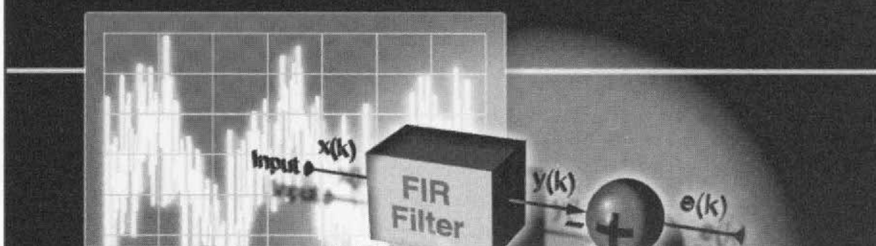
John Loughmiller KB9AT reveals the behind-the-scenes history of the famous R.L. Drake Company, focusing on the glory days, when Drake was king in amateur radio. Every ham and SWL knew R.L. Drake from the outside, but now the inside story of this incredibly interesting company is told. This book also includes 150 pages of useful circuits and modifications for many Drake amateur radios. An entertaining read and a great technical reference for every Drake owner.

universal radio inc.

Universal Radio
6830 Americana Pkwy.
Reynoldsburg, OH 43068
◆ Orders: 800 431-3939
◆ Info: 614 866-4267
www.universal-radio.com

NEW ARRL BOOK

Digital Signal Processing Technology

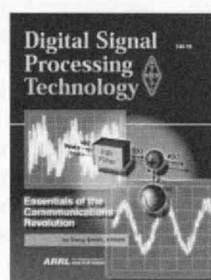


"The amateur interested in advances in DSP and communication processing can learn from the excellent presentation of this needed material."
Dennis Silage, PhD, K3DS,
Professor of Electrical & Computer Engineering, Temple University, Philadelphia, PA.

Contents:

Introduction to DSP
Digital Sampling
Computer Representations of Data
Digital Filtering
Analytic Signals and Modulation
Digital Coding Systems for Speech
Direct Digital Synthesis
Interference Reduction
Digital Transceiver Architectures
Hardware for Embedded DSP Systems
DSP System Software
Advanced Topics in DSP, and more...

Digital Signal Processing Technology
ARRL Order No. 8195 — **$44.95***



ARRL
The national association for AMATEUR RADIO

Order Toll Free
1-888-277-5289
Mon.-Fri. 8 AM-8 PM Eastern time
www.arrl.org/shop
Or, contact us to locate a dealer near you.
*shipping: $8 US (UPS) $10 International

QEX 11/2001

225 Main Street • Newington, CT 06111-1494 • USA
tel: 860-594-0355 fax: 860-594-0303 e-mail: pubsales@arrl.org

Wave Mechanics of Transmission Lines, Part 3: Power Delivery and Impedance Matching

As we conclude our tutorial, let's look at a wave-mechanical explanation of impedance matching. The analysis begins with a close examination of what happens at a match point. Discussion of vector addition of voltages emphasizes the importance of phase delays in determining steady-state conditions.

By Dr. Steven R. Best, VE9SRB

[*Editor's note: In this article, we use letters in* **boldface** *to indicate vector quantities. Italic letters indicate scalar variables.*]

In my two previous articles,[1, 2] I presented basic concepts associated with wave reflection mechanics in a transmission-line system. I developed conceptual and mathematical relationships relating the steady-state forward and reflected voltage, current and power traveling in a transmission line to the forward source voltage, current and power applied to the transmission line by a transmitter. It followed that all the transmission-line's steady-state conditions could be determined from the physical parameters of the system.

I also conceptually and mathematically demonstrated that some level of voltage, current and power are delivered rearward into a power source when its output impedance is not matched to the transmission line's characteristic impedance. The voltage and current delivered rearward into a source are simply components of the total steady-state voltage and current developed within its output stages.

This third article completes a general discussion of wave-reflection mechanics in a transmission-line system. It uses wave-reflection mechanics to explain impedance matching and relates that to concepts developed in the two previous articles. The discussion begins with how the total steady-state forward-traveling voltage, current and power are developed in a transmission line where a network is used to impedance match a transmission-line system. Finally, wave mechanics associated with achieving an impedance match using quarter-wavelength (λ/4) transformer and T-network matching circuits will be discussed and illustrated with examples.

[1]Notes appear on page 50.

48 Perimeter Rd
Manchester, NH 03103
srbest@att.net

The Total Re-Reflection Fallacy

A common misunderstanding persists regarding the wave re-reflections occurring at the output of a matching network or at a match point. It has to do with how the total steady-state forward-traveling power develops in the transmission-line section past the match point. The misconception is that a total re-reflection of the reflected power occurs at a matching network or match point as a result of an impedance match being established there. That notion leads to the erroneous conclusion that the total forward-traveling power developed in the transmission-line section past the match point is equal to the steady-state power delivered to the transmission line plus a forward power resulting from a total in-phase re-reflection of the reflected power arriving at the match point.

To discuss the concepts associated with the re-reflection of power at a matching network further, let's examine the steady-state conditions in the transmission-line system illustrated in Fig 1. In this system, an antenna having an impedance, Z_A, equal to 150 + j0 Ω is connected at the end of a loss-less transmission line, one wavelength long. For this discussion, assume that Z_0=50 Ω. In this case, the steady-state input impedance of the transmission line, Z_{IN}, is also equal to 150 + j0 Ω. The formula for determining the input impedance of any transmission line was presented in Part 1.

In Fig 1, the transmission-line input impedance, Z_{IN}, is impedance matched to the transmission line's characteristic impedance, Z_0, with a loss-less T-network circuit. The T-network circuit is connected to the transmitter with another section of loss-less transmission line also having a characteristic impedance of 50 Ω. Wattmeters (50-Ω) are connected at the T-network's input and output. The wattmeter connected at the input indicates a steady-state forward-traveling power of 100 W and a reflected or rearward-traveling power of 0 W. The wattmeter connected at the output indicates a forward-traveling power of 133.33 W and a reflected power of 33.33 W.

Looking at only these steady-state conditions, we see that no steady-state power travels rearward past the T-network input toward the transmitter, yet 33.33 W of reflected power arrive at the T-network output. At the same time, we see that 100 W of forward power arrive at the T-network input and that 133.33 W of power are traveling forward in the transmission-line section past the T-network. Given these conditions, it would seem to make sense that a total re-reflection of power occurs at the T-network and that the 33.33 W of reflected power are totally re-reflected, then add to the 100 W of forward power arriving at the T-network input, resulting in a total forward-traveling power of 133.33 W.

Before discussing this interpretation further, let's consider another example with a T-network circuit; except in this case, the T-network components are adjusted such that an impedance match does not exist at the T-network input. Let's adjust the T-network component values such that the steady-state input impedance at the T-network input is 100 Ω. At the same time, the transmitter's output level is adjusted such that the steady-state forward power arriving at the T-network input is 112.5 W. Under these conditions, the wattmeter at the T-network input indicates a forward power of 112.5 W and a reflected power of 12.5 W. The wattmeter connected at the T-network output indicates a forward power of 133.33 W and a reflected power of 33.33 W. The steady-state conditions beyond the T-network have not changed from the previous example. The question to ask now is: How should these steady-state conditions be interpreted?

One possible interpretation could be the following: The 112.5 W of forward power arrive at T-network input where 12.5 W of power are reflected and 100 W of power travel forward through the network to the transmission line connected to the antenna. The 33.33 W of reflected power arriving at the T-network are totally re-reflected and add to the 100 W of forward power, resulting in a total forward power of 133.33 W. The logic in this interpretation is consistent with the logic in the interpretation of total power re-reflection in the example with the impedance match. However, this interpretation does require us to accept that a total re-reflection of power always occurs at the T-network, regardless of whether or not an impedance match is established at its input.

If we do not accept this interpretation of these steady-state conditions and if we believe that a total re-reflection of power cannot occur when an impedance match does not exist at the network input, then another interpretation must be considered. Another interpretation could be the following: Let's consider the 33.33 W of reflected power arriving at the T-network. In the example with the impedance match, there was no reflected power traveling rearward past the T-network input, and we assumed that a total re-reflection of power occurred. In this example, 12.5 W of reflected power travels rearward past the T-network input. Let's assume that of the 33.33 W of reflected power arriving at the T-network, only 12.5 W travel rearward past the network input. This requires that a partial re-reflection of 20.83 W occurs at the T-network. At the same time, if we assume that the 112.5 W of forward power arriving at the T-network input travel through the network, then we could conclude that the 20.83 W of re-reflected power join with the 112.5 W of forward power, resulting in a total forward power of 133.33 W. This interpretation requires that we accept that the 112.5 W of forward power travel unimpeded past the network input.

In our examination of the steady-state power levels in both the impedance-matched and non-impedance-matched examples, we interpreted the relationship between the total forward power and the total reflected power based on some very broad assumptions. In each case, the math used to add the forward and re-reflected powers worked in arriving at the correct level of total forward power. However, the wave re-reflection concepts associated with each of these interpretations are considerably different. The real question we should be asking is: "Is any of these interpretations consistent with generally accepted transmission-line theory, basic circuit theory, and network theory—

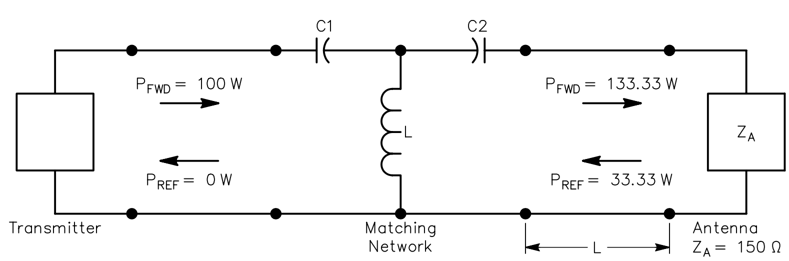


Fig 1—A basic transmission-line circuit with a matching network.

or with reality?" In answering this question, we will discover that each of the above interpretations is incorrect.

The first point to consider is that general transmission-line theory[3, 4, 5] indicates that the levels of voltage, current and power reflection or re-reflection at the end of a transmission line are direct functions of the impedance terminating the line. The reflection or re-reflection of a traveling wave in a transmission line occurs at the end of the line (assuming the line is not terminated in its characteristic impedance) or anywhere that an impedance discontinuity exists. For a total re-reflection of voltage, current or power to occur at a T-network, transmission-line theory requires that the physical or measurable impedance seen looking rearward into the matching network be a short circuit, open circuit or purely reactive. Since this generally would not be the case with a practical T-network impedance-matching circuit, the concept of total power re-reflection contradicts this fundamental aspect of transmission-line theory.

In fact, the net power delivered to the transmission-line or T-network input is the result of how the forward and rearward-traveling voltage and current develop at the input to the transmission line or T network through a process of multiple wave reflections and re-reflections. The most significant point is that the steady-state net power delivered to the transmission line or T network is simply the difference between the total steady-state forward and reflected powers (assuming Z_0 is a pure resistance). In any transmission-line system analysis, the steady-state net power delivered to the transmission line cannot be assumed or interpreted to be solely a forward-traveling power. Additional discussion on this aspect of wave-reflection mechanics will be presented in later sections of this article.

Assuming that the characteristic impedance of a transmission line is a pure resistance, the steady-state net power delivered to that transmission line, P_{DEL}, is equal to the total steady-state forward-traveling power, P_{FWD}, minus the total steady-state reflected power, P_{REF}, such that:

$$P_{DEL} = P_{FWD} - P_{REF} \qquad \text{(Eq 1)}$$

Using Eq 1, we can determine that the total steady-state power delivered to the transmission line in Fig 1 is equal to 100 W (133.33 W – 33.33 W). This relationship is true for both the impedance-matched and mismatched examples. In both cases, that is consistent with the fact that 100 W of total net power is delivered to the loss-less T-network input.

In Part 1, we used Eq 1 to develop the following relationship between delivered and forward power:

$$P_{FWD} = P_{DEL} \frac{1}{1-k^2|\rho_A|^2} \qquad \text{(Eq 2)}$$

Where ρ_A is the antenna reflection coefficient and k is determined from the total matched transmission-line attenuation, α, as follows:

$$k = 10^{-\frac{\alpha}{10}} \qquad \text{(Eq 3)}$$

The transmission-line attenuation, α, is expressed in decibels as a positive number. Since the transmission line in these examples is loss-less, $k = 1$. Given that in the above examples, $|\rho_A|$ is equal to 0.5 and that $P_{DEL} = 100$ W, we can use Eq 2 to determine that the total steady-state forward power in the transmission line is 133.33 W. It is critically important to recognize at this point that the relationships expressed in Eqs 1 and 2 are independent of whether or not an impedance match exists at the T-network input. Those relationships do not support the concept of a total re-reflection of power occurring when an impedance match is established at the T-network input.

A transmission-line system analysis must be performed with voltage and current, from which power is then derived. Let's look at the concept of power re-reflection and its relationship to the total forward power from the perspective of relating the forward and re-reflected voltages to the forward and re-reflected powers. We will consider the example where the T-network input impedance is matched to Z_0. If each wattmeter were replaced with a 50-Ω directional voltmeter and ammeter, the following traveling voltages and currents would be measured in the transmission line. At the T-network input, the forward-traveling voltage has a magnitude of 70.711 V and the forward-traveling current has a magnitude of 1.414 A. At the T-network output, the forward-traveling voltage has a magnitude of 81.650 V and the forward-traveling current has a magnitude of 1.633 A. The reflected voltage has a magnitude of 40.825 V and the reflected current has a magnitude of 0.816 A. The total voltage measured at the T-network output has a magnitude of 122.474 V—this is the vector sum of the forward and rearward traveling voltages.

Those levels of voltage magnitude, current magnitude and power within the transmission line are consistent with the relationships developed in the previous two article parts. The relationship between traveling voltage and current in any transmission-line system is given by:

$$\frac{\boldsymbol{V}_{FWD}}{\boldsymbol{I}_{FWD}} = \frac{\boldsymbol{V}_{REF}}{\boldsymbol{I}_{REF}} = Z_0 \qquad \text{(Eq 4)}$$

where $\boldsymbol{V}_{FWD}$ and $\boldsymbol{I}_{FWD}$ are the forward-traveling voltage and current, respectively, and $\boldsymbol{V}_{REF}$ and $\boldsymbol{I}_{REF}$ are the rearward-traveling voltage and current, respectively. Assuming that Z_0 is purely resistive, the traveling power in a transmission-line system is given by the following equations:

$$P_{FWD} = |\boldsymbol{V}_{FWD}||\boldsymbol{I}_{FWD}| = \frac{|\boldsymbol{V}_{FWD}|^2}{Z_0} = |\boldsymbol{I}_{FWD}|^2 Z_0 \qquad \text{(Eq 5)}$$

and

$$P_{REF} = |\boldsymbol{V}_{REF}||\boldsymbol{I}_{REF}| = \frac{|\boldsymbol{V}_{REF}|^2}{Z_0} = |\boldsymbol{I}_{REF}|^2 Z_0 \qquad \text{(Eq 6)}$$

where P_{FWD} and P_{REF} are the forward and reflected power traveling in the transmission line.

The concept of total power re-reflection at the matched T network was developed as part of an effort to relate the transmission line's steady-state conditions to the wave reflections and re-reflections occurring within the transmission line. That relationship was developed by first assuming that the forward source power applied to the transmission line at the T-network output is the steady-state net power delivered to the transmission line, which in the above Z_0-match example is 100 W. Next, the relationship between the total forward and re-reflected powers was developed by assuming that since the total steady-state forward power in the transmission line is 133.33 W, the reflected power of 33.33 W must be totally re-reflected and therefore combine with the 100 W of forward source power. These two forward-traveling powers are assumed to algebraically add in-phase, resulting in the 133.33 W of total forward-traveling power.

Let's consider the in-phase addition or combination of two forward-traveling waves to determine if it can be used to support the concept that a total re-reflection of the reflected power occurs at the T-network. This discussion will focus on the relationship between traveling voltage and power in the transmission line. Consider the Z_0 match example. From Eq 5, the total forward-traveling voltage at the T-network output has a magnitude of 81.650 V. The reflected voltage arriving at the T-network output has a magnitude of 40.825 V. Assuming that the forward source power applied to the transmission line is 100 W, the forward source voltage applied to the transmission line must have a magnitude of 70.711 V. If a total re-reflection of power occurs at the T-network, the re-reflected voltage must have the same magnitude as the reflected voltage. Therefore, based upon the assumption of total power re-reflection and in-phase forward-wave addition, the total forward-traveling voltage of 81.650 V must be the result of a voltage, having a magnitude of 70.711 V, adding in phase with a voltage having a magnitude of 40.825 V. Two in-phase complex voltages having magnitudes of 70.711 V and 40.825 V cannot add together such that the resulting total voltage has a magnitude 81.650 V. The addition of these two voltages, in a manner necessary to support the concept of a total re-reflection of power, is inconsistent with the relationships between voltage and power. The in-phase addition of two forward-traveling waves, one having a forward power of 100 W and the other having a forward power of 33.33 W, does not result in a forward-traveling wave having a total power of 133.33 W. That conclusion is consistent with general superposition theory as applied to transmission-line systems where *"... it is rare that the powers produced by two separate causes acting individually upon even a linear system can simply be added algebraically to determine the total power when both causes act together."*[5]

Vector Addition of Voltages and Currents

To understand this further, let's examine the concepts associated with the combination of two forward-traveling waves in a transmission-line system. A forward-traveling wave in a transmission line is comprised of a forward-traveling voltage and current. The forward-traveling voltage in the transmission line is a complex phasor that can be written in the general form of a vector $\mathbf{V} = x + jy = V\angle\theta$. In this case, V is the magnitude of the voltage phasor and θ, its phase angle. When two forward-traveling waves add, general superposition theory and Kirchhoff's voltage law require that the total forward-traveling voltage be the vector sum of the individual forward-traveling voltages such that $\boldsymbol{VF}_{\mathrm{TOTAL}} = \boldsymbol{V1} + \boldsymbol{V2}$ (see Notes 3, 4 and 5). The total forward-traveling voltage $\boldsymbol{VF}_{\mathrm{TOTAL}}$ is determined through the vector addition of $\boldsymbol{V1}$ and $\boldsymbol{V2}$ and is therefore a function of the individual voltage magnitudes and phases.

Understanding the relationship between voltage and power, we must accept that for a forward-traveling voltage $\boldsymbol{V1}$, the forward-traveling power $P1$ is given by:

$$P1 = \frac{|\boldsymbol{V1}|^2}{Z_0} \tag{Eq 7}$$

and that for a forward-traveling voltage, $\boldsymbol{V2}$, the forward-traveling power $P2$ is given by:

$$P2 = \frac{|\boldsymbol{V2}|^2}{Z_0} \tag{Eq 8}$$

where it is assumed that Z_0 is purely resistive.

Since the total forward-traveling voltage is given by $\boldsymbol{VF}_{\mathrm{TOTAL}} = \boldsymbol{V1} + \boldsymbol{V2}$, we can immediately see that the total forward-traveling power is given by:

$$PF_{\mathrm{TOTAL}} = \frac{|\boldsymbol{VF}_{\mathrm{TOTAL}}|^2}{Z_0} = \frac{|\boldsymbol{V1}+\boldsymbol{V2}|^2}{Z_0} \tag{Eq 9}$$

Since $\boldsymbol{V1}$ and $\boldsymbol{V2}$ are vectors, the expression $|\boldsymbol{V1} + \boldsymbol{V2}|^2$ is not necessarily equivalent to $(|\boldsymbol{V1}| + |\boldsymbol{V2}|)^2$. By examining the relationship $\boldsymbol{VF}_{\mathrm{TOTAL}} = \boldsymbol{V1} + \boldsymbol{V2}$, let's determine if it is possible that the total forward-traveling power can be found from $PF_{\mathrm{TOTAL}} = P1 + P2$ when $\boldsymbol{V1}$ and $\boldsymbol{V2}$ are in-phase.

First, let's assume that the complex voltage $\boldsymbol{V1}$ has a magnitude equal to $V1 = |\boldsymbol{V1}|$ and a relative phase angle of 0°. Let's also assume that the complex voltage $\boldsymbol{V2}$ has a magnitude equal to $V2 = |\boldsymbol{V2}|$ and a relative phase angle θ, having a value somewhere between 0° (in-phase) and 180° (out-of-phase).

If θ is between 0 and 90°, the two complex voltage vectors $\boldsymbol{V1}$ and $\boldsymbol{V2}$ would add graphically as depicted in Fig 2. In this case, we can determine $|\boldsymbol{VF}_{\mathrm{TOTAL}}|$ as follows:

$$\begin{aligned}|\boldsymbol{VF}_{\mathrm{TOTAL}}| &= \sqrt{(V1 + V2\cos\theta)^2 + (V2\sin\theta)^2} \\ &= \sqrt{V1^2 + V2^2\left(\cos^2\theta + \sin^2\theta\right) + 2V1V2\cos\theta} \\ &= \sqrt{V1^2 + V2^2 + 2V1V2\cos\theta}\end{aligned} \tag{Eq 10}$$

from which it follows that:

$$|\boldsymbol{VF}_{\mathrm{TOTAL}}|^2 = V1^2 + V2^2 + 2V1V2\cos\theta \tag{Eq 11}$$

It can be shown that this general relationship is valid for all angles of θ between 0° and 180°. From Eq 11, the general relationship for determining the total forward power resulting from the combination of two forward waves, one having an independent power equal to $P1$ and the other having an independent power equal to $P2$, is given as follows:

$$PF_{\mathrm{TOTAL}} = P1 + P2 + 2\sqrt{P1}\sqrt{P2}\cos\theta \tag{Eq 12}$$

From this theory of vector voltage addition, it is clear that the only time $PF_{\mathrm{TOTAL}} = P1 + P2$ is when $\theta = 90°$. When the voltages $\boldsymbol{V1}$ and $\boldsymbol{V2}$ are exactly in phase (θ = 0°), the total power can be determined as follows:

$$PF_{\mathrm{TOTAL}} = P1 + P2 + 2\sqrt{P1}\sqrt{P2} = \left(\sqrt{P1} + \sqrt{P2}\right)^2 \tag{Eq 13}$$

When the voltages $\boldsymbol{V1}$ and $\boldsymbol{V2}$ are exactly 90° out of phase (θ = 90°), the total power can be determined as follows:

$$PF_{\mathrm{TOTAL}} = P1 + P2 \tag{Eq 14}$$

When the voltages $\boldsymbol{V1}$ and $\boldsymbol{V2}$ are exactly in 180° out of phase (θ = 180°), the total power can be determined as follows:

$$PF_{\mathrm{TOTAL}} = P1 + P2 - 2\sqrt{P1}\sqrt{P2} \tag{Eq 15}$$

When two forward waves combine in-phase, the total power in the resulting wave does not equal the algebraic sum of the powers in the two individual waves. The in-phase addition of two forward-traveling waves, one having a power of 100 W and the other having a power of 33.33 W does not result in a wave with a total forward power equal to 133.33 W.

Based on this discussion, it is evident that the concept of total power re-reflection at the output of a matching network is incorrect. It does not describe the correct mechanism of how the total forward-traveling power in a transmission line develops through the wave re-reflections occurring there. In the system illustrated in Fig 1, there

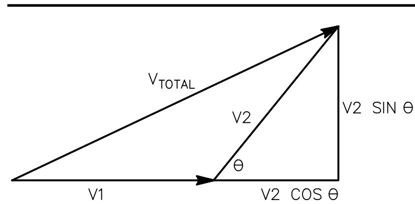


Fig 2—Vector addition of two complex voltage phasors, *V1* and *V2*.

will be some level of steady-state voltage, current and power re-reflected at the T-network output. The re-reflected voltage and current add to the forward source voltage and current applied to the transmission line, resulting in the total steady-state forward-traveling voltage and current. The important issue is to determine the actual forward source voltage and current applied to the transmission line and the actual level of voltage and current re-reflected at the T-network output. The correct method for determining these variables will be illustrated below. Examples of how to determine the steady-state forward-traveling power from the forward source voltage and current will be presented using a λ/4 transformer and a T-network matching network. At the same time, we will examine the mechanics of how steady-state impedance matching is achieved.

Impedance Matching

How does a steady-state impedance match result from the multiple wave reflections occurring between the matching device and the load? The answer is: "With any matching device or circuit, a steady-state impedance match occurs through the cancellation of multiple rearward-traveling waves at the match point."

When an antenna has an impedance other than the characteristic impedance of its transmission line, a rearward-traveling reflected wave is created at the antenna. With an impedance matching device, the objective is to prevent this reflected wave from traveling rearward to the transmitter. To understand the mechanics of how a matching device works, first consider that a transmitter applies a forward-traveling source wave that arrives at the input to the matching device. At the matching device input or match point, some portion of the forward-traveling source wave is reflected (rearward wave 1) and some portion of the forward-traveling source wave is transmitted forward past the match point toward the antenna. The portion of the forward-traveling source wave that arrives at the antenna is partially reflected rearward toward the matching device. At the matching device, some portion of this reflected wave is re-reflected back to the antenna and some portion of this reflected wave is transmitted rearward through the matching device or match point (rearward wave 2). When the system reaches the steady state, the two rearward-traveling waves at the match point (rearward waves 1 and 2) are 180° out of phase with respect to each other and a complete cancellation of both waves occurs. The result of this wave cancellation is that the total steady-state rearward-traveling wave has a net voltage and current of 0 V and 0 A, respectively, and an impedance match occurs.

Although there is some level of wave re-reflection at the matching device, the steady-state impedance match does not occur through a total re-reflection of the reflected wave. If a total re-reflection of the reflected wave from the antenna did occur, the impedance match would never develop because there would be no mechanism to cancel the reflected wave created at the matching device's input (rearward wave 1).

Impedance matching in a transmission-line system may be implemented using a number of techniques. In this article, we will consider the wave mechanics of the impedance matching process when it is achieved using either a λ/4 transformer or a T-network circuit. The concepts and formulas presented in the previous two segments are used to describe how an impedance match occurs.

Significant points illustrated in this discussion are:

1. With all impedance matching techniques, the re-reflection mechanism at the matching device is the physical impedance seen by the rearward-traveling reflected wave when it arrives at the matching device;
2. The impedance match occurs as a result of the cancellation of rearward-traveling waves;
3. A total re-reflection of the reflected voltage, current and power does not occur at the match point.

The Quarter-Wavelength Transformer

One common technique that establishes an impedance match in a transmission-line system is the use of a λ/4 transformer, as depicted in Fig 3. The basic operation of a λ/4 transformer is discussed in terms of wave mechanics. The process determines the steady-state solution for all of the transmission-line system's forward- and rearward-traveling voltages as a function of the forward source voltage. That steady-state solution is valid for a transmission-line transformer of any length or characteristic impedance.

The operation of a λ/4 transformer is actually very straightforward. The forward-traveling source wave arriving at the input to the transmission line transformer sees the transformer's characteristic impedance as an impedance discontinuity. As a result, some portion of the forward source wave is reflected rearward at the transformer input and some portion of the forward source wave is transmitted forward into the transformer. The portion of the forward source wave transmitted into the transformer travels to the antenna, where a portion of it is, in turn, reflected back toward the transformer input and a portion is delivered to the antenna. The portion of the source wave reflected at the antenna arrives back at the transformer input, where it sees the characteristic impedance of the main transmission line as an impedance discontinuity. Consequently, some portion of the reflected wave is re-reflected back toward the antenna and a portion of the reflected wave is transmitted rearward into the main transmission line. The portion of the reflected wave re-reflected back toward the antenna combines with the forward source wave and the wave reflection/re-reflection process continues, until equilibrium is reached. In the steady state, the total wave transmitted rearward into the main transmission line as a result of the reflections in the transformer is exactly 180° out of phase with the original source wave reflected at the transformer input. These two rearward-

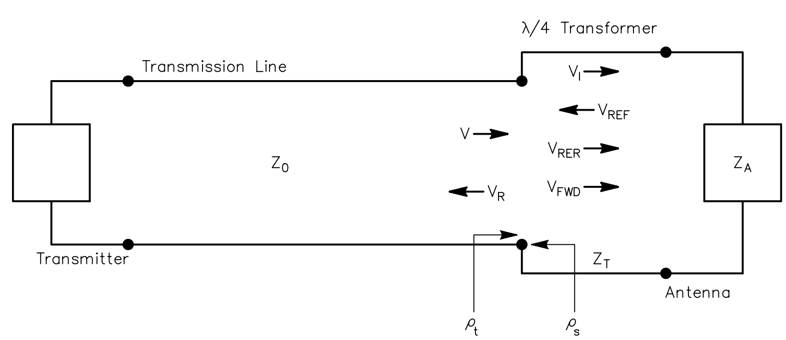


Fig 3—A transmission-line system with a λ/4 transformer.

traveling waves cancel one another; therefore, the net rearward-traveling wave in the main transmission line is zero and the steady-state match is achieved.

Let's consider an example wherein we want to match an antenna impedance $Z_A = 150 + j0\ \Omega$ to a transmission line having a characteristic impedance $Z_0 = 50\ \Omega$. To accomplish the impedance match, we use a loss-less λ/4 transformer as illustrated in Fig 3. To determine the appropriate characteristic impedance of the λ/4 transformer, Z_T, we use the following equation:

$$Z_T = \sqrt{Z_0 Z_A} \quad \text{(Eq 16)}$$

In this example, Z_T=86.60 Ω. Using the formulas previously developed, the steady-state input impedance to the 86.6-Ω, λ/4 transformer is calculated to be 50 Ω. As a result, a steady-state impedance match is achieved.

We must first correctly determine the forward source voltage applied to the transformer and then perform a detailed and consistent analysis. For ease of calculation in the analysis, we will assume that both transmission lines are loss-less and that the 50-Ω main transmission line is one wavelength long. Let's also assume that the transmitter applies a forward source voltage of 70.711 V to that transmission line.

Since the main transmission line is one wavelength long and loss-less, the 70.711 V applied by the transmitter arrives at the input to the λ/4 transformer without attenuation or relative phase delay. This forward-traveling source voltage, $\boldsymbol{V}$, arrives at the input to the λ/4 transformer and sees the transformer's characteristic impedance, Z_T, as an impedance discontinuity. The equivalent reflection coefficient at the input to the λ/4 transformer, ρ_T, is given by:

$$\rho_T = \frac{\frac{Z_T}{Z_0} - 1}{\frac{Z_T}{Z_0} + 1} \quad \text{(Eq 17)}$$

Under these conditions, ρ_T is found from Eq 17 to be 0.268. Since ρ_T = 0.268, a rearward-traveling reflected voltage is created at the junction between the 50-Ω main transmission line and the λ/4 transformer. This reflected voltage, $\boldsymbol{V}_R$, is given by:

$$\boldsymbol{V}_R = \rho_T \boldsymbol{V} \quad \text{(Eq 18)}$$

The reflected voltage developed at the input to the transformer, V_R, is therefore equal to 18.947 V (70.711 × 0.268). The reflected power developed at the match point is 7.180 W. The total forward voltage transmitted into the λ/4 transformer as the forward-traveling source voltage, $\boldsymbol{V}_I$ is determined from the sum of the incident and reflected voltages at the transformer input as follows:

$$\boldsymbol{V}_I = \boldsymbol{V} + \boldsymbol{V}_R = \boldsymbol{V}(1 + \rho_T) \quad \text{(Eq 19)}$$

The forward-traveling source voltage $\boldsymbol{V}_I$ is therefore equal to 89.658 V (70.711 + 18.947). The forward source power applied or transmitted to the transformer is determined from Eq 5 to be 92.821 W. Note that the forward source power transmitted in the transformer is not equal to the steady-state net power delivered to the transformer.

Since the λ/4 transformer is loss-less, the forward-traveling voltage arriving at the antenna will not be attenuated. Since the transformer also has an electrical length of 90°, the voltage arriving at the antenna will undergo a relative phase delay of 90° with respect to $\boldsymbol{V}_I$ at the input of the transformer. A reflected voltage is created at the antenna as a direct function of the antenna's reflection coefficient, ρ_A. The value of ρ_A can be determined using Eq 17, where Z_A is substituted for Z_T, and Z_T is substituted for Z_0. Under these conditions, ρ_A is also equal to 0.268. The reflected voltage created at the antenna arrives back at the input to the transformer as the defined voltage $\boldsymbol{V}_{REF}$. The voltage $\boldsymbol{V}_{REF}$ sees the characteristic impedance of the feeder transmission line, Z_0, as an impedance discontinuity. A re-reflected voltage, $\boldsymbol{V}_{RER}$, is created at the input to the transformer as a direct function of the reflection coefficient of the feeder transmission line, ρ_S. The value of ρ_S can be determined using the following equation:

$$\rho_S = \frac{\frac{Z_0}{Z_T} - 1}{\frac{Z_0}{Z_T} + 1} \quad \text{(Eq 20)}$$

Under these conditions, ρ_S= –0.268. The re-reflected voltage, $\boldsymbol{V}_{RER}$ combines with the forward-traveling voltage $\boldsymbol{V}_I$ (vector addition) and the process of wave reflection and re-reflection continues until equilibrium is reached.

Using the following equations developed previously, the steady-state voltages at the input to the λ/4 transformer can be determined. In these equations, L is the length of the λ/4 transformer and γ is the propagation factor of the transformer.

$$\boldsymbol{V}_{REF} = \boldsymbol{V}_I \frac{\rho_A e^{-2\gamma L}}{1 - \rho_S \rho_A e^{-2\gamma L}} = \boldsymbol{V}(1 + \rho_T) \frac{\rho_A e^{-2\gamma L}}{1 - \rho_S \rho_A e^{-2\gamma L}} \quad \text{(Eq 21)}$$

$$\begin{aligned} \boldsymbol{V}_{RER} &= \rho_S \boldsymbol{V}_{REF} = \boldsymbol{V}_I \frac{\rho_S \rho_A e^{-2\gamma L}}{1 - \rho_S \rho_A e^{-2\gamma L}} \\ &= \boldsymbol{V}(1 + \rho_T) \frac{\rho_S \rho_A e^{-2\gamma L}}{1 - \rho_S \rho_A e^{-2\gamma L}} \end{aligned} \quad \text{(Eq 22)}$$

$$\begin{aligned} \boldsymbol{V}_{FWD} &= \boldsymbol{V}_I + \boldsymbol{V}_{RER} = \boldsymbol{V}_I \frac{1}{1 - \rho_S \rho_A e^{-2\gamma L}} \\ &= \boldsymbol{V}(1 + \rho_T) \frac{1}{1 - \rho_S \rho_A e^{-2\gamma L}} \end{aligned} \quad \text{(Eq 23)}$$

In this example, the steady-state reflected voltage arriving at the transformer input, $\boldsymbol{V}_{REF}$, is determined from Eq 21 to be –25.882 V. From Eq 6, the steady-state reflected power arriving at the input to the transformer is 7.735 W. From Eq 22, the steady-state re-reflected voltage developed at the input to the transformer, $\boldsymbol{V}_{RER}$, is 6.935 V. The steady-state re-reflected power developed at the transformer input is 0.555 W. The total steady-state, forward-traveling voltage developed at the transformer input, $\boldsymbol{V}_{FWD}$, is 96.593 V from Eq 23. The total steady-state forward-traveling power developed at the transformer input is 107.736 W from Eq 5.

This level of total forward-traveling power can also be determined from the combination of two in-phase powers (waves) using Eq 13. In this case, the forward source power of 92.831 W combines in-phase with the re-reflected power of 0.555 W, resulting in a total forward-traveling power of 107.736 W. Again, it is significant to notice that this "addition" of power is determined entirely from the vector addition of forward voltage and current. It must be understood that it is really the two forward voltages and currents that combine, rather than the two forward powers.

The level of steady-state rearward-traveling voltage transmitted back into the 50-Ω feeder transmission line is defined as $\boldsymbol{V}_{BACK}$, and is determined from the sum of $\boldsymbol{V}_{REF}$ and $\boldsymbol{V}_{RER}$, as follows:

$$\boldsymbol{V}_{BACK} = \boldsymbol{V}_{REF} + \boldsymbol{V}_{RER} \quad \text{(Eq 24)}$$

$$\boldsymbol{V}_{\mathrm{BACK}} = \boldsymbol{V}(1+\rho_{\mathrm{T}})\frac{\rho_{\mathrm{A}} e^{-2\gamma L}}{1-\rho_{\mathrm{S}}\rho_{\mathrm{A}} e^{-2\gamma L}} + \boldsymbol{V}(1+\rho_{\mathrm{T}})\frac{\rho_{\mathrm{S}}\rho_{\mathrm{A}} e^{-2\gamma L}}{1-\rho_{\mathrm{S}}\rho_{\mathrm{A}} e^{-2\gamma L}} \qquad \text{(Eq 25)}$$

$$\boldsymbol{V}_{\mathrm{BACK}} = \boldsymbol{V}(1+\rho_{\mathrm{T}})(1+\rho_{\mathrm{S}})\frac{\rho_{\mathrm{A}} e^{-2\gamma L}}{1-\rho_{\mathrm{S}}\rho_{\mathrm{A}} e^{-2\gamma L}} \qquad \text{(Eq 26)}$$

The rearward-traveling voltage, $\boldsymbol{V}_{\mathrm{BACK}}$, is –18.947 V. This rearward-traveling voltage is the exact negative of the +18.947 V reflected at the input to the transformer ($\boldsymbol{V}_{\mathrm{R}}$). Therefore, the total steady-state rearward-traveling voltage in the 50-Ω feeder transmission line is 0 V. Since the total steady-state rearward-traveling voltage is 0 V, there is no net rearward-traveling wave; hence, a steady-state impedance match occurs. This is the rearward voltage-cancellation mechanism that causes the effective steady-state input impedance to be 50 Ω at the input to the λ/4 transformer.

Eq 26 for $\boldsymbol{V}_{\mathrm{BACK}}$ is valid for a transformer design of any Z_{T} and length L. For an impedance match to occur, the value of $\boldsymbol{V}_{\mathrm{BACK}}$ must simply be equal to the negative of V_{R} as defined by Eq 18. Note that for the loss-less λ/4 transformer, substituting $e^{-2\gamma L} = -1$, $\rho_{\mathrm{A}} = \rho_{\mathrm{T}}$, and $\rho_{\mathrm{S}} = -\rho_{\mathrm{T}}$, Eq 18 for $\boldsymbol{V}_{\mathrm{BACK}}$ simplifies to $\boldsymbol{V}_{\mathrm{BACK}} = -V\rho_{\mathrm{T}} = -V_{\mathrm{R}}$. Note also that if the quarter-wave transformer were lossy, a perfect impedance match would not occur without an adjustment to the transformer's characteristic impedance.

Another important aspect of wave interactions in impedance matching is how the law of conservation of energy is satisfied at the match point. That law requires that the total power arriving at the match be equal to the total power leaving it, assuming that no power dissipation occurs at the match point. In the case of the λ/4 transformer, two steady-state waves arrive at the match point. One is a steady-state forward source wave with a power of 100 W; the other is a steady-state reflected wave with a power of 7.735 W. Since a forward source power of 100 W arrives at the match point, we determine that 7.180 W are reflected at the match point and that 92.821 W are transmitted forward into the transformer. Since 7.735 W of reflected power arrives at the match point, we determine that 0.555 W of power is re-reflected into the transformer and that 7.180 W are delivered rearward into the main transmission line. The two forward waves having 92.821 W and 0.555 W of forward power combine in-phase to become a forward wave with a total power of 107.736 W. The two rearward waves, each having a power of 7.735 W, combine 180° out of phase, resulting in a rearward wave with a total or net power of 0 W. Thus, the law of conservation of energy is satisfied.

The two basic facts to understand in an impedance-matching process are:

1. The thing causing re-reflection at the input to the λ/4 transformer is the impedance seen by the rearward-traveling reflected wave as it arrives at the transformer input,
2. A total re-reflection of voltage, current and power does not occur at the match point and is not necessary for the impedance match to occur.

The T-Network Tuner

An impedance-matching process similar to that described above occurs when a T-network or other impedance matching circuit is used to match an antenna impedance to a main transmission line's characteristic impedance. Let's now consider an example using a T-network circuit to match an antenna impedance Z_{A}=150 Ω to Z_0=50 Ω. The T-network circuit configuration is illustrated in Fig 4. The input terminal is designated Input A and the output terminal is designated Output B. To illustrate the impedance-matching process that occurs with the T-network circuit, let's again assume that the 50-Ω main transmission line is loss-less. We will assume that the T-network is arbitrarily placed 1.20 λ away from the antenna such that the steady-state input impedance at that point is equal to 18.213 –j14.273 Ω. This is the effective steady-state load or line impedance seen by the T-network at point B. The frequency of operation assumed for this example is 21.2 MHz.

In this example, the components in the T-network are assumed to be loss-less. Suitable matching-component values necessary to achieve the steady-state impedance match are determined to be: $C1$ = 27.8568 pF (Z = –j269.496 Ω), L = 0.782384 µH (Z = j104.216 Ω) and $C2$ = 50.0 pF (Z = –j150.146 Ω). Using simple circuit theory, the steady-state input impedance at point A is calculated to be the desired 50-Ω impedance match. To understand how this steady-state impedance match occurs with the T-network circuit, we must analyze the multiple wave reflections occurring between the T-network output and the antenna. A circuit diagram for this example is presented in Fig 5 to support this analysis.

The forward source voltage arriving at the T-network input, $\boldsymbol{V}$, is equal to 70.711 V, as was the case with the λ/4 transformer example. From the perspective of the forward source wave arriving at the T-network input, the only physical load impedance at the T-network output is the characteristic impedance of the transmission line, Z_0. Therefore, the input impedance seen by the forward source wave at point A in the T-network circuit is not equal to the steady-state matched impedance of 50 Ω. With a Z_0 load connected at point B in the T-network, the input impedance at point A is found through circuit theory to be Z = 117.810 –j57.060 Ω. This impedance has a complex reflection coefficient, ρ_{T}, equal to 0.466 – j0.182. The magnitude of this reflection coefficient, $|\rho_{\mathrm{T}}|$, is equal to 0.5.

The forward-traveling source voltage of 70.711 V sees the impedance Z when it arrives at point A and a rearward-traveling reflected voltage is created at the T-network input. This reflected voltage, $\boldsymbol{V}_{\mathrm{R}}$, is 32.940 –$j$12.843 V. The reflected power in this case is 25 W. The total voltage developed at the input to the T-network is equal to the sum of incident and reflected voltages and is 103.651 –j12.843 V. This voltage is the total forward source voltage transmitted into the T-network creating the forward source voltage, $\boldsymbol{V}_{\mathrm{I}}$, in the transmission line connecting the T-network and the antenna. Using simple circuit theory, the voltage $\boldsymbol{V}_{\mathrm{I}}$ is found to be –53.740 + j29.361 V. The forward source power applied to the transmission line is, from Eq 5, 75 W. Notice again that the forward source power transmitted through the network to the transmission line is not the steady-state delivered power of 100 W. The fact that the forward source power is 75 W, rather than 100 W, is significant in correctly interpreting the steady-state conditions and the relationship between the total steady-state forward and re-reflected powers.

As with the λ/4-transformer example, we can use Eqs 21,

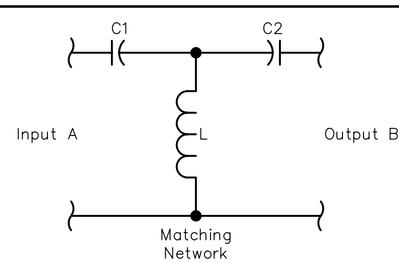


Fig 4—A typical T-network tuner configuration.

22 and 23 to determine the system's steady-state conditions from the forward source voltage $\boldsymbol{V}_{\mathrm{I}}$. The issue now is to correctly determine the values of ρ_{S} and ρ_{A}. The value of ρ_{A} is easily determined from the antenna impedance to be 0.50. The value of ρ_{S} is determined from the physical impedance seen by the rearward-traveling reflected voltage when it arrives at the output of the T-network. This impedance is determined using circuit theory by looking rearward into the T-network from point B where the input to the T-network at point A is terminated by Z_0. Looking rearward into the T-network from point B, point A is effectively terminated in Z_0 because it is assumed that the impedance seen looking rearward from point A toward the transmitter is Z_0. The choice of Z_0 in this calculation is independent of the steady-state input impedance at point A, also Z_0. The impedance seen looking rearward into the T-network from point B is Z_{S} = 18.213 + j14.273 Ω, which has a complex reflection coefficient, ρ_{S}, equal to –0.405 + j0.294. The magnitude of this reflection coefficient, $|\rho_{\mathrm{S}}|$, is 0.5. Notice that all of the reflection coefficients in the system have the same magnitude because there are no losses within the system.

Using Eq 23, the total steady-state, forward-traveling voltage developed at the input to the transmission line connecting the T-network and the antenna is $\boldsymbol{V}_{\mathrm{FWD}}$ = –71.653 + j39.148 V. Using Eq 5, the total steady-state forward-traveling power developed at the transmission-line input is 133.33 W. Using Eq 21, the steady-state rearward-traveling reflected voltage developed at the T-network output is $\boldsymbol{V}_{\mathrm{REF}}$ = 40.489 + j5.223 V. Using Eq 6, the total steady-state reflected power arriving the T-network output is 33.33 W. Additionally, using Eq 22, the steady-state re-reflected voltage developed at the T-network output is $\boldsymbol{V}_{\mathrm{RER}}$ = –17.913 + j9.787 V. The re-reflected power developed at the T-network output is calculated from Eq 5 to be 8.33 W.

In this case, the re-reflected voltage developed at the tuner output ($\boldsymbol{V}_{\mathrm{RER}}$) is in phase with the forward source voltage applied to the transmission line ($\boldsymbol{V}_{\mathrm{I}}$). Using Eq 13, the correct total steady-state forward-traveling power of 133.33 W can be determined through the in-phase combination of the forward source wave applied to the transmission line (75 W power) and the re-reflected wave developed at the matching-network output (8.33 W power). The total forward source power is not a result of a total re-reflection of 33.33 W adding to a delivered power of 100 W.

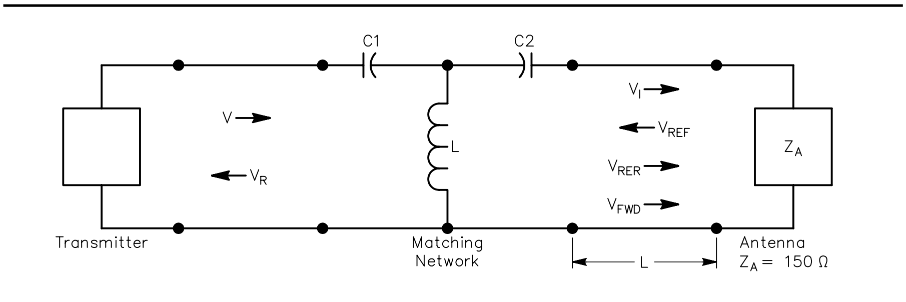


Fig 5—A typical transmission-line system with a T-network circuit.

Because of the reflected voltage arriving at the output of the T-network at point B, there is some level of rearward-traveling voltage, $\boldsymbol{V}_{\mathrm{BACK}}$, delivered to the main transmission line. The rearward-traveling voltage delivered back into the main line is determined using simple circuit theory. The total rearward voltage developed at point B in the T-network is the sum of $\boldsymbol{V}_{\mathrm{REF}}$ and $\boldsymbol{V}_{\mathrm{RER}}$. This rearward-traveling voltage is 22.576 + j15.010 V. The level of rearward power delivered back into the T-network is 25 W. Using circuit theory, the rearward-traveling voltage delivered back into the main transmission line, $\boldsymbol{V}_{\mathrm{BACK}}$, is –32.940 + j12.843 V. This rearward-traveling voltage is the exact negative of the reflected voltage, $\boldsymbol{V}_{\mathrm{R}}$, created at the input to the T-network. Therefore, the total steady-state, rearward-traveling voltage in the main transmission line is 0 V. This cancellation of all rearward-traveling voltages within the main transmission line is the mechanism that causes the effective steady-state input impedance to be 50 Ω at the input to the T-network.

As with the λ/4-transformer example, the two important facets are:

1. The mechanism for re-reflection at the T-network output is the termination seen by the rearward-traveling reflected wave as it arrives at the T-network output, and
2. A total re-reflection of voltage, current and power does not occur.

Summary

This series of articles has presented a detailed discussion of how the forward-traveling power is developed at the input to a transmission line through reflections and re-reflections of power traveling in a transmission line. Basic concepts of impedance matching were discussed and the mechanism for reflection cancellation with an impedance-matching device was illustrated.

Major points of conclusion include the following:

- A total re-reflection of power does not occur at the output of a matching device or at an impedance-match point. It was demonstrated that the concept of total power re-reflection is inconsistent with both transmission-line and circuit theories.
- The level of total forward-traveling voltage in a transmission line is a direct result of the vector summation of the forward source voltage applied to the transmission line and the partial re-reflected voltage created at the matching network output terminals or match point.
- Two in-phase forward-traveling powers in a transmission line do not algebraically add to give the total forward power.
- A steady-state impedance match in any transmission-line system occurs via the direct cancellation of all rearward-traveling voltages and currents at the match point. A total re-reflection of power at the match point is not necessary for the impedance match to occur.

Acknowledgment

I would like to thank Mr Jeff Anderson, WA6AHL, for his valuable comments and suggestions regarding the content and format of this article. I would also like to thank Mr. William Klocko, N3WK, for his valuable discussions regarding the subject of wave reflections in transmission-line systems.

Notes

1. S. Best, "Wave Mechanics of Transmission Lines, Part 1: Equivalence of Wave Reflection Analysis and the Transmission Line Equation," *QEX*, Jan/Feb 2001, pp 3-8.
2. S. Best, "Wave Mechanics of Transmission Lines, Part 2: Where Does Reflected Power Go?," *QEX*, Jul/Aug 2001, pp 34-42.
3. R. Chipman, *Schaum's Outline of Theory and Problems of Transmission Lines*, (New York: McGraw-Hill, 1968).
4. W. Johnson, *Transmission Lines and Networks*, (New York: McGraw-Hill, 1950).
5. R. Adler, Lan Jen Chu, and Robert M. Fano, *Electromagnetic Energy Transmission and Radiation*, (New York: John Wiley and Sons, 1960).

More Properties of Circular Waveguide

Some things you ought to think about before you buy all that copper pipe!

By William B. Bridges, W6FA

The recent paper by Paul Wade, W1GHZ[1] was an excellent reminder that readily obtainable copper pipe is an economical alternative to rectangular or elliptical wave-guide for amateurs. Wade's experi-mental approach to the design of components for such waveguide should encourage many other microwave amateurs to go and do likewise. There are a few properties of circular waveguide that should be understood at the outset, though, lest the eager experimenter find that the waves at the output are not what he or she expects.

[1]Notes appear on page 54.

Mail Code 136-93
California Institute of Technology
Pasadena, CA 91125
w6fa@caltech.edu

Propagation Modes and Cutoff Frequencies

One might hear a nagging voice of suspicion raised by Wade's observation in his first paragraph: "Unfortunately, commercial circular waveguide is rare and unlikely to be found surplus." Why is this so, we might wonder, when it is obvious that round pipe is much cheaper to make than rectangular or elliptical cross section cylinders?

The answer is that circular waveguide is not used commercially because of the unavoidable multimode nature of circular waveguide. Wade correctly observes that the circular TE_{11} mode has the lowest cut-off frequency f_c—the frequency below which a mode ceases to propagate—of 8.55 GHz for 3/4-inch copper pipe (0.81 inch ID). The theoretical field distribution for a cross-section of that mode is shown in Fig 1. The next-order mode, TM_{01}, has a cut-off frequency of 11.08 GHz for the same pipe and an entirely different field distribution. Thus, if we use the waveguide above 8.55 GHz but below 11.08 GHz, we have only the TE_{11} mode. So what's the problem?

The problem is that an *infinite number* of TE_{11} modes are possible, all with the same cutoff frequency. They differ only in the angular orientation of the fields within the pipe! Simply imagine the distribution of Fig 1 rotated to different orientations in the circular pipe. There is no "anchor" for the mode in a circular cross-section—except at its source of excitation.

The situation is very different for rectangular waveguide, where the lowest-order mode is termed TE_{10}, as shown in Fig 2; the cutoff frequency is set by the largest dimension of the rectangle ($\lambda_c = 2a$, where λ_c is the cutoff wavelength). There is one and only one field distribution possible for this rectangle if the frequency is low enough so that only the TE_{10} mode propagates. A second mode—TE_{01}, which looks like the TE_{10}, but rotated 90° in space—can propagate at a higher frequency, determined by the shorter dimension of the rectangle ($\lambda_c = 2b$). If the rectangle is stretched into a square, both of these modes are "cut on" (as opposed to "cut off"—*Ed.*) over the same frequency range, and the waveguide is no longer "single-mode" at any frequency. This is the reason that rectangular waveguide with about a 2:1 aspect ratio is used commercially: It is truly single-mode over almost a 2:1 frequency range with a well-defined polarization, or direction of the electric field. The same is true for elliptical waveguide, which you can visualize by squashing Fig 1 vertically into a 2:1 ellipse. Now there are rotational features to anchor the mode.

Wade refers to this polarization ambiguity problem for the circular TE_{11} mode. He cautions that the probes used to excite and detect the wave must be parallel to the electric field of the TE_{11} mode—for example, in the discussion of his slotted line. It is true that the probe-coupled coax-to-waveguide transition Wade shows in his Fig 3 will excite a single TE_{11} mode, and that mode should maintain its angular orientation, or polarization, as it propagates down a straight piece of perfect pipe.

The problem arises when the pipe is not perfect (imagine a dent) or is bent (intentionally or unintentionally). If the dent or bend is in the E or H plane,[2] nothing too serious happens. Some of the power in the TE_{11} mode is scattered into the cut-off modes to match boundary conditions neces-

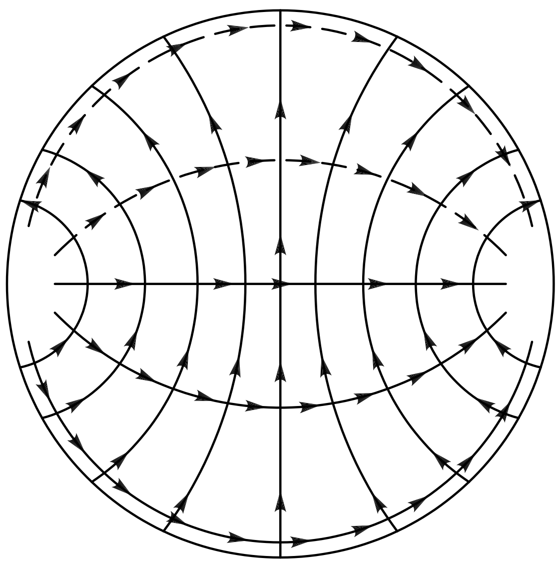


Fig 1—A sketch of the transverse electric (solid lines) and magnetic (dashed lines) fields for a cross section of the circular TE_{11} mode where the transverse fields are strongest. The electric field lines end in charges on the wall. The magnetic field lines close on themselves by running longitudinally down the side walls at 3 o'clock and 9 o'clock to the next transverse field maximum, a half-guide-wavelength away. Both electric and magnetic fields at the walls are 63% of their strength at the horizontal diameter. The arrowhead directions indicate that this mode is traveling into the paper.

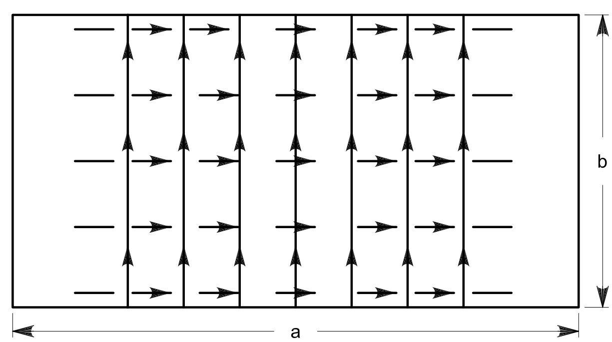


Fig 2—A sketch of the fields for the rectangular TE_{10} mode. All the comments for Fig 1 apply except that the fields are uniform in strength from top to bottom.

Fig 3—An end view of the beer-can polaplexer, showing the 726 klystron at 4 o'clock, the local oscillator injection screw at 6 o'clock, the crystal mount and BNC IF output at 7 o'clock and the sliding coax shorted stub at 1 o'clock.

Fig 4—A side view of the beer-can polaplexer. Notice that the receiving antenna (crystal and sliding short) are about $\lambda_c/4$ from the closed end of the can, and the klystron and transmitting antenna are about $3\lambda_c/4$ from the closed end. Judging by the label, I must have removed the bottom of the beer can rather than the top! (This can predates pull-tabs.)

sary to describe the dent or bend. These modes do not propagate and represent only locally stored energy (an equivalent reactive element added to the guide). The net result is small mismatch for the TE_{11} mode, with some reduction in transmission and some reflection. If the dent or bend is at an angle with respect to these two planes though, energy is scattered into TE_{11} modes with other polarizations and they propagate. What comes out the end of the pipe may be quite different from what you put in!

Some simple cases illustrate the point. Consider a small screw (say a #4-40) tapped through the pipe wall 45° from the E plane. This will scatter some power from the excited TE_{11} mode into another TE_{11} mode rotated 90° from the E plane. The amount of power diverted into this mode will depend on how deeply this screw penetrates into the pipe. What comes out the end of the pipe will be the superposition of these two TE_{11} modes, which will also look like a TE_{11} mode, but rotated by some angle that depends on the depth of penetration of the screw. Thus, we have a variable polarization rotator!

The Beer-Can Polaplexer

Such a component has practical applications. Figs 3 and 4 are photographs of a "beer-can *polaplexer*"[3] I made when I was a graduate student in the late 1950s. A 16-oz beer can with its top removed is used to make two coax-to-waveguide transitions at right angles to one another. The brand of beer is unimportant, but a tin-plated can is essential so you can solder to it. The wire spanning the full diameter of the can is located about $\lambda_g/4$ from the bottom of the can. It is terminated at one end with a variable-length shorted coax stub (for tuning) and at the other with a 1N21 diode used as a mixer. The BNC connector is the IF output.

The wire probe with a free end originates from the coaxial pin of a type-726A reflex klystron plugged into an octal socket on the outside of the can. This probe is located $3\lambda_g/4$ from the bottom of the can and oriented at 90° to the other wire. There is also a #6-32 screw 45° from either probe that extends into the can through a nut soldered to the outside of the can.

The operation is simple: The 726 klystron and its probe serve as the transmitter radiating a TE_{11} mode out the open end of the can. The other probe is the receiving antenna for an orthogonally polarized TE_{11} mode, and the #6-32 screw couples a little of the transmitted power into the receiving antenna to serve as a local oscillator for the crystal detector. The screw is adjusted for optimum crystal current—about 1 mA, as I recall. The set-up on the other end of the microwave link is identical, but with its beer can rotated 90°. Its klystron is offset in frequency by the IF used following the mixer—typically 100 MHz, so a commercial FM radio can serve as the IF. Here is an early example of amateur circular waveguide at 3.5 GHz, with two orthogonal TE_{11} modes.

An interesting variation on the beer-can polaplexer can be used to create circularly polarized radiation, something that may be useful in an antenna feed. Imagine feeding the two probes in the polaplexer device directly with microwave signals of the same frequency and equal amplitudes, but with a variable phase delay. Adjust that phase delay so that the two orthogonal TE_{11} modes are 90° out of time phase, as well as being 90° out of space phase. The result is a circularly polarized TE_{11} mode. A signal in such a mode will radiate at the open end as a circularly polarized spherical wave, or it may be sent down a pipe as a guided mode.

Bends in Circular Waveguide

The problem of bends is more complex. Rectangular TE_{10} waveguide can be bent in a very tight radius (a few wavelengths) without problems because of the "tight hold" the rectangular walls have on the mode. This is not the case for circular TE_{11} guide. The solution for a TE_{11}-like mode in a bent pipe is not easy to express mathematically, but an important conclusion can be stated simply. The guide wavelength is slightly changed by the bend, and the amount of change depends on the relationship between the plane of polarization and the plane of the bend. We can best think of this in terms of two possible TE_{11} modes: One has its electric field exactly in the plane of the bend (an "E-plane bend"), and one has its electric field oriented perpendicular to the plane of the bend (an "H-plane bend"). These two modes will have slightly different guide wavelengths in the bend region, and hence will suffer slightly different phase shifts going through the same bend.

If our actual mode is either one of these, then the only effect of the bend is to change the total phase shift through the bend slightly—no big deal. However, if our desired TE_{11} mode is polarized at some arbitrary angle with respect to the plane of the bend, we are in for a surprise. Think of our incident TE_{11} mode as the vector sum of an E-plane bend component and an H-plane bend component. When these two components combine on the other side of the bend, there will be a relative time phase shift between them. What we get at the output depends on this relative phase shift. If it is an integral multiple of 360°, we get our original TE_{11} mode back. If it is an odd multiple of 180°, we get a linearly polarized TE_{11} mode that is rotated 90° in the pipe from the input mode! If the time phase shift is an odd multiple of 90°, we get a circularly polarized TE_{11} mode. If it is anything else, we say it is "elliptically polarized," with the parameters of the ellipse determined by the phase shift. The practical message here is that if you want to maintain your original TE_{11} mode, use only E-plane or H-plane bends.

A Polarization Converter

If you want to make a plane-polarized to circularly polarized converter without resorting to a bend, consider the following idea. I have not actually tried it, but in principle, it should work.

Put your TE_{11}-excited copper waveguide horizontally in a vise, with the polarization direction of the mode at 45° to the direction of motion of the vise jaws. Monitor the polarization of the output, for example, by rotating a wire-grid polarizer (WGP) between the radiating end of the pipe and a detector. You should see a plane polarized wave at 45°, indicated by a sharp null in transmission when the wires of the WGP are parallel to the E field. Now tighten the vise jaws, squeezing the pipe to make it slightly elliptical in shape, while monitoring the polarization of the output.

You should see the output become elliptically polarized. This is indicated by some transmission variation with rotation of the WGP, but no deep nulls. At some degree of squeezing, the polarization becomes circular; this is indicated by no transmission variation with rotation of the WGP. The polarization becomes elliptical again upon more squeezing, and so forth. The result also depends on the width of the squeezed region; that is, the width of your vise jaws. As in the bend, this idea works because the guide wavelengths of the waves polarized along the major and minor axes of the slightly elliptical pipe are slightly different. I'd appreciate hearing from anyone who actually makes this work!

Matching and Coupling Devices

Many of the matching devices used

with rectangular TE_{10} waveguide such as the slide-screw tuner, the three-screw tuner, the E-H tuner and the magic-tee junction, should be realizable in circular TE_{11} waveguide, provided the polarization of the modes is kept in the symmetry planes of these devices. Some things won't work, though; for example, the offset matching posts used to improve some magic-tee junctions, or anything requiring asymmetry. I would think that multihole, parallel-propagating directional couplers could be made, but not two-hole cross-guide couplers. It will be interesting to see what the amateur community accomplishes!

Another kind of coax-to-waveguide transition is possible for TE_{11} circular waveguide. A shorting plane is used to close the end of the guide, just as in Wade's junction. An appropriate coax fitting (for example, an SMA) is mounted in the center of this end shorting plane, coaxial with the pipe. A small loop of wire (smaller than the diameter of the pipe) is connected from the center conductor of the coax fitting to the ground flange, with the plane of the loop perpendicular to the end-shorting plane. The TE_{11} mode that is excited will be polarized in the plane of the loop. For best match, adjust the wire size, loop diameter and possibly the shape of the loop. This kind of transition also works just fine in rectangular TE_{10} waveguide, provided you align the plane of the loop parallel to the short wall of the guide.

Comparing Figs 1 and 2 should suggest how a transition between round and rectangular waveguide could be made. Simply think of gradually distorting the conducting sidewalls of the guide from a circle to a rectangle. I have a commercially-made transition that looks like this, (also indicating that circular waveguide was actually used for something commercial at one time). I imagine you could make a reasonable substitute with thin, flexible copper sheet.

I think you could also make a reasonable transition with soldered-together rigid printed circuit board by a progression of shapes. First, flare the rectangle to a square, then gradually chop down the corners to make a regular octagon and finally butt the octagon to the circular pipe. I suggest soldering conducting flanges to both the octagon and the circular ends to make good contact in the butt joint. Longitudinal currents flow in the walls of the guide; these must have a good conducting path down the guide. Again, I haven't tried to make this kind of transition experimentally, but I would bet that it works pretty well.

The equation for the guide wavelength plotted by Wade in his Fig 2 is quite simple:

$$\lambda_g = \frac{\lambda}{\left[1-\left(\frac{\lambda}{\lambda_c}\right)^2\right]^{\frac{1}{2}}}, \qquad \text{(Eq 1)}$$

where $\lambda = c/f$ is the wavelength in free space, and λ_c is the cut-off wavelength. This equation is true for any TE or TM waveguide mode. All you need to know is the cutoff wavelength for the guide. The units of wavelength can be inches, millimeters, centimeters, meters or anything else, as long as they are consistent in the equation.

There are several older (that is, before 1970) textbooks that describe circular waveguides. My favorites are *Fields and Waves in Modern Radio*, (editions 1944, 1956) by Simon Ramo and John R. Whinnery, and *Fields and Waves in Communication Electronics* (editions 1965, 1984, 1994) by Simon Ramo, John R. Whinnery and Theodore Van Duzer, both published by John Wiley and Sons of New York. I used this text both as a student and as a teacher. All editions contain the mathematics of circular waveguides and illustrations of the fields, as well as much more, including antennas and microwave networks. The older editions should be available at booksellers stalls at swap meets.

Other Propagation Modes

An interesting challenge to the Amateur Radio community would be to utilize another circular waveguide mode, the circular TE_{01} mode, sometimes called the "circular electric mode," because the electric field lines form concentric circles in the waveguide cross-section. This is the only waveguide mode that exhibits a *decreasing* attenuation with increasing frequency. This amazing property was discovered by George Southworth in the 1930s and was developed extensively from 1950 to 1970 by the AT&T Bell Laboratories for cross-country transmission. The goal was to build a transmission system with 2 dB/km attenuation or less. The Bell Labs results are summarized in a special issue of the *Bell System Technical Journal* devoted to the WT4 system.[4] However, this system was never widely deployed because optical fibers with even less loss—and many fewer technical problems—became available first.

With the circular TE_{01} mode, you could envision sending power to a microwave antenna several hundred feet up on a tower with virtually no loss. The problems are manifold, though, as indicated by the 20 years of Bell Labs research. The TE_{01} mode is the fifth to cut on. (The cutoff wavelength is $0.82d$, where d is the diameter of the pipe). If you have a pipe large enough to support the TE_{01} mode, then the TE_{11}, TM_{01}, TE_{21} and TM_{11} are also cut on, and they are typically much more lossy. Any power scattered into these modes by dents, scratches or slight bends in the guide will be quickly lost. Worse still, the TM_{11} mode has exactly the same guide wavelength as the TE_{01} mode, so the coupling between these two modes is quite strong, and it doesn't take much of a dent or bend to do it. Even worse still, if you make the pipe big enough to realize very low loss (something like 10-inch diameter for the 2 dB/km value at 10 GHz), then you have *several hundred* modes cut on! Schemes for dealing with these problems were worked out (see Reference 4 and the references therein for the details), but they are tricky to implement.

I would like to thank Paul Wade for his interesting article. It's nice to know there are still some waveguide experimenters out there.

Notes

[1]P. Wade, W1GHZ, "Understanding Circular Waveguide—Experimentally," *QEX*, Jan/Feb 2001, pp 37-48.

[2]The *E plane* is defined by two lines: that of the electric-field direction in the center of the guide and the direction of propagation. The *H plane* is perpendicular to the E plane and defined by the direction of the magnetic field in the center of the guide and the direction of propagation.

[3]The "beer-can polaplexer" was described in several *QST* articles:

W. Baird, W6VIX, "A Radio Club for Microwave Enthusiasts," *QST*, Dec 1957, pp 45-47.

A. Bredon, W6BGK, "Let's Go Microwave," *QST*, June 1958, pp 11-14.

K. Peterson, K3KRU, "Practical Gear for Amateur Microwave Communication," *QST*, June 1963, pp17-20.

A very similar system of polarization and frequency-duplex operation using circular waveguide (but not beer cans) was "Experimental Transceiver for 5650 MHz," by C. J. Prechtel, W8DRR, *QST*, Aug 1960, pp 11-15.

You can view "The Polaplexer Revisited," by Ed Munn, W6OYJ, at **www.ham-radio.com/sbms/sd/ppxrdsgn.htm**.

[4]*Bell System Technical Journal*, Vol 56, pp 1825-2207, Dec 1977. (The entire issue is devoted to the WT4 Millimeter Wave System, 20 papers).

RF

By Zack Lau, W1VT

A Simple 10-Meter Satellite Turnstile Antenna

This very simple antenna actually took several years for me to develop. It is based on a very clever idea in Maxwell's *Reflections, Transmission Lines and Antennas*,[1] called the self-phased quadrature feed. In chapter 22, Walt points out that if you combine two orthogonal elements with impedances $R+jR$ and $R-jR$ in parallel, not only will the resulting impedance be R, there will be a 90° phase shift. This is exactly what you want for circularly polarized antennas.

I first experimented with loop antennas, attempting to make a very simple eggbeater style antenna. Unfortunately, no combination of different sized square loops seemed to work. I did find that you could add a loading coil to one loop—this provided a satisfactory design. However, loops for 10 meters are somewhat big and bulky.

I next tried phasing a pair of dipoles. Merely making one dipole larger and one smaller did not work—making the dipole larger increased the radiation resistance and making it smaller reduced the radiation resistance. The scheme was workable if you reduced the size of the dipole so that the impedance was $R-jR$. Then, an inductance of $2R$ could be added to a second dipole, so its impedance was $R+jR$. If a precise match to 50 Ω is desired, the shorter dipole can be further shortened until its resistive component is 50 Ω. Additional inductance can then be added to bring the reactance up to $-jR$. Fig 1 shows this optimized version. However, adding loading inductors requires a suitable weatherproof housing to be constructed. If you intend to actually build one, I'd suggest using one of the popular antenna analyzers to adjust each dipole to the proper impedance. I'd also

[1]This title is now out of print. Its successor, *Reflections II—Transmission Lines and Antennas* is published by WorldRadio Books (ISBN: 0-9705206-0-3). You can order it from the ARRL at **www.arrl.org/catalog**.

225 Main St
Newington, CT 06111-1494
zlau@arrl.org

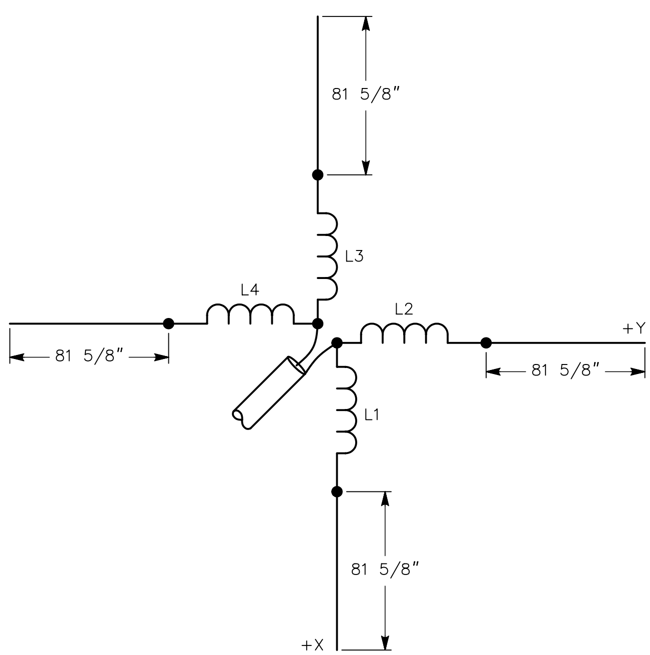


**Fig 1—Top view of self-phased turnstile using inductively loaded dipole elements. Height is 15 feet above ground.
L1, L3—0.42 µH L2, L4—0.70 µH**

10M Simple Satellite Turnstile Model Files

For Fig 1

```
;RHCP version
file : oscar.ant
Turnstile for Oscar Satellites
Over Ground
29.5 MHz
2 copper wires, feet
size1 = 6.8
height =15
height1=height+.00667
diam =0.00667
10     -size1 0 height      size1 0 height      diam
10          0 -size1 height1      0 size1 height1 diam

2 sources
Wire 1, center
wire 2, center
2 loads
wire 1, center 0.85 uH Q=100
wire 2, center 1.39 uh Q=100
```

For Fig 2

```
Over Ground
29.400 MHz
5 copper wires, inches
height = 120                  ;heights of inverted vee ends
hi = 75                       ;inverted vee height referenced to ends
h =hi+height                  ;height of the antenna apex
l = 91                        ;dipole lengths
l1 = 73                       ;length of inverted vee is
                                ;sqrt(hi^2+l1^2)
l2=0.2+l1
diam = .0808                  ;#12 wire
10 0 0.2 h                    l 0.2  h diam
10 0 0 h                      -l 0 h diam
10 0 0.2 h                    0 l2 height  diam
10 0 0 h                      0 -l1 height diam
1  0 0 h                      0 .2 h diam
1 sources
wire 5, center
```

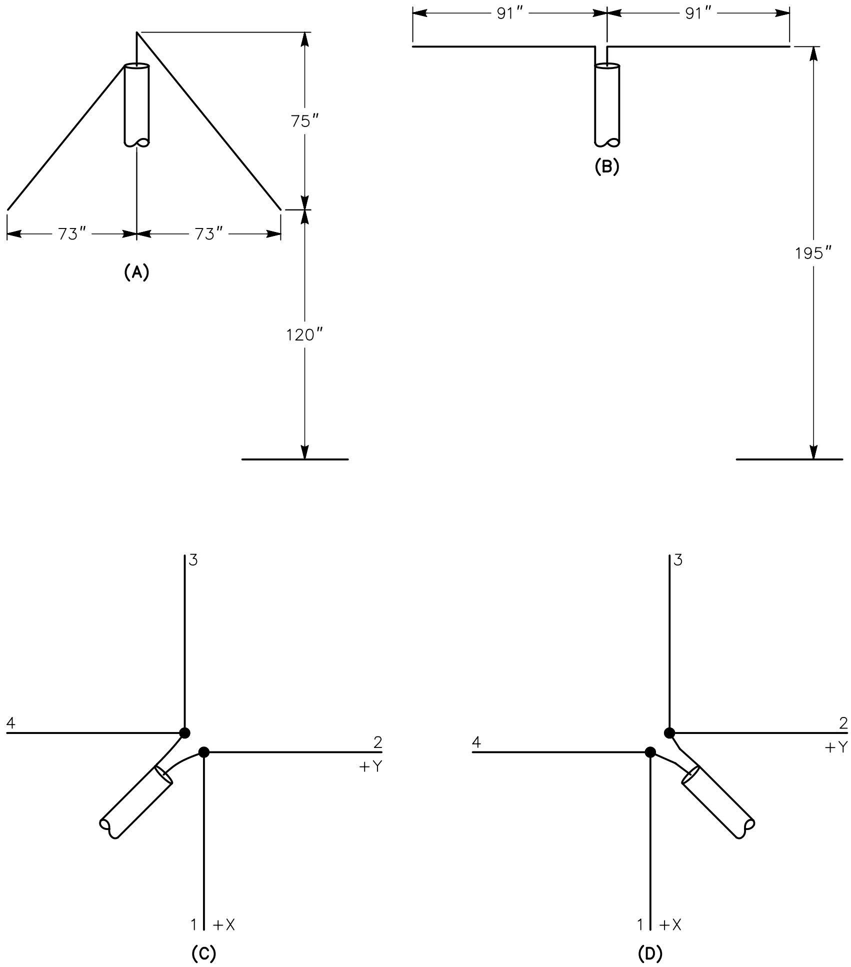
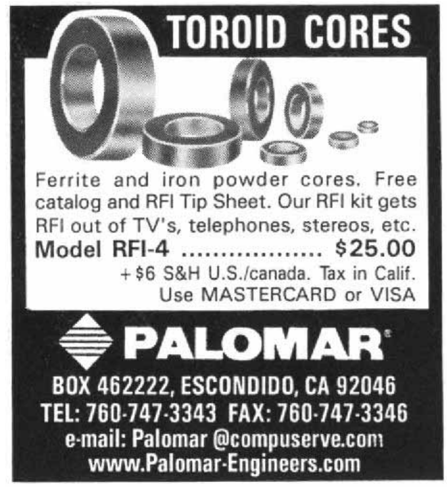


Fig 2—(A) Front view of self-phased turnstile antenna using a combination of dipole and inverted-V elements. (B) A side view of a self-phased turnstile using a combination of dipole and inverted-V elements. (C) Top view RHCP; (D) top view LHCP.

suggest using a simple choke balun made out of a coil of coax—shield currents may make the impedance measurement difficult.

My final idea was making an inverted V out of the longer dipole. By adjusting the angle between the elements, the radiation resistance can be adjusted to match the radiation resistance of the shorter dipole. By adjusting the element lengths and angle, the impedance $R+jR$ can be obtained. Thus, an extremely simple circularly polarized antenna can be constructed by combining a dipole and an inverted vee in parallel. Fig 2 shows the dimensions of a 29.4-MHz version for satellite work. It is tough to imagine a simpler right-hand circularly polarized antenna. It can be modified for left-hand circular polarization by swapping the feedpoint connections of just the dipole or inverted V.

However, for satellite work, it doesn't really matter which circularity you end up with, since linear polarization is typically used at the satellite on HF bands. A true circularly polarized HF antenna is just too big for amateur satellites. The linear to circular polarization mismatch is actually an advantage, since the mismatch is more constant. A fixed linear polarization on the ground would be subject to deeper fades as the orientation of the satellite antenna changed. This is useful to know—why worry about whether your antenna is LHCP or RHCP if it doesn't matter? □□



2001 Index

Issues 204 through 209 (All dates are 2001 unless otherwise indicated.)

Tech Notes (Bertini)

Upcoming Conferences

United States Postal Service

Statement of Ownership, Management, and Circulation

1. Publication Title	2. Publication Number	3. Filing Date
QEX	0 8 8 6 - 8 0 9 3	September 26, 2001
4. Issue Frequency	**5. Number of Issues Published Annually**	**6. Annual Subscription Price**
Bimonthly in: Jan, Mar, May, Jul, Sep, Nov	6	$24.00

7. Complete Mailing Address of Known Office of Publication (*Not printer*) (*Street, city, county, state, and ZIP+4*)
225 Main St., Newington, Hartford County, CT 06111-1494
Contact Person: Debra Jahnke
Telephone: (860) 594-0297

8. Complete Mailing Address of Headquarters or General Business Office of Publisher (*Not printer*)
225 Main St., Newington, Hartford County, CT 06111-1494

9. Full Names and Complete Mailing Addresses of Publisher, Editor, and Managing Editor (*Do not leave blank*)

Publisher (*Name and complete mailing address*)
Mark J. Wilson, 225 Main Street, Newington, CT 06111-1494

Editor (*Name and complete mailing address*)
Doug Smith, 225 Main Street, Newington, CT 06111-1494

Managing Editor (*Name and complete mailing address*)
Robert Schetgen, 225 Main Street, Newington, CT 06111-1494

10. Owner (*Do not leave blank. If the publication is owned by a corporation, give the name and address of the corporation immediately followed by the names and addresses of all stockholders owning or holding 1 percent or more of the total amount of stock. If not owned by a corporation, give the names and addresses of the individual owners. If owned by a partnership or other unincorporated firm, give its name and address as well as those of each individual owner. If the publication is published by a nonprofit organization, give its name and address.*)

Full Name	Complete Mailing Address
American Radio Relay League, Inc.	225 Main St., Newington, CT 06111-1494

11. Known Bondholders, Mortgagees, and Other Security Holders Owning or Holding 1 Percent or More of Total Amount of Bonds, Mortgages, or Other Securities. If none, check box → ☒ None

Full Name	Complete Mailing Address

12. Tax Status (*For completion by nonprofit organizations authorized to mail at nonprofit rates*) (*Check one*)
The purpose, function, and nonprofit status of this organization and the exempt status for federal income tax purposes:
☒ Has Not Changed During Preceding 12 Months
☐ Has Changed During Preceding 12 Months (*Publisher must submit explanation of change with this statement*)

PS Form **3526**, October 1999 (*See Instructions on Reverse*)

13. Publication Title: QFX
14. Issue Date for Circulation Data Below: Sep/Oct 00 - Jul/Aug 01 | Sep/Oct 01

15. Extent and Nature of Circulation		Average No. Copies Each Issue During Preceding 12 Months	No. Copies of Single Issue Published Nearest to Filing Date
a. Total Number of Copies (*Net press run*)		8,824	8,204
b. Paid and/or Requested Circulation	(1) Paid/Requested Outside-County Mail Subscriptions Stated on Form 3541. (*Include advertiser's proof and exchange copies*)	6,046	5,172
	(2) Paid In-County Subscriptions Stated on Form 3541 (*Include advertiser's proof and exchange copies*)	0	0
	(3) Sales Through Dealers and Carriers, Street Vendors, Counter Sales, and Other Non-USPS Paid Distribution	263	247
	(4) Other Classes Mailed Through the USPS	1,328	1,259
c. Total Paid and/or Requested Circulation [*Sum of 15b. (1), (2), (3), and (4)*]		7,637	6,678
d. Free Distribution by Mail (*Samples, complimentary, and other free*)	(1) Outside-County as Stated on Form 3541	85	85
	(2) In-County as Stated on Form 3541	0	0
	(3) Other Classes Mailed Through the USPS	63	4
e. Free Distribution Outside the Mail (*Carriers or other means*)		25	0
f. Total Free Distribution (*Sum of 15d. and 15e.*)		173	89
g. Total Distribution (*Sum of 15c. and 15f*)		7,810	6,767
h. Copies not Distributed		1,014	1,437
i. Total (*Sum of 15g. and h.*)		8,824	8,204
j. Percent Paid and/or Requested Circulation (*15c. divided by 15g. times 100*)		97.78%	98.68%

16. Publication of Statement of Ownership
☒ Publication required. Will be printed in the Nov/Dec 2001 issue of this publication. ☐ Publication not required.

17. Signature and Title of Editor, Publisher, Business Manager, or Owner — Date: September 26, 2001

I certify that all information furnished on this form is true and complete. I understand that anyone who furnishes false or misleading information on this form or who omits material or information requested on the form may be subject to criminal sanctions (including fines and imprisonment) and/or civil sanctions (including civil penalties).

Instructions to Publishers

1. Complete and file one copy of this form with your postmaster annually on or before October 1. Keep a copy of the completed form for your records.
2. In cases where the stockholder or security holder is a trustee, include in items 10 and 11 the name of the person or corporation for whom the trustee is acting. Also include the names and addresses of individuals who are stockholders who own or hold 1 percent or more of the total amount of bonds, mortgages, or other securities of the publishing corporation. In item 11, if none, check the box. Use blank sheets if more space is required.
3. Be sure to furnish all circulation information called for in item 15. Free circulation must be shown in items 15d, e, and f.
4. Item 15h., Copies not Distributed, must include (1) newsstand copies originally stated on Form 3541, and returned to the publisher, (2) estimated returns from news agents, and (3), copies for office use, leftovers, spoiled, and all other copies not distributed.
5. If the publication had Periodicals authorization as a general or requester publication, this Statement of Ownership, Management, and Circulation must be published; it must be printed in any issue in October or, if the publication is not published during October, the first issue printed after October.
6. In item 16, indicate the date of the issue in which this Statement of Ownership will be published.
7. Item 17 must be signed.

Failure to file or publish a statement of ownership may lead to suspension of Periodicals authorization.

PS Form **3526**, October 1999 (*Reverse*)

Letters to the Editor

Beyond Fractional-N, Part 2
(May/Jun 2001)

This is an informative and interesting article. There appear to be some "typos" in Fig 4, a schematic. If you look closely, you can see that the filters in the dc bus are shown as band-pass instead of low-pass (FL 2, 3, 4 and 5). In addition, the wires are crossed between the charge pump and the op amp. Moreover, if C38 and C71 = 0.68 µF, the pole with R43 is at 5 kHz. This makes the loop unstable. The circuit would make most op amps go unstable because of too small a load resistance or too big a load capacitance.

Then in Fig 6, there is a usage I don't like, but I am not sure it's wrong. ω_n is labeled hertz, when it should be radians/second. Could you clear up these points in an otherwise excellent article?—*William Cross, KA0JAD, 7100 E Evans Ave #411A, Denver, CO 80224*

KW7CD responds:

I am traveling, but have a copy of the article here. In this case, the proof is in the pudding. This synthesizer works well and outperforms an FT-1000D synthesizer—in phase noise alone—by at least 10 dB. I have done the measurements with world-class equipment.

It is possible that one of the draftsman (mine or ARRL's) got carried away with the EMI filters showing band-pass rather than low-pass. These filters are specific EMI units that are not usually found in amateur gear. They were chosen specifically to reject certain frequency ranges, so they could be low-pass types.

In addition, I think the "wires-are-crossed" comment is wrong. There is no mistake that I see on the runs of the charge-pump signals to the op amp. CPR, pin 20 of the PLL (U4) goes to pin 2 of OP27 (–) and CPO, pin 21 goes to pin 3 (+) of the op amp (U5). It may be confusing to some, but the draftsman made the best use of the space, that's all. The lines cross, but they are connected correctly.

Strictly speaking, ω_n is expressed in radians per second. The math is otherwise correct, though. Some op amps may go unstable depending on the R43 value; however, this circuit works and is stable in this case. The trap combination is useful and is used extensively by the industry and even in amateur equipment on the market today. This circuit overcomes the so-called "hump" in phase noise possible at the intersection of the VCO's natural phase-noise point and the loop filter's corner frequency, when we have a very tight loop such as this one. To eliminate this hump, we usually move the corner frequency of the loop filter out; however, this spoils the phase noise of the synthesizer. In our case, we want to have our cake and eat it, too.

We wouldn't need this circuit if we could afford to move the corner frequency of the loop out of the picture, but that would worsen the phase-noise performance. We need to reach a compromise between the loop's lock-up time and the close-in phase-noise performance of the synthesizer. These parameters are contradictory to say the least, as any experienced PLL engineer would recognize. We need to have a great appreciation for the imperfect tradeoffs that are being made in these circuits. The result in this case is a close-in phase-noise measurement on the DDS-driven PLL loop of better than –130 dBc at 1-kHz offset. This phase-noise performance rivals anything being manufactured with this very wide frequency bandwidth.

Equally important are the implications about the spurious (harmonic and nonharmonic) performance in a design such as this. For the record, the measured harmonic spurious of the published DDS-driven PLL is –80 dBc, while the nonharmonic spurious is –98 dBc. The BFO DDS part of the synthesizer has a harmonic spurious performance of –90 dBc for the second harmonic and a –95 dBc for the third harmonic. The nonharmonic spurious performance of the BFO DDS is –110 dBc. I think these are outstanding numbers by any standard.—*Cornell Drentea, KW7CD, 757 N Caribbean, Tucson, AZ 85748;* **CDrentea@aol.com**

Empirically Speaking
(Mar/Apr 2001)

This is in response to your editorial suggesting that analysis and simulation are not the end of the job, but that measurement is also needed. I have found that the 30-year-old book, *Communications Circuit: Analysis and Design* by Clarke and Hess (Reading, Massachusetts: Addison-Wesley, 1971) is still the best reference on the design of oscillators. It has a detailed analysis of the clamp-based JFET—the building block for almost all oscillators built by hams. Their analysis makes several simplifying assumptions: Are they valid? To find out, I made some measurements on an MPF-102 and compared them with the nifty graph in Clark-Hess. The following is the result.

In Fig 1, the X axis shows input voltage *V1* (peak, not RMS) relative to pinch-off voltage *V*p (also known as $V_{GS(OFF)}$). An X-axis value of 0.5 represents an input voltage whose peak-to-peak magnitude is just equal to the pinch-off voltage; an input voltage larger than this cuts off the drain

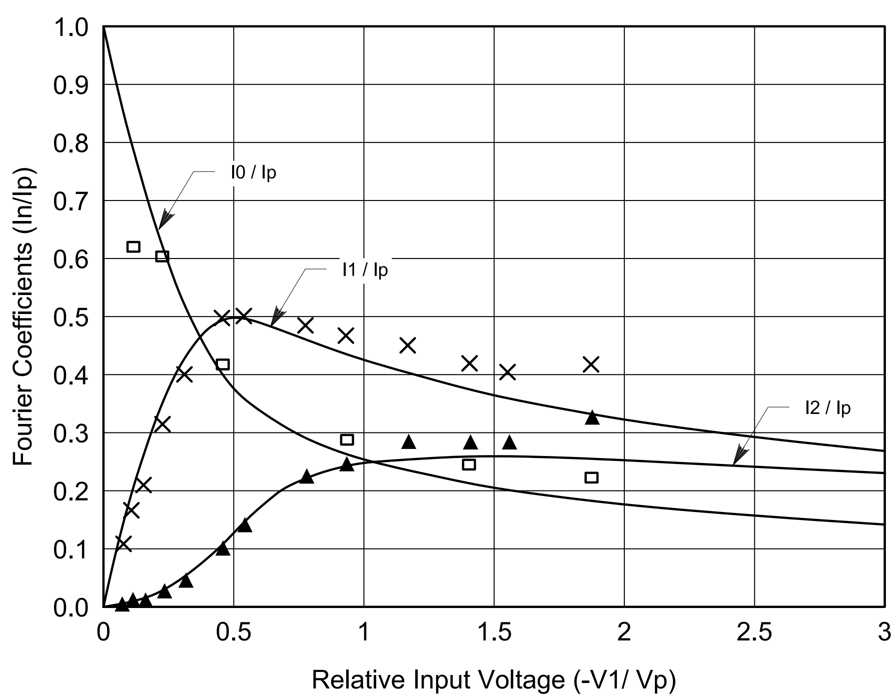


Fig 1—N1EKV's measured performance of an MPF-102 (see text).

current for part of each cycle.

The Y-axis shows the normalized Fourier coefficients for dc, fundamental and second-harmonic components of the JFET drain current. They are normalized as peak (not RMS) current at each frequency component: *I0*, *I1* and *I2*, relative to the peak instantaneous current.

The theoretical curves assume the gate-source diode in the JFET is ideal, having zero forward voltage drop; so the peak current is equal to I_{DSS}. In a real JFET, the gate can go about 0.7 V positive; so the peak current, *I*p, is greater than I_{DSS} and the input voltage *V*p is greater than $V_{GS(OFF)}$. For the real JFET, whose data are plotted, I_{DSS}=6.4 mA but *I*p=10.4 mA (measured). $V_{GS(OFF)}$=–2.5 V but *V*p=–3.2 V.

We learn from the curves that the best operating point for a JFET in an oscillator is at 0.5 on the relative-input-voltage scale. A higher input voltage gives *less* fundamental output and greater second-harmonic output.—*Byron E. Blanchard, N1EKV, 16 Round Hill Rd, Lexington, MA 02420-3634*

A Spreadsheet for Remote Antenna-Impedance Measurement (Sep/Oct 2001)

Ron Barker, G4JNH's, article assumes a characteristic impedance of the line and transmission line loss (*L*= known line loss). Actually, with two measurements of the type Barker proposes, it is possible to obtain both the transmission line loss and the transmission line characteristic impedance. Barker supplies the results of two such measurements, so the calculation is possible.

If the transmission line equation is solved for the hyperbolic tangent of the propagation factor, there results:

$$\tanh(\gamma l) = Z_0 \frac{Z_L - Z_G}{Z_L Z_G - Z_0^2} \qquad \text{(Eq 1)}$$

where the subscripts L and G refer to the impedance at the load (antenna) and generator (in), γ is the complex propagation factor and *l* is the line length. By equating the hyperbolic tangents for two measurements, a solution for the characteristic line impedance is obtained:

$$Z_0^2 = \frac{(Z_{L2} - Z_{G2}) Z_{L1} Z_{G1} - (Z_{L1} - Z_{G1}) Z_{L2} Z_{G2}}{(Z_{L2} - Z_{G2}) - (Z_{L1} - Z_{G1})} \qquad \text{(Eq 2)}$$

where the subscripts refer to the two sets of measured values. Having found Z_0, its value can be substituted back into the first equation to find the propagation constant. Using Barker's measurements,

	Measurement 1	*Measurement 2*
R_L	20	120
R_G	71.1	24.9
X_G	33.3	–13.3

I find

$Z_0 = 47.6 - j0.372\Omega$

and $\tanh(\gamma l) = 0.796 + j2.530$

Taking the inverse hyperbolic tangent,

$\gamma l = \alpha l + \beta l = 0.1004 + j1.2219$

Converting from nepers and radians to decibels and wavelengths, the measured attenuation is 0.872 dB and the measured line length is 0.195 λ. Checking these results by calculation, the impedance at the input of the line with a 120-Ω load yields $Z_G = 24.94 - j13.31$. The agreement seems satisfactory.

These results compare well with Barker's results. Only the measured characteristic impedance differs noticeably from the values assumed or measured by Barker. I find the agreement between the results of the measurements and those assumed/calculated by Barker to be quite amazing. His impedance measurements must have been done with considerable skill.—*Albert E. Weller, WD8KBW, 1325 Cambridge Blvd, Columbus, OH 43212;* **a-nweller@worldnet.att.net**

The author responds:

I would like to thank Mr. Weller for his thoughtful contribution. My objective was to devise a spreadsheet for a procedure that amateurs had been practicing for years using paper Smith charts. There was no intent to improve on that method other than to make it quicker, easier and less error-prone. I believe the spreadsheet meets that objective.

I had looked at the transmission-line equation as it is presented on pages 27-29 and 27-30 of *The ARRL Antenna Book* 17th edition, but I didn't have the necessary mathematical know-how to handle the complex hyperbolic functions; so the spreadsheet was based on the older and easier-to-manipulate equations.

Since last year, when I submitted the article, I have conducted many more tests using both the 20-Ω and 120-Ω terminations and in many cases, the difference in Smith-chart line length has been higher than the 0.006 reported in the article; frequently in the region of 0.010 and occasionally as high as 0.015. In all cases, the characteristic impedance of the line was taken to be as specified and the loss read from the graphs in *The ARRL Antenna Book*.

These observations and the earlier attempts to determine line loss from the known impedances at both ends, which I mentioned in the article, led me to conclude that I needed to use measured, rather than specified, values for Z_0 and to be able to handle a Z_0 having a reactive component. I wasn't making much progress until I read Dr. Steven Best's excellent article in the Jan 2000 issue of *QEX*. When I saw the transmission-line equation as given in his Eq 20, I realized that it was in a form that could manipulated—without understanding the mysteries of hyperbolic functions—to reveal complex Z_0, loss and electrical length from two impedance measurements with known terminations, as Mr. Weller describes in his letter.

In addition to the two solutions that he shows, I did a third solution for Z_L so the antenna impedance could be derived once the values of Z_0 and $\tanh(\gamma l)$ had been established; however, I chose a different approach to get at the values of line length and loss than he did. Whereas he took $\tanh^{-1}$, I used the reflection-coefficient approach. I used Dr. Best's Eq 2, which gives the reflection coefficient in rectangular form, and converted it into polar form. The electrical length is then equal to half of the difference between the angles of reflection coefficient at the two ends of the line and the loss is equal to 10 $\log_{10}$ of the ratio of the two magnitudes of reflection coefficient.

I have compiled a new spreadsheet to do all of the above calculations taking advantage of the complex-algebra facility in MS *Excel* and have used it to repeat the calculations conducted by Mr. Weller. Here they are:

$Z_0 = 47.6 - j0.372\Omega$

$\tanh(\gamma l) = 0.796 + j2.530$

loss = 0.872 dB

line length = 0.194 λ

The measurements on the dummy antenna from Table 1 in the article were also entered into the new spreadsheet and gave an impedance of 32.5–*j*43.8 Ω compared to a known 36.0–*j*48.0 Ω.—*Ron Barker, G4JNH;* **ron.g4jnh@talk21.com**

A Correction

It appears that Eq 4 for input reactance is incorrect as printed on page 4 of the Sep/Oct 2001 *QEX* article. The numerator at the upper right should be a sum and not a difference of terms, as shown here.—*Dr. Bill Echols, WA2NYR, 111 Carpenter Dr, Jackson, MS 39212-9671;* **bill.echols@ieee.org**

$$X_{IN} = \frac{X_a\left(1-\tan^2\theta\right) + \left(Z_\eta - \frac{R_a^2 - X_a^2}{Z_0}\right)\tan\theta}{\left(1 - \frac{X_a}{Z_0}\tan\theta\right)^2 + \left(\frac{R_a}{Z_0}\tan\theta\right)^2}$$

(Incorrect Eq 4)

$$X_{IN} = \frac{X_a\left(1-\tan^2\theta\right)+\left(Z_\eta - \frac{R_a^2+X_a^2}{Z_0}\right)\tan\theta}{\left(1-\frac{X_a}{Z_0}\tan\theta\right)^2+\left(\frac{R_a}{Z_0}\tan\theta\right)^2}$$

(Correct Eq 4)

The Art of Making and Measuring LF Coils

(Sep/Oct 2001)

Congratulations on the inclusion of the LF article in the Sep/Oct issue. While the article gives some valuable insights into coil design, I feel it is necessary to highlight a major practical shortcoming with respect to the construction of the coil and former.

The author's assertion that a wood former has low losses at LF is indeed true; however, from the text it appears that only 4 W have been used, essentially in lab-based experiments. Given that 100 W at 136 kHz is a common practical minimum (outside the USA—*Ed*), 300 W is average and 1 kW is not unusual, I suggest that wood is completely unsuitable for practical use with an LF transmitter.

Wood naturally absorbs moisture and as the coil is usually placed at the base of the antenna, this will rapidly reduce the Q of the coil, increasing losses. The authors suggest that 1.5 kV was developed across the coil with 4 W. Even if the wood is varnished against the ingress of water, 100 W+ will easily develop over 10 kV in a typical installation. These very high voltages will cause arcing to the wooden structure. As the surface carbonizes, the conductivity increases rapidly and the thing catches fire. This is a very real safety risk. Wood should be avoided at all costs. Plexiglas or similar polymers are far superior.

Here's a final cautionary note: Close-wound, enameled copper wire will have insufficient insulation to withstand the voltage developed between adjacent turns. A variable 0.5 to 3 mH basket-weave coil constructed from Litz wire for my LF station had to be dismantled and reassembled with PVC sleeving covering the Litz. The original coil failed at 200 W; the rebuilt unit has been tested to 1200 W without problem.—*David Bowman, G0MRF;* **www.g0mrf.freeserve.co.uk**

The authors respond:

We are happy to read your questions; they are very important and useful for a clearer understanding of LF topics. Effectively, many of our coils have been realized only for low-power, lab-based experiments; but the final coil of Fig 8 has been used at the 120-W level ($R_G + R_S$=30 Ω, I_{ANT}=2 A), sheltered by a pail. The support is wooden; but practically speaking, this is an *air* coil! Another, higher-power version of the same coil, for IK2WAQ, was realized on a Delrin support.

We use only highly insulated Litz wire for our operational load coils. With 6-8 kV @ 137 kHz on about 100 turns, we have only 60-80 V per turn. Many thanks!—*Paolo Antoniazzi, IW2ACD,* **paolo.antoniazzi@st.com**; *Marco Arecco, IK2WAQ*

A Mathematical Model for Regenerative RF Amplifiers

(Jul/Aug 2001)

Having designed, built and experimented with regenerative circuits for more than 50 years, I found Bill Young's mathematical analysis extremely interesting. The circuit chosen for analysis, with feedback via a coupling or "tickler" coil connected to the drain, is basically the same as that patented by Armstrong in 1913 and used in the first regenerative receivers. Controlling feedback by means of a shunting resistor (R_R) across the feedback winding is an improvement on the series-connected variable capacitor used in the earliest circuits. With this latter method—rotating the reaction control, as it was known—the tuning of the receiver tended to shift and added to the difficulties of adjustment.

Unless I have missed something, Bill concentrates on the factors influencing the gain of the FET amplifier stage and ignores the direct effect on tuned-circuit Q of the shunt resistor used as the feedback control. When considering regenerative circuits, one should always remember that the dramatic improvement in performance results solely from the increase in the Q of the single tuned circuit. (The signal magnification and selectivity displayed by a tuned circuit are directly related to its Q.) The application of positive feedback, also known as reaction or regeneration, reduces the resistive losses in the tuning coil, and basic Q factors of around 100 can thereby be increased to more than 1000. Under laboratory conditions, positive feedback has been used to increase the Q of a coil to 8000; but levels as high as this cannot be realized in practical radio receivers.

In a receiver where separate Q-multiplier and detector stages are used—and I agree with Bill that this is usually the best arrangement—the Q-multiplier amplifier (FET, bipolar transistor or valve) need only have a very modest amount of gain: just sufficient to almost overcome the losses in the coil. If reception of SSB signals is required, there must be enough gain to completely overcome resistive losses and maintain oscillation, so as to restore the carrier.

I have found that the smooth control of regeneration, up to and through the onset of oscillation, with complete freedom from backlash or detuning effects, is best achieved by adopting the Hartley rather than the Armstrong oscillator, and by adjusting the gain of the amplifier to control the level of feedback. With this arrangement, feedback is taken from the source (or emitter or cathode) and connected to a tap on the tuning coil via the source-bias resistor. A separate feedback winding may be adopted when this is more convenient.

The use of a dual-gate MOSFET is ideal for circuits of this kind, as its gain can be varied by adjusting the voltage on its gate 2. If a JFET must be used, gain and regeneration are best controlled by adjusting its drain voltage.

Over the years, I have published a number of circuits for regenerative receivers, preselectors and loop aerials in British journals. They have all used the Hartley circuit with regeneration control being effected by adjusting the gain of the maintaining amplifier. Provided that the source bias resistor is selected (use a preset) to suit the particular MOSFET used, no difficulty is experienced in obtaining smooth regeneration that can easily be set close to the onset of oscillation, and which is so necessary to the efficient operation of this type of circuit. I would urge your readers to try a Q multiplier of this kind.—*Raymond Haigh, Technical Author;* **Raymo96725@aol.com**

WD5HOH responds:

Thanks for your comments on my article. Yes, I neglected the tuned circuit and its Q entirely. My model is just a beginning. A model that takes into account the tuned circuit would yield more information, of course. I have also been experimenting with regenerative receivers, off and on over several decades. I tend to try to develop as much gain as possible, and I usually place an RF amplifier stage ahead of the regenerative stage. Narrow bandwidth at high gain follows if sufficiently stable control of regeneration has been achieved. I didn't think of my regenerative amplifier stage followed by a detector as a Q multiplier; but, of course, it is.

There is a British book about experimental valve (vacuum-tube) receivers that I remember reading in junior high school in the 1950s. I do not remember the author's name, but it contained many variations on the basic regenerative receiver circuit. Do you remember such a book?

Also, do you recall ever observing that a valve or tube regenerative stage is photosensitive? I unintentionally directed the beam of a flashlight onto the glass envelope of an oscillating regenerative stage years ago and noticed that the beat note shifted.—*Bill Young, WD5HOH, 343 Forest Lake Dr, Seabrook, TX 77586;* **blyoung@hal-pc.org**

Bill,

Thanks for your reply. I suspect that a number of enthusiasts are still experimenting with this century-old circuit, perhaps because it gives so much for such a modest outlay in components. I too use a stage of RF amplification to isolate the *Q* multiplier from the aerial and to provide a measure of gain. I have found that a good arrangement is a PNP transistor in grounded-base mode with its collector taken to the negative rail via the tuning coil.

Your comment about the photosensitive nature of valve regenerative detectors is most interesting, and it's something that I have never encountered before. At the risk of seeming immodest, I have to say I have quite a collection of literature on the subject but have not seen it mentioned: There is always something new.

The book you mention also sounds most interesting, but I can't identify it from your brief recollection. I, too, remember and sadly miss interesting publications read in my youth, when valve regenerative receivers were still built in some numbers by amateurs.—*Raymond*

Deconvolution in Communication Systems
(Sep/Oct 2001)

The title of Ref 8 was even funnier than what we printed. Our spelling checker threw up on it, as you will understand. The correct title is "The Quefrency Alanysis of Time Series for Echos: Cepstrum, Pseudo-Autocovariance, Cross-Cepstrum and Saphe Cracking," *Proceedings of the Symposium on Time Series Analysis* (New York: John Wiley & Sons, 1963)—*Doug Smith, KF6DX,* QEX *Editor*

Will a Digital Audio Amplifier work in a Ham Environment?
(Tech Notes, Sep/Oct 2001)

I started reading your article in the Sept/Oct 2001 issue of *QEX* magazine on class-D audio amplifiers. I am sorry to report that it looks like the graphs in Fig. 2 are totally confusing. The graph at the top with the sine waves superimposed over the triangle waves is confusing when comparing it to the text. I think curve B should be A, curve C should be B, curve D should be C, and so on. Line D seems to agree with the last line in the second paragraph on page 54. I am sorry all these errors occurred, as it may totally confuse anyone not familiar with the operation of comparators. Do you have a list of corrections or corrected Fig 2 available anywhere? I haven't finished reading the article yet, but I am enjoying the work you did in trying to describe class-D audio amplifiers and using the TPA2000D2 IC.—*Todd Roberts, WD4NGG, PO Box 21413, Hilton Head Island, SC 29925-1413;* **ToddRoberts2001@aol.com**

Sam responds

You are correct that the text related to Fig 2 is incorrect. It should have read as follows starting with the first column on page 54. The figures themselves are correct when the words are changed. Thanks for bringing this to my attention.

"…comparator. The comparator output goes high when one input signal is greater than the other, and is 0 when it is lower. Fig 2 shows how the comparator output is comprised of a string of pulses whose width is dependent on the amplitude of the incoming signal. Analyzing just how the volume and frequency information was encoded into the pulse stream was educational. Referring to Fig 2 again, for a low-volume signal of set frequency (see the sine wave "B" in Fig 2A, and the pulse waveform in Fig 2B), the pulse widths do not vary much when going from high to low volume. Sine wave "C" in Fig 2A shows a signal with the same frequency, but at a much higher magnitude. Here the pulse width (Fig 2C) varies a lot. Sine wave "D" has the same magnitude as sine wave "C," but at twice the frequency. The pulse-width variations (Fig 2D) are als very large, but they happen twice as often."

The caption under Fig 2 also needs to be changed. Curve A becomes Curve B; Curve B becomes Curve C; Curve C becomes Curve D.—*Sam Ulbing, N4UAU, 5200 NW 43rd St Ste 102-177, Gainesville, FL 32606;* **n4uau@arrl.net**

The Q of Single-Layer, Air-Core Coils: A Mathematical Analysis
(Sep/Oct 2001)

There is a minor error in Eq (13) on page 34. The numerator 25530.29 should be 25330.29.—*George Murphy, VE3ERP, 77 McKenzie St, Orillia, ON L3V 6A6, Canada;* **ve3erp@encode.com**

Next Issue in *QEX/Communications Quarterly*

We are finding that articles placed around the Web are sometimes the only evidence potential readers have that *QEX* exists. We also find that experimenters in faraway places—such as Russia, for example–are hungry for information about homebrewing and state-of-the-art circuits and systems. Often, the Web is their best connection to such information. Therefore, we are considering a change to our publication agreements with authors that would allow greater freedom to post or distribute copies of published articles. Your articles and letters have been outstanding, and they certainly deserve a wider readership. In keeping with our philosophy of promoting the free exchange of ideas among experimenters, we hope to be able to grant greater latitude to you to help us get the word out.

We are trying to lay the groundwork for a better *QEX*. As mentioned previously, part of that involves greater editorial effort on our part. You may see us enlist some additional help. We are fortunate to have volunteers who are ready and willing to assist.

As we start our third decade, it is notable that one of the oldest forms of radio receiver continues to be one of the most popular. In the Jan/Feb 2002 issue, Andre Jamet, F9HX, takes the super-regen to SHF where it stands out as a practical way to build a microwave receiver.

Jack Hardcastle, G3JIR, takes a thorough look at crystal parameters and ways we can specify and measure them. Those are valuable things to learn that sometimes befuddle even experienced RF designers.

Upcoming Conference

Now is the time to begin preparing papers for the Sixteenth Annual Southwest Ohio Digital Symposium. We'd like presentations about and demonstrations of Spread Spectrum, APRS, Packet, PSK-31, Throb, Picolo, MFSK-16, Hellschriber, RTTY, Coherent CW, Wigwag, Semaphore—whatever *you* are working, or working on!

This oldest regional digital symposium will be held Saturday 12 January, 2002, 9 AM to 4 PM, at Thesken Hall, Middletown Campus, Miami University, Middletown, Ohio. It is sponsored by The Center for Chemistry Education, Miami University, Middletown, DIAL Radio Club and The Ohio Packet Council. For more information contact Hank Greeb, N8XX, **n8xx@arrl.net**. Visit us on the Web at **w3.one.net/~rkuns/swohdigi.html**.

EZNEC 3.0

All New *Windows* Antenna Software by W7EL

EZNEC 3.0 is an all-new antenna analysis program for Windows 95/98/NT/2000. It incorporates all the features that have made ***EZNEC*** the standard program for antenna modeling, plus the power and convenience of a full Windows interface.

EZNEC 3.0 can analyze most types of antennas in a realistic operating environment. You describe the antenna to the program, and with the click of a mouse, ***EZNEC 3.0*** shows you the antenna pattern, front/back ratio, input impedance, SWR, and much more. Use ***EZNEC 3.0*** to analyze antenna interactions as well as any changes you want to try. ***EZNEC 3.0*** also includes near field analysis for FCC RF exposure analysis.

See for yourself

The ***EZNEC 3.0*** demo is the complete program, with on-line manual and all features, just limited in antenna complexity. It's free, and there's no time limit. Download it from the web site below.

Prices - Web site download only: $89. CD-ROM $99 (+ $3 outside U.S./Canada). VISA, MasterCard, and American Express accepted.

Roy Lewallen, W7EL *phone* 503-646-2885
P.O. Box 6658 *fax* 503-671-9046
Beaverton, OR 97007 *email* w7el@eznec.com

http://eznec.com

Mike's Electronics

ICOM®

Receivers

1001 North West 52nd St
Ft Lauderdale, FL 33309

Phone: 800-427-3066
Fax: 954-491-7011
mspivak@bellsouth.net

QEX

American Radio Relay League
225 Main Street
Newington, CT 06111-1494 USA

For one year (6 bi-monthly issues) of **QEX**:

In the US

☐ ARRL Member $22.00
☐ Non-Member $34.00

In Canada, Mexico and US by First Class mail

☐ ARRL Member $35.00
☐ Non-Member $47.00

Elsewhere by Surface Mail (4-8 week delivery)

☐ ARRL Member $27.00
☐ Non-Member $39.00

Elsewhere by Airmail

☐ ARRL Member $55.00
☐ Non-Member $67.00

QEX Subscription Order Card

QEX, the Forum for Communications Experimenters is available at the rates shown at left. Maximum term is 6 issues, and because of the uncertainty of postal rates, prices are subject to change without notice.

Subscribe toll-free with your credit card **1-888-277-5289**

☐ Renewal ☐ New Subscription

Name ______________________ Call __________

Address ______________________________

City ____________ State or Province ________ Postal Code ________

Payment Enclosed ☐

Charge:

☐ MasterCard ☐ VISA ☐ American Express ☐ Discover

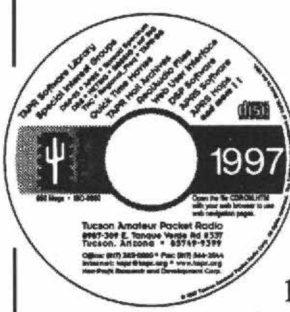
Account # ______________________ Good thru __________

Signature ______________________ Date __________

Remittance must be in US funds and checks must be drawn on a bank in the US. Prices subject to change without notice.

11/98

Join **the effort in developing Spread Spectrum Communications for the amateur radio service.** *Join* TAPR and become part of the largest packet radio group in the world. TAPR is a non-profit amateur radio organization that develops new communications technology, provides useful/affordable kits, and promotes the advancement of the amateur art through publications, meetings, and standards. Membership includes a subscription to the *TAPR Packet Status Register* quarterly newsletter, which provides up-to-date news and user/technical information. Annual membership US/Canada/Mexico $20, and outside North America $25.

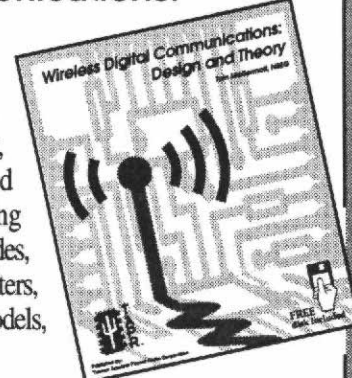


TAPR CD-ROM

Over 600 Megs of Data in ISO 9660 format. TAPR Software Library: 40 megs of software on BBSs, Satellites, Switches, TNCs, Terminals, TCP/IP, and more! 150Megs of APRS Software and Maps. RealAudio Files. Quicktime Movies. Mail Archives from TAPR's SIGs, and much, much more!

Wireless Digital Communications: Design and Theory

Finally a book covering a broad spectrum of wireless digital subjects in one place, written by Tom McDermott, N5EG. Topics include: DSP-based modem filters, forward-error-correcting codes, carrier transmission types, data codes, data slicers, clock recovery, matched filters, carrier recovery, propagation channel models, and much more! Includes a disk!

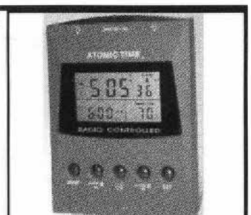


Tucson Amateur Packet Radio
8987-309 E. Tanque Verde Rd #337 • Tucson, Arizona • 85749-9399
Office: (940) 383-0000 • Fax: (940) 566-2544 • Internet: tapr@tapr.org www.tapr.org
Non-Profit Research and Development Corporation

ATOMIC TIME™

...self setting
...correct time
...atomic clock

SALE
Atomic Watch hard mineral lens, hi-tech polymer case black leather band $109.95



atomic dual alarm clock w. temperature day and date, black 3.5x4.5x2 $29.95

World s most exact time... atomic clocks, atomic watches and weather stations

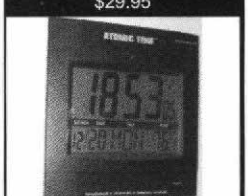
- for any time zone
- synchronized to the u.s. atomic clock in colorado
- accurate to 1sec. in 1 mil. years
- engineered in germany

complete line of atomic clocks

JUNGHANS MEGA CERAMIC Watch
JUNGHANS MEGA CARBON Watch
JUNGHANS MEGA CLOCKS
JUNGHANS SOLAR WATCHES
ATOMIC SPORTS WATCHES
ATOMIC SCHOOL/OFFICE CLOCKS
ATOMIC INDUSTRIAL CLOCKS
Oregon Scientific Weather Stations, Weather Forecast, World Time, NOAA Radios, Radio Controlled Clocks...

call for our FREE Brochure
or go to **www.atomictime.com**
credit card orders call toll free
1-800-985-8463
30 Day Money Back Guarantee
send checks incl. s&h $6.95 to
ATOMIC TIME, INC.
1010 JORIE BLVD.
OAK BROOK, IL 60523

atomic radio with 2 alarms and tempature, day, date, LCD $39.95

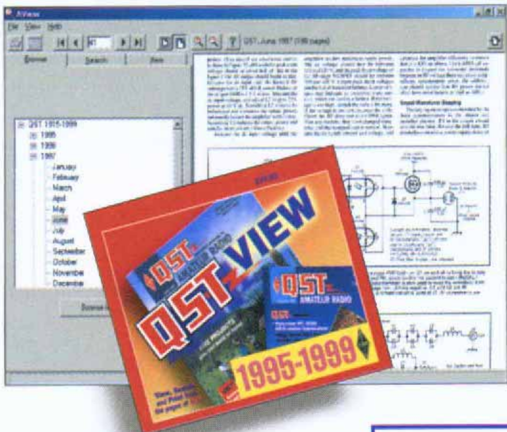


jumbo digit atomic clock w. temperature & day and date,wall or desk 8.5 x8.5 x1 • $49.95

NEW
Junghans atomic carbon, stainless bezel, sapphire lens LCD day, date - carbon/leather band • $279.00

black arabic 12 wall clock for home or office • $59.95 (wood $69.95)

CD-ROM Collections



NEW 1995-1999 set!

QST View

The perfect way to enjoy past issues of *QST*! Each CD-ROM set includes all the ads, articles, columns and covers—scanned to provide black-and-white images that can be viewed on your computer screen or printed. Every Page!

Features:

Search for articles by title and author.

Select specific year and issue.

Browse individual articles or columns.

QST View CD-ROM sets

Years	Order No.
1915-29	#7008
1930-39	#6710
1940-49	#6648
1950-59	#6435
1960-64	#6443
1965-69	#6451
1970-74	#5781
1975-79	#5773
1980-84	#5765
1985-89	#5757
1990-94	#5749
1995-99	#8497

$39.95* PER SET

Requires Microsoft *Windows*.

NEW 5-year set: 1995-1999

- **IMPROVED** viewer (works with all sets!)
- **BETTER** Search & Browse
- **SUPERIOR** image quality!

SAVE over $80 when you order the **entire *QST* View Collection. Includes all 12 CD-ROM sets!**

QSTV ~~$479.40~~ **Only $399**

Call our toll-free number, ***1-888-277-5289****, to order or to find a dealer near you!*

ARRL Periodicals on CD-ROM

An entire year of ARRL journals on CD-ROM!

Each CD includes: QST — **The official membership journal of ARRL**

NCJ — **National Contest Journal**

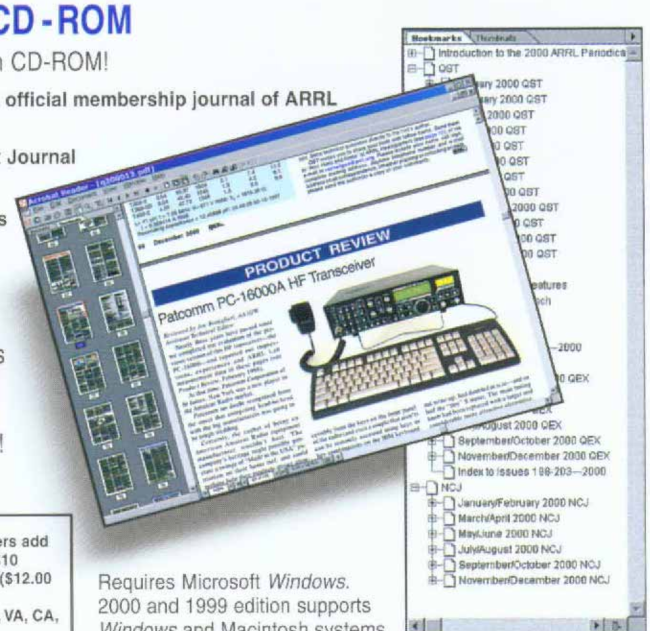
QEX — **Forum for Communications Experimenters**

Features:

Search the full text of every article by entering titles, call signs, names—almost any word.

See every word, photo, drawing and table—including color images!

Print what you see.



Requires Microsoft *Windows*. 2000 and 1999 edition supports *Windows* and Macintosh systems.

ARRL Periodicals CD-ROM sets:

Year	Order No.
2000	#8209
1999	#7881
1998	#7377
1997	#6729
1996	#6109
1995	#5579

JUST $19.95* PER SET!

***Shipping & Handling** charges apply: US orders add $5 for one CD, plus $1 for each additional CD ($10 max). International orders add $2.00 to US rate ($12.00 max).
Sales Tax is required for orders shipped to CT, VA, CA, and Canada.

QEX and NCJ Collection CD-ROMs

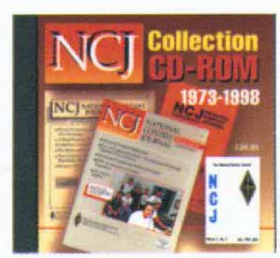
Back issues from these journals have been scanned to provide black-and-white images that can be read on your computer screen or printed. All the articles, ads, columns and covers are included. Search—Select—Browse—it's all there!

The QEX Collection CD-ROM.
For Communications Experimenters! Includes all *QEX* issues from 1981 through 1998.

The NCJ Collection CD-ROM.
Contests, scores and more! Includes all *NCJ* issues from 1973 through 1998.

System Requirements: Pentium or equivalent IBM-compatible PC, and Microsoft *Windows*.

QEX Collection CD-ROM
#7660 .... $39.95*



NCJ Collection CD-ROM
#7733 .... $39.95*

ARRL *The national association for* *AMATEUR RADIO*
www.arrl.org/shop

225 Main Street, Newington, CT 06111-1494 tel: 860-594-0355 fax: 860-594-0303

In the US call our toll-free number **1-888-277-5289** 8 AM-8 PM Eastern time Mon.-Fri.

Name: ham radio, born January, 1968.

Why **ham radio** (magazine)? *The electronics and communications industry is moving forward at a tremendous clip, and so is amateur radio. Single sideband has largely replaced a-m, transistors are taking the place of vacuum tubes, and integrated circuits are finding their way into the ham workshop. The problem today, as it has always been, is to keep the amateur well informed.*— Editor Jim Fisk, W1DTY (SK), from the preview issue of *ham radio* magazine, February, 1968 (last issue published in June, 1990).

hr focus on communications technology . . .

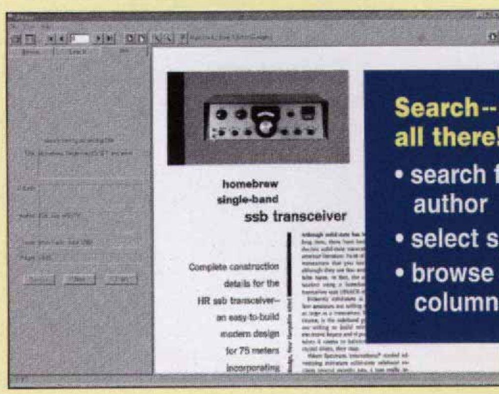
Introducing Ham Radio CD-ROMs!

System Requirements: Pentium or equivalent IBM-compatible PC, and Microsoft Windows™ 95, 98, NT 4.0, Me, or 2000.

Now you can enjoy quick and easy access to back issues of this popular magazine! These CD-ROM sets include high quality black-and-white scanned pages, easily read on your computer screen or printed. All the articles, ads, columns and covers are included.

Readers will enjoy a wealth of material that spanned the gamut of Amateur Radio technical interests: construction projects, theory, antennas, transmitters, receivers, amplifiers, HF through microwaves, test equipment, accessories, FM, SSB, CW, visual and digital modes.

The complete set covers more than 30,000 pages!



Search--Select--Browse—it's all there!

- search for articles by title and author
- select specific year and issue
- browse individual articles and columns

Only $59.95 per set:*   Each set includes four CDs!

Ham Radio CD-ROM 1968-1976 ARRL Order No. 8381
Ham Radio CD-ROM 1977-1983 ARRL Order No. 8403
Ham Radio CD-ROM 1984-1990 ARRL Order No. 8411

SAVE $80! when you order the complete set:*

All 3 Ham Radio CD-ROM Sets (1968-1990)
ARRL Order No. HRCD **$149.85**

***Shipping/handling fee:** US orders add $5 for one set, plus $1 for each additional set ($10 max, via UPS). International orders add $2.00 to these rates ($12.00 max, via surface delivery). Sales tax is required for orders shipped to CA, CT, VA, and Canada.

ARRL *The national association for AMATEUR RADIO*

225 Main Street, Newington, CT 06111-1494
tel: 860-594-0355   fax: 860-594-0303

In the US call our toll-free number **1-888-277-5289** 8 AM-8 PM Eastern time Mon.-Fri.
Quick order **www.arrl.org/shop**

Ham Radio CD-ROM, © 2001, American Radio Relay League, Inc. Ham Radio Magazine © 1968-1990, CQ Communications, Inc.