$5

QEX

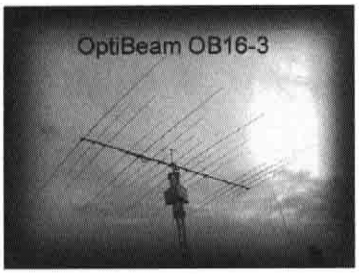
INCLUDING: COMMUNICATIONS QUARTERLY

Forum for Communications Experimenters

January/February 2003

XQ2FOD's elegant autotuning mobile antenna—inside!

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



OPTIBEAM

High Quality, High Performance, Multi-Band Yagis

Quality Made in Germany!

The OptiBeam Product Line:

Model	Elements	Bands	Boom Length
OB6-3M (Moxon)	6	20-15-10	10 feet
OB7-3	7	20-15-10	14 feet
OB11-3	11	20-15-10	20 feet
OB16-3	16	20-15-10	33 feet
OB9-5	9	20-17-15-12-10	17 feet
OB4-2W	4	17-12	12 feet
OB7-2W	7	17-12	17 feet
OB9-2W	9	17-12	33 feet

RF Applications, Inc

VFD Power and SWR Meters

- Vacuum Fluorescent Display
- 2,955 and 5,000 Watt HF Version
- 300 Watt VHF Version
- 65 Element Bar Graph— Better than a Meter!
- Settable VSWR Alarm
- VSWR Alarm Relay Option
- RF Power Monitor Option
- Vanity Option (Callsign)
- Remote Sensor
- 12 VDC Operation
- Accurate Peak/Hold Readings

Also P3000, P5000 and WINWATT

The VFD's display holds your peak PEP for about two seconds after you stop transmitting.

RF switching, phasing and contesting products from

ARRAY SOLUTIONS

RATPAK & SixPak—6 way 5 kW antenna switches

Filters—Bandpass and BCB by W3NQN

SCK—CW and Phone message keyer, and a lot more

SO2R Master—Finally a SO2R controller for high performance contesting

StackMaster—stack/phase up to 4 mono band antennas

StackMatch—the world's leader in stacking/phasing devices for mono or Tribanders

Phasing Systems—for 2 and 3 element and 4 square vertical arrays

AS80FS 80-Meter Vertical

- Full Sized 1/4 λ on 80
- Free Standing
- Rigid Base
- Wind Rated for 100+ MPH
- Handles 10 kW+
- Winch Up/Down (with optional removable winch)
- Radial Ring Included
- 160 and 40 Meters with Optional Tuner

WXØB ARRAY SOLUTIONS

Antennas and Antenna Accessories by: Cal-Av, Titanex, Bencher, M2, low band specialist AY Technologies and RFI and surge protection devices by I.C.E.

NEC-Win—The most powerful and leading antenna modeling programs on the market.

nittany scientific inc NSI

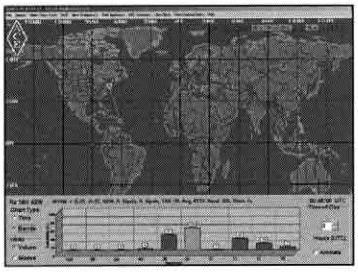
NEC-Win Plus - Developed for beginners, hobbyists, and field engineers. Includes almost unlimited wire segments in the basic program, polar plots, rectangular plots, input impedance and VSWR, tabular data, Synthesis Light and Necvu 3D, and the NEC2 calculating engine.

NEC-Win Pro - Developed for the researcher and professional engineer, it Includes polar plots, Smith Chart, tabular data, and 9 rectangular plots. Includes and supports the full NEC2 command set.

GNEC - Supports NEC4. (A NEC4 license is required.)

Basic Antenna Modeling Tutorial by LB Cebik - A hands-on tutorial. A fantastic book which includes a disk of antenna examples for the lessons. Intended to be used with NEC-Win Plus, but it is sufficient for learning modeling in general. Learn to be an antenna modeling expert with LB Cebik's most informative teaching style.

ACE-HF



Professional Propagation Software for HF Radio Operators

- Easy-to-use propagation software powered by VOACAP.
- Best frequency, circuit summary and MUF charts.
- Animated area coverage maps based on time-of-day, frequency or sunspot number.
- Animated circuit quality graphs of S/N ratio, S-units, reliability, required power gain and elevation angle.
- Modify any circuit parameter in seconds.
- ACE-HF PRO version adds a database of more than 35,000 receive locations, a DXCC list and speed select. Military/commercial versions are also available.

PRO.SIS.TEL.

BIG BOY ROTATORS

The most powerful antenna rotators available anywhere.
Amateur, Military, Commercial, Industrial

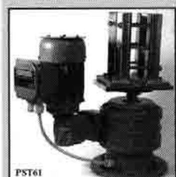
Two controllers to choose from:

Controller "B"

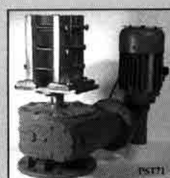
- Paddle Key
- Preset with True 360° Rotary Encoder
- Soft Stop
- Large Green Display

Controller "C"

- Keypad Commands
- Nine R/W Memories
- Voice Confirmation
- Soft Start/Stop
- RS-232 Built In



ProSisTel Rotators were designed to perform under tremendous stress with abnormally large antenna loads - up to 81 sq. ft. Perfect for large 80 Meter beams, long-boom Yagis and log periodics, stacked arrays, and rotating towers. Several AZ and EL models to choose from.



ProSisTel Rotators give you incredible starting and rotating torque with tremendous braking resistance. They use double-worm technology, far exceeding any other amateur rotator on the market. And we back them up with a **Two-Year Warranty (USA).**

WXØB ARRAY SOLUTIONS

www.arraysolutions.com

Phone 972-203-2008 sales@arraysolutions.com FAX 972-203-8811

Contesting products for the dedicated Contesters, DXing products for the deserving

WXØB ARRAY SOLUTIONS

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 $24; nonmembers $36. 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 216

Mark J. Wilson, K1RO
Publisher

Doug Smith, KF6DX
Editor

Robert Schetgen, KU7G
Managing Editor

Lori Weinberg, KB1EIB
Assistant Editor

Zack Lau, W1VT
Ray Mack, WD5IFS
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:

Joe Bottiglieri, AA1GW, *Account Manager*
860-594-0329 direct
860-594-0200 ARRL
860-594-4285 fax

Circulation Department

Debra Jahnke, *Circulation Manager*
Kathy Capodicasa, *Senior Fulfillment Supervisor*
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;

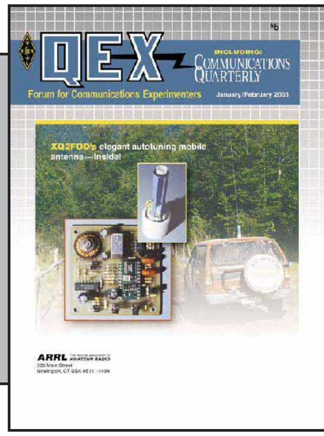
Canada by Airmail: ARRL member $40, nonmember $52;

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

Members are asked to include their membership control number or a label from their *QST* 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 ©2002 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

An alternative approach to screwdriver antennas, the story starts on p 3.

Features

Columns

Jan/Feb 2003 *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.

Empirical Outlook

QEX Forums On Line

A friend of *QEX*, Chris Karpinski, AA1VL, has created a new Web site at **www.neoamateur.org.** It is devoted to experimentation and technical issues related to Amateur Radio. He set up at least two on-line forums there for you, our readers. While not officially sanctioned by ARRL, the forums have our support for the following reasons.

Sixty days can be a long time to wait for corrections to published material and from time to time, we may put them there so that readers might get them quickly. Further, we feel that an on-line forum logically supports a mission of *QEX*: to facilitate the exchange of ideas and information among Amateur Radio experimenters. Finally, it provides a place for feedback, opinions and other items for those who prefer the speed and informality of on-line publication.

We understand that you sometimes have questions about authors' assertions or about sources of parts and information. Our policy is to correct all errors and allow authors a chance to defend their claims. We encourage criticism of published material where productive, but we view negative personal remarks with disdain. Please consider that when you post things to the forums. Chris welcomes suggestions for new groups under the *QEX* heading that pertain to particular projects or topics.

Regrettably, Peter Bertini, K1ZJH, is unable to continue as Contributing Editor. We plan to keep Tech Notes, though, and we welcome your submissions for it. The column is a neat vehicle for articles of 3000 words or less, summarizing new techniques, ideas or materials. Keep those projects going!

On November 22, 2002, Didier Chulot, F5MJN, operating F8KGG in Paris, France, and yours truly in eastern Tennessee made a two-way contact using the digital voice system described in this issue. Believed to be the first transatlantic contact of its kind, the feat took place on 21,218 kHz using about 400 W EIRP at each end. Signals were of readability 5 and strength S-5 to S-7. QPSK and 16-QAM modes were used, achieving a peak symbol rate of 2400 bauds. I recorded a mean opinion score of 3.5 using the 1200 bit/sec QPSK mode. Further tests are ongoing at the time of this writing.

In This Issue

Software radio, digital audio and other advanced digital techniques continue to shine. That does not mean *QEX* is going entirely digital; it is just that digital techniques constitute an especially fruitful area for experimentation these days. We still have a lot to print about analog electronics.

Interestingly, Manfred Mornhinweg, XQ2FOD, wrote about his 13.8-V, 40-A switching power supply for *The ARRL Handbook*. Here, he takes us inside the design and construction of his multiband mobile HF antenna. Learn how he overcame the battle against low efficiency, while operating his automatically tuned unit over more than two octaves. The article features many interesting mechanical ideas and plenty of illustrations.

Ulrich Rohde, KA2WEU, focuses on pragmatic issues of receiver IMD measurement in the concluding segment of his series. He discusses why IMD does not always behave as predicted and presents information about an interlaced dual AGC system.

We have been declaiming that switch-mode power supplies deserve more coverage. Contributing Editor Ray Mack, WD5IFS, is addressing that as he brings us Part 2 of his series on switchers. This segment contains something we have been wanting: a discussion of magnetic design in power conversion. Ray shows how to select materials and how to build inductors and transformers that work. Follow the rules and get a healthy, nonsmoking circuit that lasts a long time!

Leif Åsbrink, SM5BSZ, returns with a second installment on *Linrad*. We've had the introduction; now we get to the nitty gritty. Software unites with some relatively simple front-end hardware in this segment.

Cédric Demeure and Pierre-André Laurent describe the new digital audio-broadcasting standard called DRM, for Digital Radio Mondiale. Their Amateur Radio system is a subset of the standard, operating in a 3-kHz bandwidth. It is capable of simultaneous digital voice and data at over 3000 bits/second on HF. The authors discuss practical advantages along with some of the technical details.

In RF, Contributing Editor Zack Lau, W1VT, describes a 6-meter Yagi antenna.—Happy New Year!—*Doug Smith, KF6DX*; **kf6dx@arrl.org**.

An Automatically Tuned 7-30 MHz Mobile Antenna

Here's a beautiful, weatherproof alternative to screwdriver antennas for the mechanically adept.

By Manfred Mornhinweg, XQ2FOD

For any ham who likes to travel, and who often travels by car, a mobile installation soon becomes a necessity. This is especially true when traveling on roads like the one depicted in Fig 1! While in inhabited areas there is usually some VHF repeater coverage, in many other places VHF radios are useless. HF can provide reliable contacts, and lots of fun, from anywhere. That led me into setting up a combined VHF-HF mobile station as soon as I got my first car.

The VHF antenna is easy enough: An NMO mount on the center of the roof takes either a $\lambda/4$ whip for modest operation in cities, or a $^5/_8$-λ element for better performance when out on the highways. The HF portion involves a lot more difficulties. Many commercial mobile HF antennas are less than practical because of their size, weight and stiffness; others have low performance. So, I preferred to homebrew antennas from the start.

Over the years—and when changing cars—I have gone through four iterations in my aim for a "perfect" antenna. While of course nothing can be really perfect, the antenna presented here is the result of 10 years of experience in both building and using HF mobile antennas, and it has me completely happy!

Any ham who has ever operated HF mobile knows that mobile antennas, especially those for the lower bands, tend to be notoriously narrow-banded. In addition, most mobile antennas require replacing resonators, or at least plugging jumpers, to change bands. I used to solve the problem of insufficient bandwidth by using an automatic antenna tuner, but this did not solve the band-changing problem. Much too often, it happened that someone invited me over to a different band or the propagation closed down on the current band. Not wanting to pull over, stop, get out into the rain and change a jumper, I simply switched off the radio.

That happens no more. The antenna presented here covers all ham

Fig 1—The antenna has survived well during trips through rugged country.

Buenaventura Osorio 720 Dpto. 1
La Serena, Chile
mmornhin@gmx.net

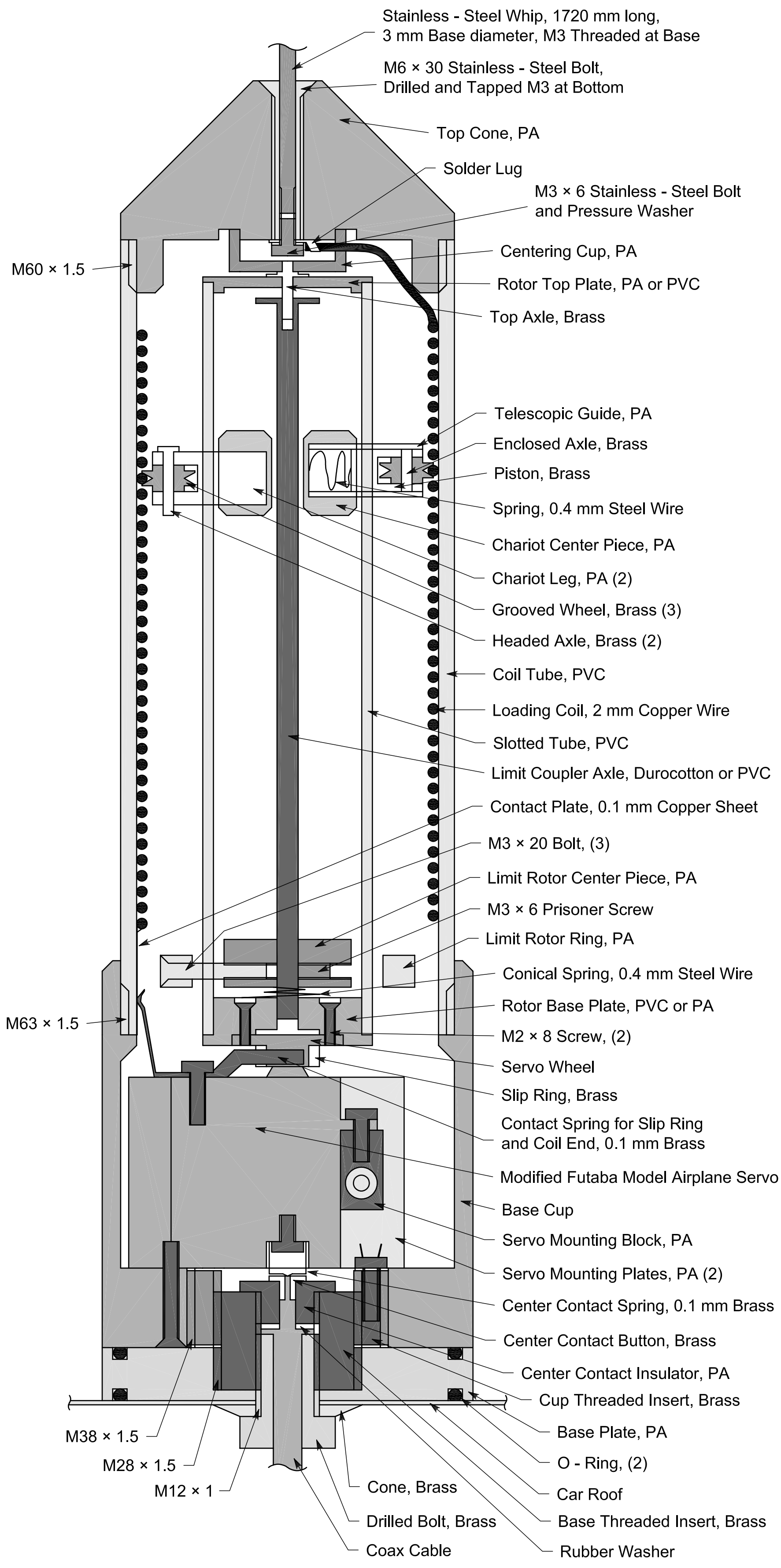


Fig 2—Mechanical details of the matching network.

bands from 40-10 meters with a maximum SWR of 1.3:1 and good efficiency. Since remotely adjusting an antenna while driving could be dangerous, I included fully automatic tuning, commanded from the transceiver. The antenna is very lightweight, tall but flexible and looks simple while actually being quite sophisticated. The project involved a major amount of work, but it gave me many hours of entertainment.

The Basic Concept

This base-loaded vertical antenna mounts on the car's roof. The loading coil is designed as a variable inductor, with a three-legged chariot that travels up and down inside the coil. Grooved brass wheels run on the coil turns, driven by a slotted rotor tube. A modified model-airplane servo (inexpensive and widely available) is used as a compact and powerful motor. Two limit switches shut off the motor current before the chariot can run off the ends of the coil. The motor is fed by positive and negative dc voltages applied through the coax cable. The bulk of the antenna can be quickly removed from the car without any tools, while only the mounting base is left in place, protected by a cover. The entire antenna is sealed against the weather by silicone caulk in the permanently assembled threads and O-rings at the other places. Only plastic and stainless steel are exposed to the weather. An impedance-matching coil is included, resulting in very low SWR. To maximize efficiency, the loading coil is large and—to the best extent possible—metal parts are avoided in close proximity to the coil. All mechanisms and coils are inside the assembly, resulting in a smooth, simple appearance.

The control unit can be mounted anywhere between the antenna and the transceiver. It need not be mounted within the reach of the operator, since it is remotely operated from the transceiver. It is connected via one coax cable to the antenna and one coax cable plus one power/data cable to the transceiver. It contains a Basic Stamp microcontroller, an SWR sensor, a dc-dc converter with bipolar output for powering the motor and associated components.

The microcontroller communicates with the Kenwood TS-50 transceiver, emulating an AT-50 antenna tuner as seen from the transceiver side. If a different transceiver were used, a relatively simple software modification would be necessary. Commanding the system to tune to any frequency within the covered range is done in exactly the same way as one would do it with one of the factory-made automatic antenna tuners: Go to a nearby clear frequency and press **AT TUNE**!

Building this antenna requires the use of a lathe, but it doesn't need to be a particularly large or complex lathe. I built my antenna using a simple, relatively inexpensive desktop hobby lathe, and the antenna was the first major mechanical project I attempted, which shows that no unusual mechanical skills are necessary to successfully build it! On the other hand, this definitely isn't a weekend project.

Mechanical Design

Please refer to Fig 2 to be able to follow. The 172-cm stainless steel whip, which has a base diameter of 3 mm and tapers out to a slim 1.5 mm, screws into a drilled and tapped stainless-steel bolt. The bolt, in turn, is permanently screwed into the top cone, made from polyamide (PA, also known under the tradenames of Nylon and Technyl, among others). This top cone screws into the coil tube, which has the loading coil wound on its inside, with a pitch of 3 mm. The coil tube is simply a piece of polyvinylchloride (PVC) water pipe of 63-mm outer diameter and 3-mm wall thickness. In countries that use inch-size water pipes, there are almost exactly equivalent pipes available of $2^{1}/_{2}$-inches outer diameter and $^{1}/_{8}$-inch wall thickness. Winding the coil on the inside of the pipe is one of the more challenging parts of building this antenna, but I give the detailed instructions later in the article.

The chariot travels on the coil by means of three grooved wheels that run on the wire. Two of these wheels are fixed on their respective legs. The third has a spring-loaded telescoping mount, which compensates for the unavoidable imprecision in the dimensions. The chariot is supported only by the three wheels floating freely inside the rotor tube. The tube has three lengthwise slots through which the chariot legs poke. Just like the coil tube, the slotted tube is also made from standard PVC water pipe. This pipe has a 32-mm outer diameter and 2-mm wall thickness, being equivalent to $1^{1}/_{4}$-inch pipe.

The top of the rotor tube is terminated by a PVC plate that has a brass axle inserted, which engages in a cup-shaped piece. This centers the rotor assembly by engaging a circular groove in the top cone. The lower end of the rotor tube is terminated in a PVC base plate, which is in turn fixed to the output wheel that comes with the model aircraft servo, a Futaba FP-S148. The servo provides the motion. As the servo turns the rotor, the

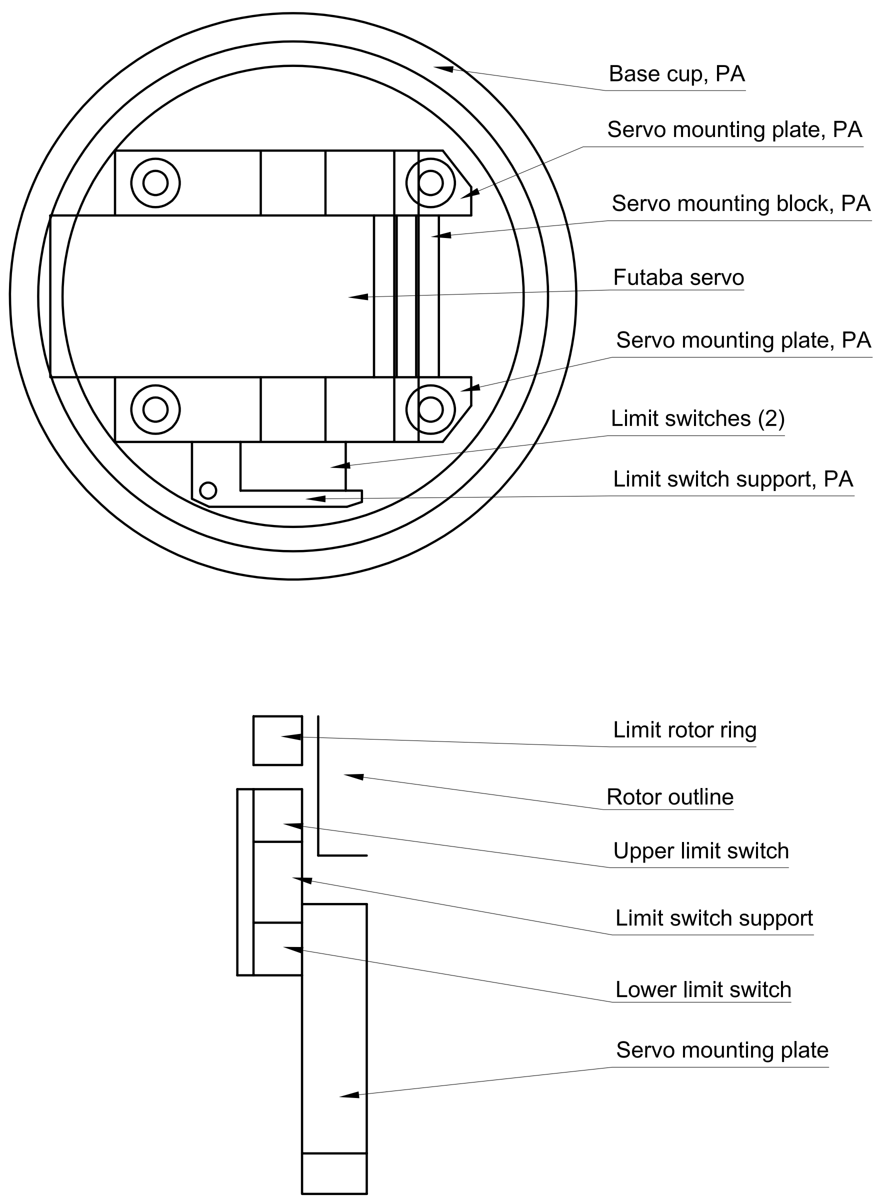


Fig 3—Base-cup and servo details. The upper half shows a cross sectional view through the base cup, from which the dimensions and positions of the servo mounting parts can be extracted. The lower half shows the exact location of the limit switches relative to the other parts.

Fig 4—The coil assembly must rotate as the glue dries. I used the winding machine at low speed, and directed the beam of an infrared lamp on the coil tube, in order to speed up the setting of the glue.

edges of the three slots in the rotor tube push against the chariot legs, making it turn. The chariot then is guided up and down by the coil turns. The lowermost of the grooved wheels doubles as moving tap on the coil.

When the chariot reaches the upper end of the allowable range, the chariot centerpiece will lift the limit coupler axle, which has a longitudinal play of 2 mm both up and down from the resting position. The axle will carry along the limit rotor centerpiece and with it will lift the limit rotor ring, which will activate the lower limit switch and stop the servomotor. Likewise, when the chariot reaches the lower end of the range, it will push down the limit rotor so that the ring will activate the other limit switch. To avoid confusion, the limit switches are not shown in the drawing; instead, a detail drawing is provided in Fig 3. Use that one and the photos to visualize this part.

Also not shown in the drawing is a flexible wire that connects the axle of the lowermost chariot wheel to the slip ring, passing through a slanted hole in the rotor base plate. A contact spring slides on the slip ring, providing the connection to the movable tap on the coil. The same contact spring connects to a contact plate soldered to the lower end of the coil. Thus, the unused part of the coil is shorted out. From this contact spring, a toroidal impedance matching coil is connected to ground and a coupling capacitor is connected from the spring to the center contact spring, which has direct contact to the coax feed center conductor on the base.

The servo, limit switches, all associated components, impedance matching coil, coupling capacitor and the contact springs form a compact unit that mounts with four bolts inside the base cup, made from PA like most of the parts. The coil tube screws into the base cup, closing the assembly. The base cup carries a brass threaded insert, which engages to a similar but smaller threaded insert in the mounting base. This brass-to-brass thread is

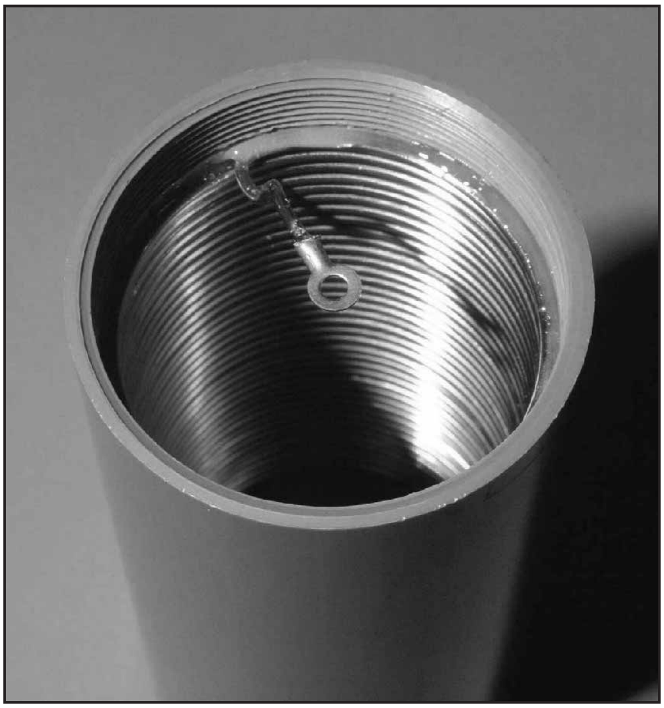


Fig 5—The top end of the loading coil after the glue has dried. Notice extra epoxy that strengthens the last turn.

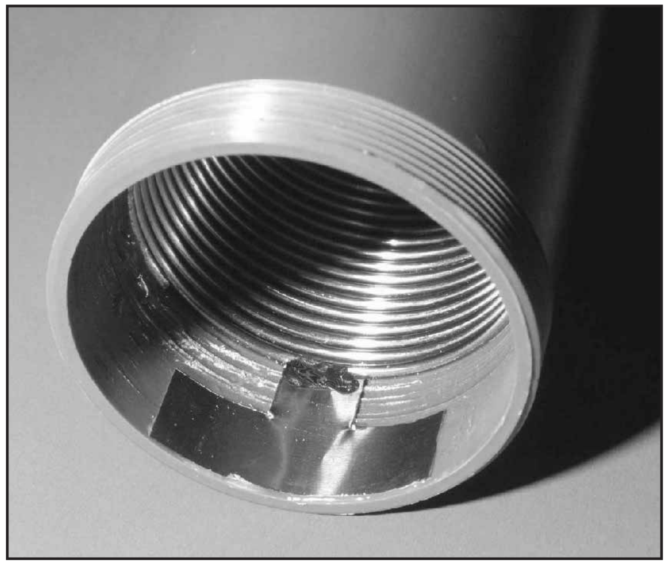


Fig 6—The lower end of the coil is cut to the dimensions given in the plan, and then reinforced with a dab of epoxy. A contact plate made from 0.1-mm copper sheet is soldered to the last turn of the coil and epoxied to the tube.

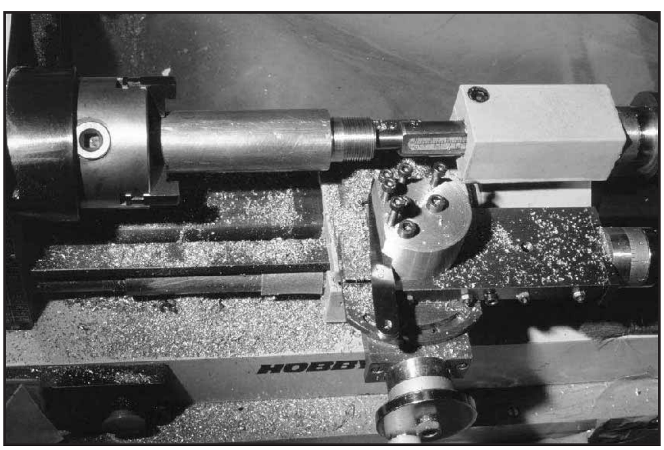


Fig 7—A piece of brass stock in the lathe, just after cutting the external thread of the insert that goes in the mounting base.

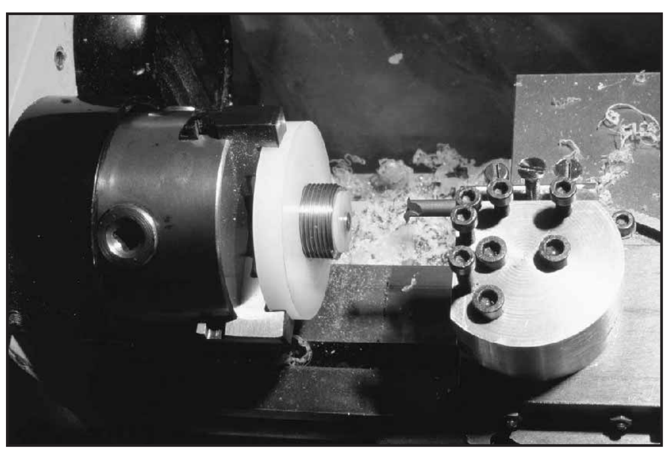


Fig 8—The entire base was assembled while still mounted in the lathe.

the one that is unscrewed to remove the main part of the antenna from the car roof. The brass-to-brass contact is electrically safe and mechanically wear-resistant.

The base cup threaded insert is locked in place by a single M3 bolt, which also carries a solder lug for the ground connection of the antenna innards. The other threaded insert screws into the base plate and is fixed by a brass M12 bolt through a 12-mm hole from below the car roof. This bolt has a center hole through which the RG-58 coax cable enters the base. A PA insulator, which is also M12-threaded, presses the coax shield against the drilled bolt, using a rubber washer or small O-ring to assure good contact. The dielectric of the coax cable enters the hole of the center insulator, and the center conductor is soldered to a small brass center contact button, on which the contact spring of the main antenna unit rests.

The base plate is sealed against the car roof using an O-ring. A 2-mm plastic layer acts as a cushion between the brass threaded insert and the car roof to avoid damaging the paint. When reselling the car, the hole can be closed with a little cap making it almost invisible. The base plate has another O-ring embedded that seals it against the main antenna unit.

The all-important ground connection to the car roof is made by a brass cone that encloses a toothed washer, which bites into the car roof from below. This connection of brass to steel is vulnerable to galvanic corrosion, so it is very important to apply a coating of thermally stable grease (I used silicone grease) before assembly. This grease will avoid corrosion and contact problems.

You will probably want to print the drawing. The full-resolution file is available on the *ARRLWeb* in GIF format.[1] It is dimensioned so that if you print it at 600 dpi, you should end up with a true-sized print. It will fit either on a DIN A4 sheet or on a US legal-size one, but not on a letter-size one. Be aware that some viewers, and especially some browsers, may have trouble handling a file of this resolution (27 megapixels), so you may have to get a better viewer. *IrfanView*, available from **www.download.com**, is highly recommended for such purposes.

The best way to get the exact dimensions is by downloading the original AutoCAD R14 DWG file. You will need AutoCAD or some compatible software to open this file, but then you can get all dimensions, easily and in full precision. Otherwise, you can measure them on the print. It helps to know that most dimensions are exact multiples of one millimeter.

Fig 3 is a detail drawing. It is also available on *ARRLWeb*, as is the AutoCAD drawing file. The upper half shows a cross-sectional view through the base cup, from which the dimensions and positions of the servo mounting parts can be extracted. The lower half shows the exact location of the limit switches relative to the other parts. The lower switch is mounted by two M2 bolts going through the limit-switch support and the switch, self-tapped into the servo mounting plate, while the upper switch is held in place by two bolts self-tapped into the support. The location of the bolts depends on the exact microswitches you find, so I didn't draw them.

The lathe I used is a small German-made desktop unit intended for hobby use that can turn pieces up to 75-mm diameter and about 250-mm length. I bought this lathe with the specific idea of finally being able to build a good mobile antenna and other radio-related mechanics. I learned the basics of tooling operations while building this antenna.

The Loading Coil

There are several advantages to winding the loading coil inside the support tube. It will be protected from the weather and dirt; it will look clean, and it will allow all of the tuning mechanisms to use the room inside while contacting the wire. Therefore, I had to come up with a way of making such a coil! It needs to be quite precise, because the wheels of the chariot must be able to freely run on the coil turns, without ever causing a short between two turns.

[1]You can download some of the photos, drawings and source code for the BS2 from *ARRLWeb* at **http://www.arrl.org/qexfiles/**. Look for 0103Morn.zip.

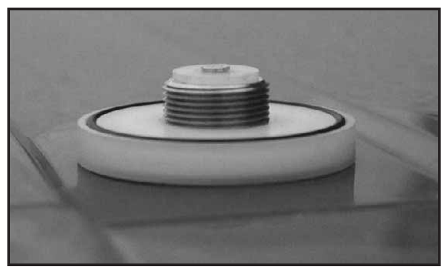


Fig 9—The base plate mounted on the car.

Fig 10—When the antenna is dismounted, a simple cap screwed onto the base protects the contacts from weather.

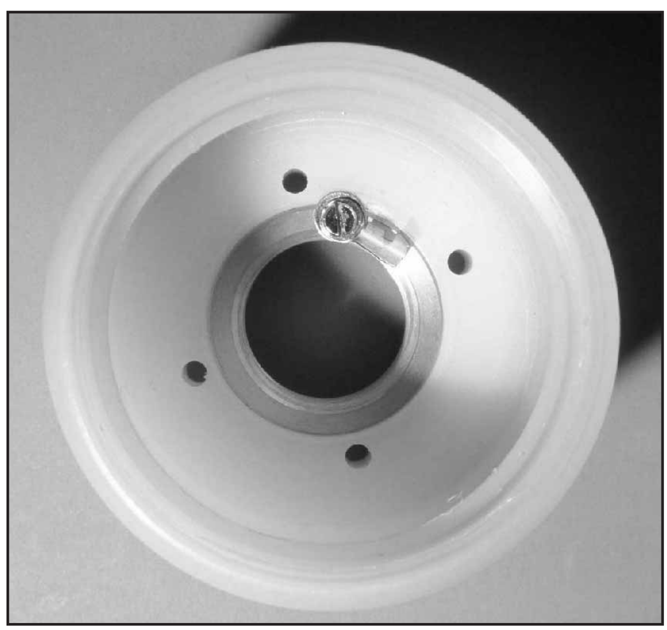


Fig 11—(right) The base cup is made from solid PA. This requires removing much material from the inside.

I first made a winding core from inexpensive, soft polyethylene (PE). This core has a diameter of 55 mm, which is 2 mm less than the internal diameter of the coil tube. The core is slightly longer than the coil tube; the exact length is not critical. Into this core, I cut a rectangular thread with a pitch of 3 mm per turn. The width of the cut is 0.1 mm more than the wire diameter, so that the wire can move freely in the thread, and the depth is equivalent to the wire diameter.

Then I wound the coil on this core. I used a winding machine I got from CE5FSB, back in my university times, but hand winding could also have been used. The wire is the kind used for electrical installations in homes. It's solid copper wire of 2.5-mm^2 cross sectional area, insulated in PVC. I removed the insulation while winding by slitting it open with a sharp knife. It's good to avoid touching the bare wire, because the skin oils would later weaken the bond with the glue used to mount the coil in the tube. The wire ends are fixed to the core using two screws installed in the two ends of the core.

After machining the coil tube, making its two threads, the coil must be installed inside. This is done with slow setting epoxy glue (I used 90-minute Araldite), in this way: Firstly, the tube is sanded on the inside using coarse sandpaper to improve adhesion of the epoxy glue. A layer of epoxy is applied to the inside of the tube. This layer should be 0.2-mm thick. If it's thinner, the structure will end up weaker; and if it's much thicker, the epoxy will hinder the free motion of the chariot wheels. I made a spatula of plastic scraps, which has the blade cut to the exact radius of the glue layer (28.3-mm), and covers about 90° of the tube. This spatula has two little holes at the corners, in which I anchored small pieces of 0.2-mm magnet wire. These serve as spacers between the spatula and the tube, making it very easy to apply the required even layer of glue.

Then four pieces of 1-mm magnet wire were inserted into the tube at 90° from each other, and bent over the ends of the tube. These four wires serve as spacers, allowing you to slide the core with the coil into the tube without touching the glue! Now, from each end of the assembly, four small pieces of spacing wire are inserted. These wires keep the core and coil centered while the long spacing wires are pulled out.

The anchors of the coil wire at the ends of the core are now removed, and the wire is cut off cleanly, so that it can retreat in the thread of the core. If all is correct, the coil will spring open and contact the glue, while being held at the precise pitch by the edges of the core's rectangular thread! The low friction of PE helps making this step easy. In any case, you can help by pushing the coil wire ends in. It's important that the coil springs open to the full inner diameter of the tube, because otherwise the chariot may jam later.

During this procedure of settling the coil against the tube, the short spacer wires are kept in position by adhesive tape or other means. It's very important to keep the core well centered at all times, because otherwise it may bind to the glue. While PE does not adhere well to epoxy glue, you may still manage to get enough glue on it to cause trouble, so be careful!

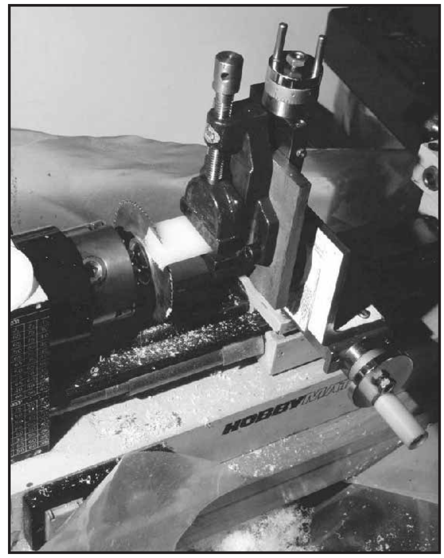
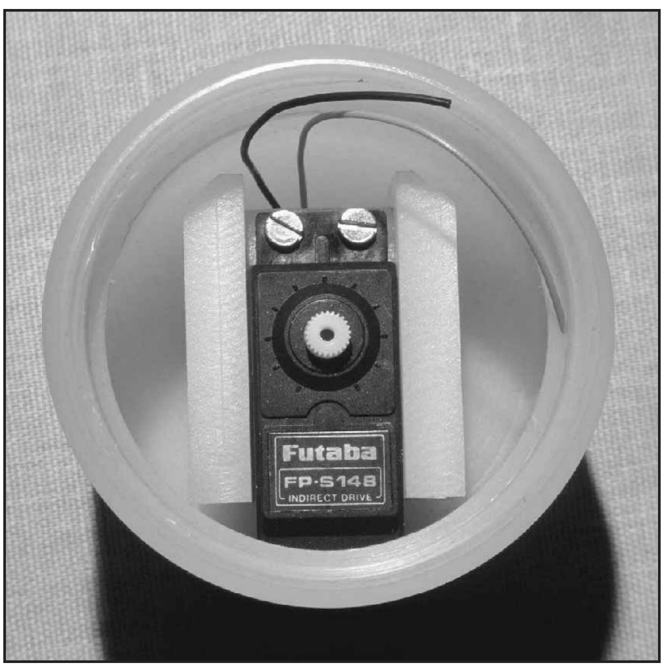


Fig 12—The hobby lathe cuts a servo mounting plate to rough size.



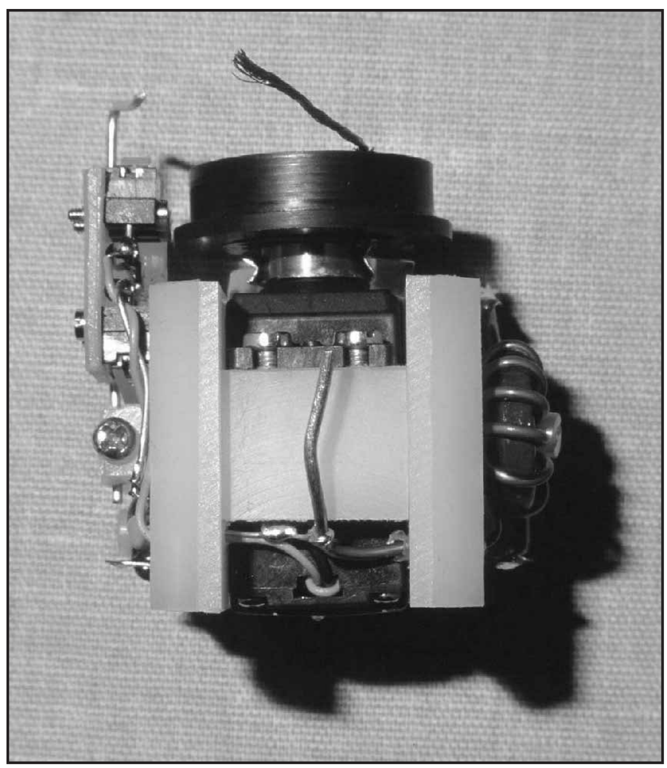


Fig 13—The servo with its mounting plates test-fitted in the base cup. Notice that one edge of each plate is chamfered.

Fig 14—(Right) The servo assembly seen from the front (grounding) side.

After the coil has fully expanded, the glue should be allowed to set while the assembly is constantly rotating. This is necessary to avoid the risk of having the glue flow down by gravity, causing an uneven layer and later trouble. I did this with the whole assembly in the coil winder. As shown in the photo (Fig 4), I used the winding machine at low speed, and directed the beam of an infrared lamp on the coil tube to speed up the setting of the glue. Excuse me for the messy photo.

When the glue has fully set, it's time to remove the core. This is the moment when it's most important that the core was made from slippery PE, and that the thread was wide enough–because the core must be literally screwed out of the coil. I clamped the core end in a vise and turned the coil tube with both hands.

After the core has come out, you can inspect your coil. Wow! Mine came out looking much better than I had expected. Fig 5 shows the top end of the loading coil. Any excess wire is removed, then the last quarter turn of wire is reinforced with additional epoxy to cover the wire (no wheel will

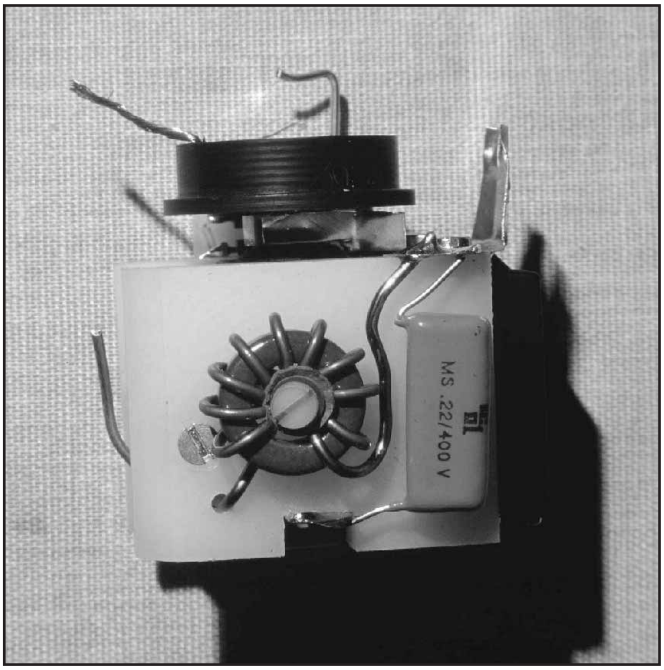


Fig 15—A side view shows the RF portion of the circuit.

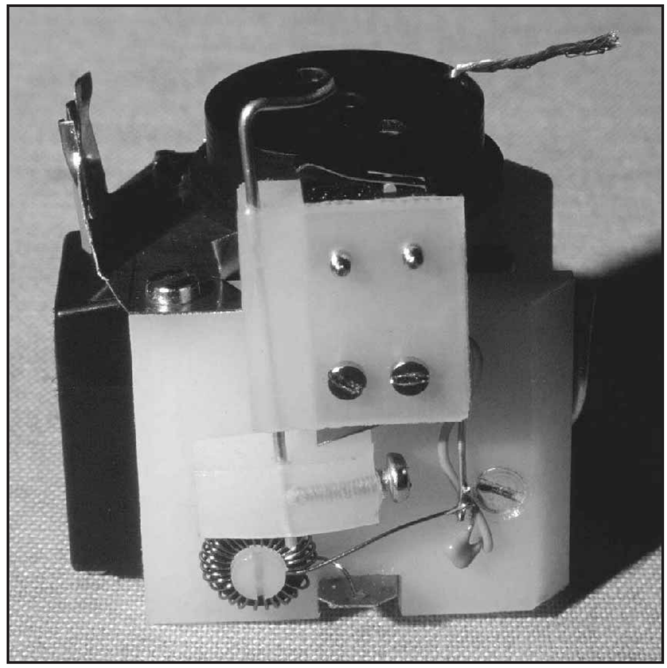


Fig 16—The limit switch side can be seen here. L3 and C4 are visible. The most interesting feature is the steel wire and associated parts, used to link the lower limit switch to the top surface of the limit rotor ring.

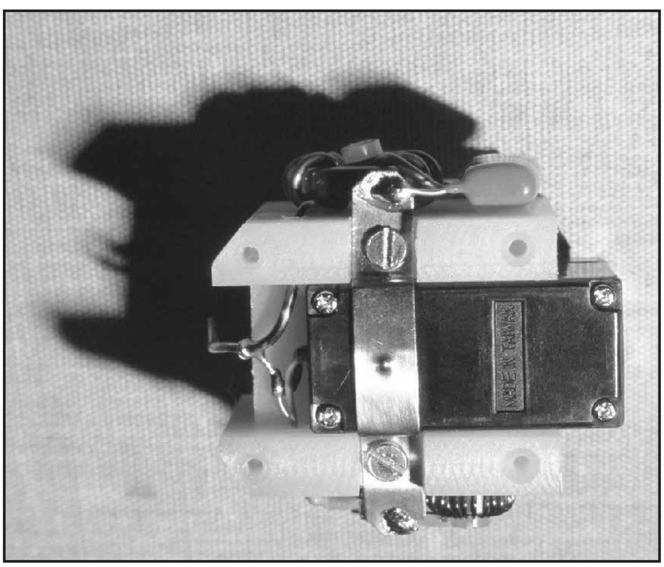


Fig 17—A bottom view of the servo assembly (see text).

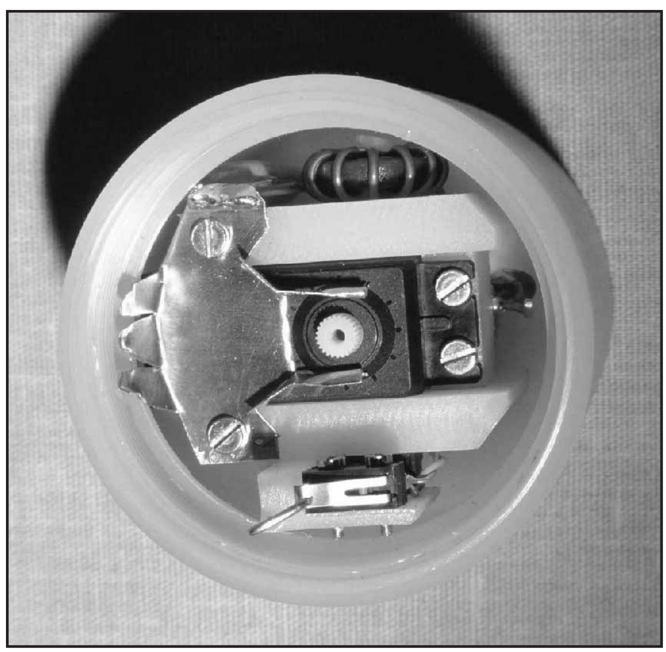


Fig 18—(Right) The servo assembly mounted inside the base cup. Notice how everything fits precisely! There is relatively little room wasted.

travel here). The wire end is bent to fit the plan, and a solder lug is installed. The lug must be centered and flush with the end of the tube.

Likewise, the lower end of the coil is cut to the dimension given in the plan, and then reinforced with a dab of glue. A contact plate made from 0.1-mm copper sheet is soldered to the last turn of the coil and epoxied to the tube (Fig 6). If part of a turn remains unused, that's no problem; but make sure that the coil has the full dimension (amount of turns) shown in the plan. Even one missing turn would cut off a large part of the 40-meter band.

The Mounting Base

I have heard experienced mechanics say that it is difficult to make threads with a lathe. Expecting trouble, I was soon surprised that even my very first thread was perfectly usable! Fig 7 shows a piece of brass stock in my lathe, just after finishing the external thread of the insert that goes in the mounting base.

In Fig 8, the entire base has been assembled, while still mounted in the lathe, to test the fit of the plastic threads. Cutting precise threads in plastic is a little more difficult than in brass, because the soft material tends to flex away from the tool, so it's necessary to cut a little deeper to compensate. Frequent test fitting during the process prevents cutting too deeply. The tool used to cut inner threads can be seen here. It needs to be very sharp for proper finish quality in plastic.

Fig 9 shows how the completed base looks, when mounted on the car roof. Sorry, I did not wash the car before installing the antenna! Notice how the O-ring extends slightly above the surface of the base plate.

When the antenna is dismounted, a simple cap is screwed onto the base, to protect the contacts from dirt, water and corrosion. Fig 10 shows how the car roof looks with the antenna dismounted. Only the quarter-wave whip for VHF remains in place at all times, since it fits easily into most garages. The other wires in this photo are power and telephone lines in the background.

The Base Cup

The base cup (Fig 11) is made from solid PA, and it requires removing a lot of material from the inside. Those who have access to a large lathe and large drill bits will be able to do this very quickly; but for me, it meant removing almost all the material in successive passes of small tools.

A threaded insert, larger than that of the base but otherwise similar, is screwed into the cup. It takes an M3 bolt that serves a double purpose. It holds a solder lug providing a ground connection to the antenna innards, while also locking the threaded insert in place, so it won't unscrew from the cup when installing or removing the antenna from its base.

Note that the four holes used for the bolts that hold the servo mounting plates are *not* in a square arrangement, but a rectangular one. These holes can be drilled using a drill stand or using the lathe configured as milling machine. They must be drilled in the proper locations, so that the contact spring for the coil end will properly rest on the coil contact plate. Likewise, the threaded hole in the brass insert must be drilled in the proper location, so that the solder lug will be roughly centered under the end of the servo when the insert is correctly screwed into the cup.

The Inner Assembly

A hobby lathe is a very versatile tool! Fig 12 shows it configured for sawing a servo mounting plate to the rough size. Later, the circular saw was replaced by a cylindrical router bit, and then the precise finishing was performed. Note that setting up a lathe in this way allows full milling functionality.

PA plastic generates lots of heat by friction against a tool. It will easily melt, thus losing precision and finish quality. To avoid this, it's best to cool the tool with some liquid while working, but in my wooden-floored apartment, this isn't practical. The solution is to work with very sharp tools at low speed. I did not buy plastic sheet stock for this project. Instead, I turned my own plates from the same round 75-mm PA stock used to make the base plate, base cup and top cone.

In Fig 13, the two servo mounting plates and the mounting block have been joined to the servo and test-fitted in the base cup. Note that one edge of each plate is chamfered. This is necessary to be able to introduce the assembly into the cup, because the coil tube seats against a rim that has a narrower internal diameter than the rest of the cup. The rear end of the servo is tucked under this rim, so that its output shaft is precisely centered in the cup.

The servo is modified in three ways:

1. The white output shaft internally carries a gear and a cam. This cam originally is used to keep the shaft from rotating over more than three-quarters of a turn. This cam must be removed to allow continuous rotation.

2. The rear of the servo originally carries a mounting flange equal to that at the front (top in Fig 13) with two screw holes. The rear-mounting flange simply would not fit inside the cup, so it is removed.

3. The entire servo electronics, including the feedback potentiometer, are removed. Instead, the decoupling

Fig 19—The lathe milling slots in the PVC tube. The quality is rough, but adequate.

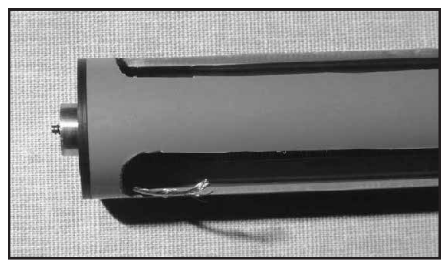


Fig 20—The bottom end of the rotor. The base plate is held in place by friction.

Fig 21—The assembly of Fig 20 as seen from the front.

components C2, C3 and FB1 are installed inside.

Fig 14 shows the servo assembly seen from the front (grounding) side. The servo has been fitted with its wheel, the rotor base plate and slip ring. Also, the contact springs, the entire limit switch and impedance matching assemblies have been installed. This is the complete assembly that goes into the base cup. The limit switches are on the left side. They have the diodes D1 and D2 installed between their pins. The top switch will later directly interface to the bottom side of the limit rotor ring, while the bottom switch employs a wire lever to engage to the limit rotor ring's top side.

On the right side is the toroidal inductor, L2, and the coupling capacitor, C1. The ground connections of both sides are brought to the front via small holes in the two servo mounting plates and joined there. The large wire pointing upwards is the ground connection. It will later be bent down and soldered to the solder lug attached to the cup threaded insert, as the last step of building this antenna. The live connections of both sides are soldered to the two ends of the center contact spring. The slip ring contact spring can be seen contacting the slip ring on both sides, under the rotor base plate.

The side view (Fig 15) shows the RF portion of the circuit. RF energy from the center contact spring is routed through a large capacitor to the contact spring on top. That spring applies this RF to the base of the loading coil, to the movable tap via the slip ring and to the impedance matching coil that has its other end grounded. Note that in principle, it's not necessary to use a 400-V capacitor here but I did it for safety reasons. Should I ever strike power lines with this antenna, or even get a lightning hit, the combination of the toroidal inductor to ground and this capacitor in series should protect the radio to a very high degree.

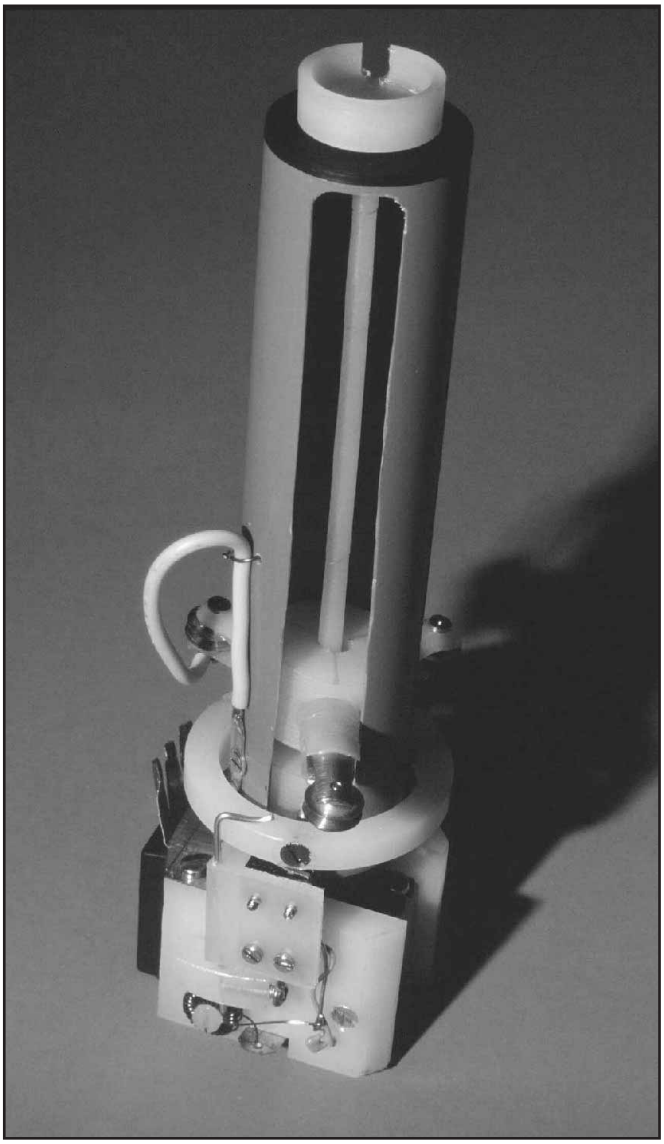


Fig 22—The complete innards of the antenna, from the centering cup on the top to the servo assembly! You can see how the limit rotor ring activates the two limit switches, and how the flexible wire connects the wire coming from the slip ring to the lowermost grooved wheel.

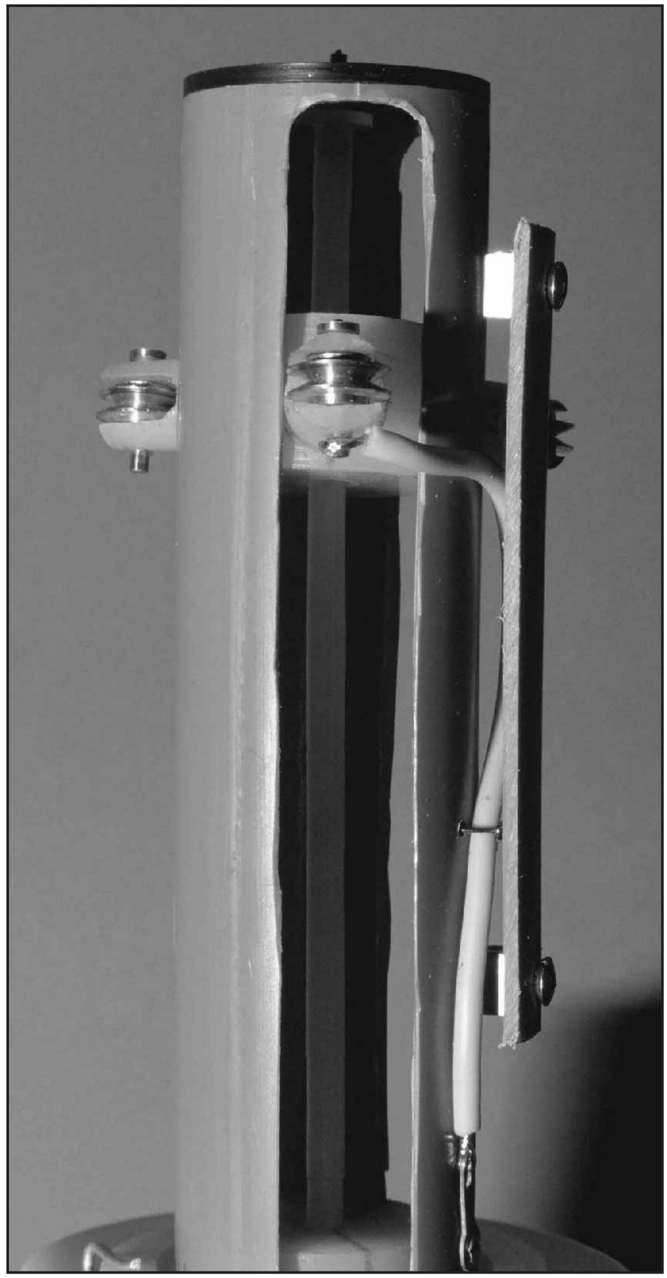


Fig 23—After a few tests, the flexible wire began rubbing against the coil, which slowed the tuning process and forced the mechanics. The solution is a simple guide for the wire, which I made from a segment of PVC tube. It is mounted with two M3 screws self-threaded into the slotted tube, and some junk-box spacers. This photo also shows how the chariot wheel heights are staggered to follow the coil pitch.

The stranded wire protruding from the rotor base plate is soldered into a hole in the slip ring, and will later be connected to the lowermost grooved wheel, which doubles as the movable tap on the loading coil.

The limit switch side can be seen in Fig 16. L3 and C4 are visible. The most interesting feature is the steel wire and associated parts, used to link the lower limit switch to the top surface of the limit rotor ring.

Fig 17 shows the bottom view of the servo assembly. The center contact spring is featured prominently here. It is screwed to the two servo mounting plates, and has a raised contact point, that produces a sure one-spot contact to the center contact button of the base. Note the four holes drilled into these plates. These take the screws that attach the assembly to the base cup. They are countersunk to facilitate assembly. Without this, it proved difficult to fit all four screws, given the unavoidable imprecision of home construction.

In Fig 18, the servo assembly is mounted inside the base cup. Notice how everything fits precisely! There is relatively little room wasted. When properly built, the assembly can just barely slip into and out of the cup, without forcing anything. Notice how the fingers of the coil end contact spring are bent into a rounded shape. This allows the coil tube to be installed, with these spring fingers sliding off and on the contact plate during assembly.

The lathe is configured as milling machine to make the slots in the slotted tube (Fig 19). A simple PE core helps support the tube. Because of the softness of the PVC tube, the finish quality of this slotting job was not very good; but this is of no consequence to the performance of the system, as long as the chariot legs can slide freely.

These slots are 14 mm wide, giving the chariot legs enough clearance to avoid binding even if the chariot is displaced from center because of some imprecision. The slots run almost the full length of the tube, excepting only the areas covered by the top and base plates of the rotor.

Fig 20 shows the bottom end of the rotor. The base plate has been pressure-fitted into the tube, but it could also have been glued or fixed with some screws. The slip ring, its sliding surface finished using fine sandpaper and polishing compound, was epoxied onto the servo wheel. This wheel is screwed to the rotor base plate, using two tiny M2 bolts that will make their own threads in the holes provided in the servo wheel.

The screw that holds the servo

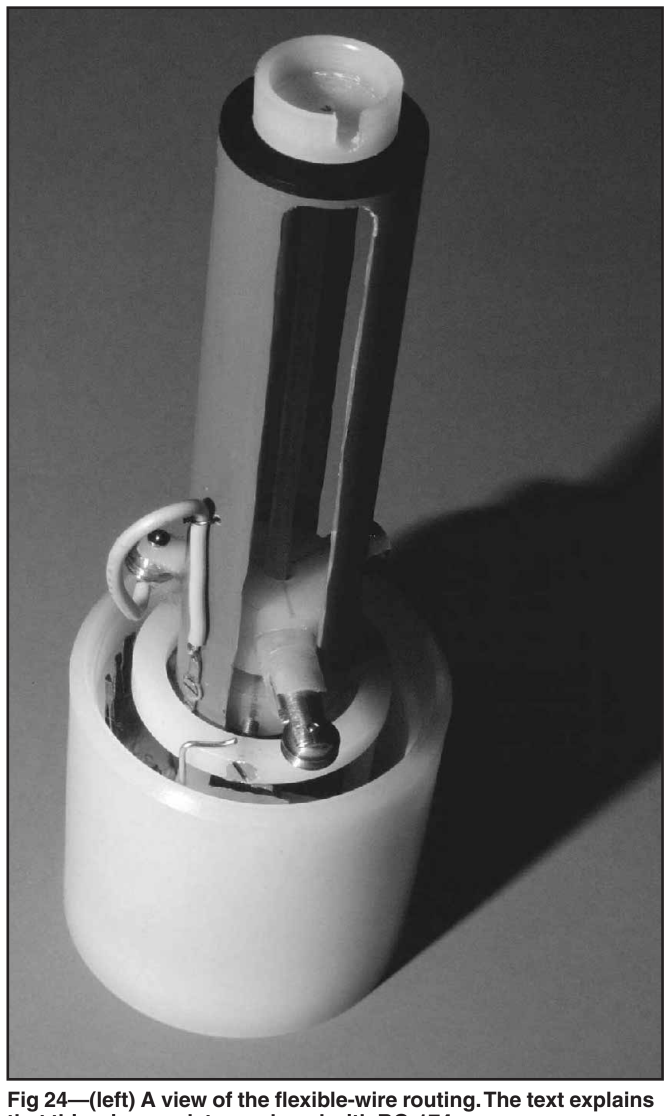


Fig 24—(left) A view of the flexible-wire routing. The text explains that this wire was later replaced with RG-174.
Fig 25—(above) The complete assembly (except for the wire guide) installed in the base cup.

wheel to the servo must be inserted before attaching the rotor base plate to the servo wheel. For mounting the wheel on the servo, the screw head can be accessed through the hole that will later take up the limit coupler axle. The stranded wire runs through a slanted 1-mm hole drilled into the rotor base plate, and continues through a similar hole drilled straight into the slip ring, where the contact is soldered before gluing the parts together.

Fig 21 shows the same assembly from the front. Some solder residue can be seen.

Fig 22 illustrates the complete innards of the antenna, from the centering cup on the top to the servo assembly. You can see how the limit rotor ring activates the two limit switches and how a wire connects the wire coming from the slip ring to the lowermost grooved wheel. More about this wire later.

The limit coupler axle on this picture is the first one I made. It is nylon, which proved to be a little too flexible for best performance. I later replaced it by one made from "durocotton," which is much stiffer. You may ask, "Why not make it from metal?" The reason is that any long conductive piece inside the coil would degrade the Q, lowering the antenna's efficiency. Also there would be risk of arcing at the top of the axle. The entire reason for placing the limit switches down in the base and activating them with mechanical linkages made from plastic, is to preserve the coil's high Q.

To adjust the limit-switch trip points, first the conical spring is bent such that the limit rotor will rest in the center of its range. This allows it to move 2 mm in each direction. Then the lever of the upper limit switch is bent to make it trip when pushing the limit rotor down. Finally, the lower limit switch is adjusted by loosening the screw fixing the plastic block to the coupler wire and sliding the block to a suitable position so that lifting the limit rotor will trigger the switch.

The chariot floats loosely inside the slotted tube. It has considerable clearance, so it won't bind if there is some eccentric motion because of imprecision. The wheel that pokes forward in Fig 22 is the spring loaded, telescoping one.

After a few tests, the simple jumper wire proved insufficient. It bent away from the correct shape and caused friction by scraping against the coil. This did not keep the antenna from working properly, but it slowed the tuning process and forced the mechanics. For that reason, I conceived a simple guide for the wire, made from a segment of PVC

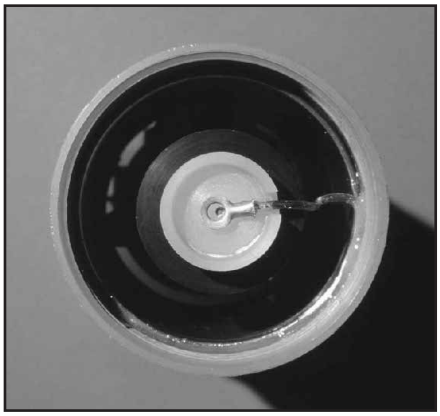


Fig 26—The loading coil assembly has been installed. This was done only for the photo, which shows how the top lead of the coil runs through a slot in the centering cup.

tube, which eliminated this problem. It is mounted with two M3 screws self-threaded into the slotted tube, and some junk-box spacers as shown in Fig 23.

Notice how the three legs are installed at different heights in the chariot centerpiece. Since the coil pitch is 3 mm and the legs are 120° apart from each other, the height difference between each leg and the next is 1 mm. When this is done correctly, the chariot will end up perfectly leveled inside the coil.

Fig 24 is a frontal view, allowing you to see how the wire bends in this guide. It does not bind anywhere. I did all the tests of the antenna using this yellow wire, which has a finely stranded conductor and thick rubbery insulation. In the end, I replaced this wire with a piece of RG-174 coax that has the braid and center conductor shorted. Thanks to the braid, this coax cable has much greater torsional stiffness, without being much more difficult to bend. This proved an advantage in this application.

In addition, the direct connection to the headed axle was used only for the tests. It would eventually break because the wire would be flexing here. Before final assembly of the antenna, I looped the coax a half turn around the chariot leg, soldered it on the top side and wound a few loops of magnet wire around, which I then secured with epoxy. I expect this to last a long time.

Fig 25 shows the complete assembly (except for the wire guide) installed in the base cup. The loading-coil assembly has been installed. This was done only for Fig 26, which shows how the top end of the coil runs through the slot in the centering cup.

Final assembly of the unit starts with the coil tube removed from the base cup. It requires first screwing the top cone into the coil tube, sealing the thread with silicone caulk. The drilled and tapped stainless steel bolt is screwed into the top cone, also sealing it with silicone. Then, using a long screwdriver, the M3 stainless steel bolt that connects the top of the coil can be inserted through the coil. This screw is sealed with low or medium strength Loctite thread locker.

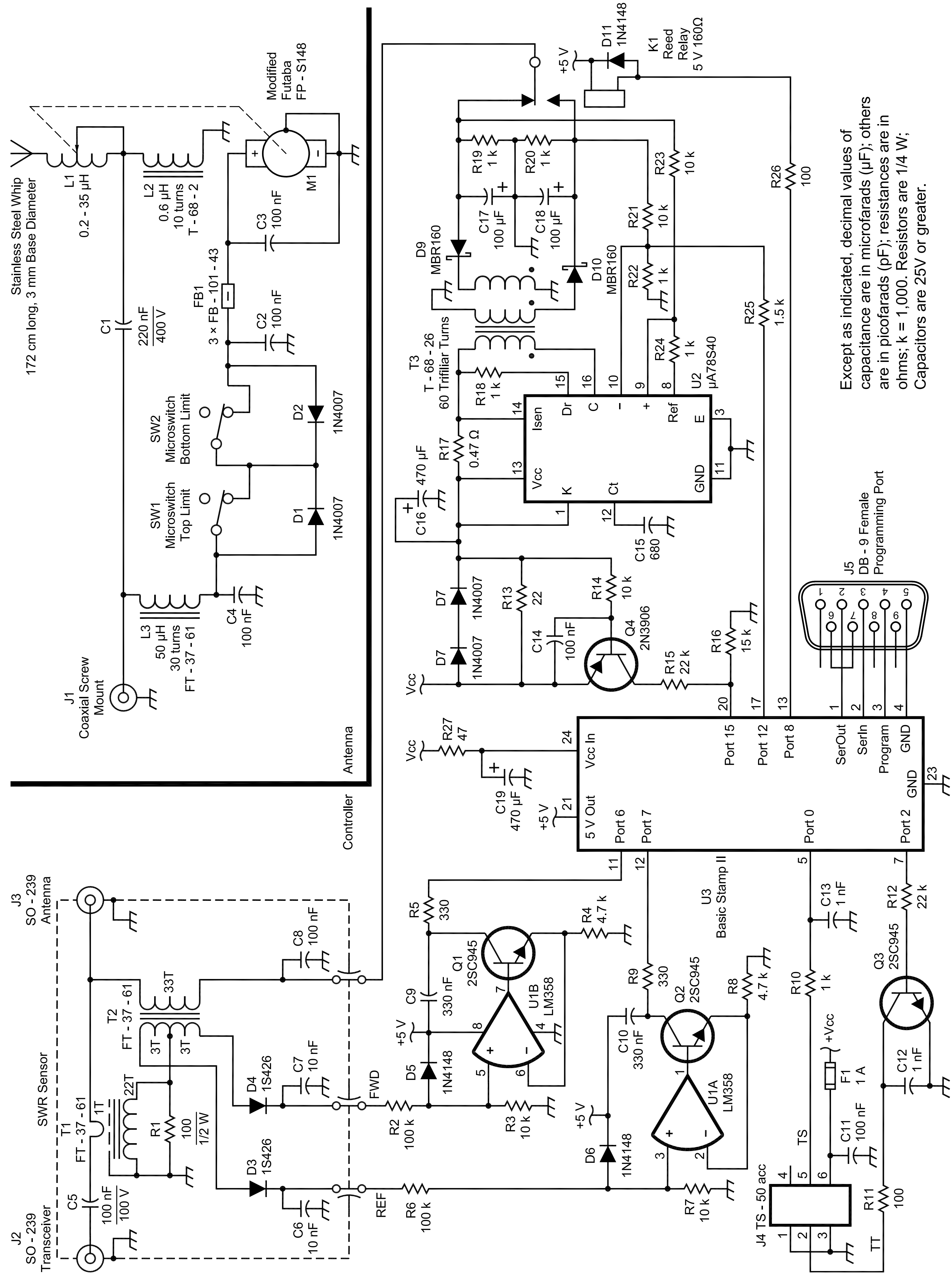
Holding the base cup with installed innards vertically, the loading coil with the top parts is lowered onto it. The telescoping chariot leg is compressed, and the chariot inserted into the coil tube. Then the coil is gently rotated, while manually guiding the grooved wheels, one after another, uppermost first, so they will climb onto the start of the coil wire, in the proper sequence. When this has happened, the coil tube's lower thread is engaged with the base cup, and screwed down a few turns. Before the centering cup at the top jams (it will rarely be perfectly centered during assembly), you need to apply a trick I came up with: Connect a 6-V power supply to the unit! The servo will run, the chariot will screw itself up the coil, and any imprecise centering of the centering cup will transform into an orbiting motion. While the servo runs, *slowly* finish screwing down the coil tube. The centering cup will cleanly latch its notch on the end of the coil wire, and then cleanly engage in the circular groove in the top cone! It's incredibly easy to do, once one has figured out the trick of doing it with a running motor. Before that, I was close to pulling my last hairs out!

Leave this base thread without silicone caulk, until you have completed all testing. Once the antenna has been found to work perfectly, open it a last time and reassemble it with some silicone caulk in this thread.

The Controller

In principle, it would be perfectly possible to use this antenna simply by installing a breakout box equipped to feed selectable-polarity 6-V dc to the coax cable with control via some "Up" and "Down" manual-tuning buttons. These could then be used to manually tune the antenna from inside the car while watching a SWR meter or even while listening. Nevertheless, this is inconvenient and even dangerous if done while driving. Therefore, I built a fully automatic controller for my antenna. Fig 27 is the schematic.

The brain of this circuit is U3, a Basic Stamp II microcontroller. It interfaces to the transceiver, getting commands and returning the proper


Stainless Steel Whip
172 cm long, 3 mm Base Diameter
L1
0.2 - 35 µH
L2
0.6 µH
10 turns
T - 68 - 2
Modified
Futaba
FP - S148
M1
C1
220 nF
400 V
C3
100 nF
FB1
3 × FB - 101 - 43
C2
100 nF
SW2
Microswitch
Bottom Limit
D2
1N4007
SW1
Microswitch
Top Limit
D1
1N4007
J1
Coaxial Screw
Mount
L3
50 µH
30 turns
FT - 37 - 61
C4
100 nF
Antenna
Controller
D11
1N4148
K1
Reed
Relay
5 V 160Ω
+5 V
R19
1 k
R20
1 k
R23
10 k
C17
100 µF
C18
100 µF
R21
10 k
R26
100
D9
MBR160
D10
MBR160
R22
1 k
T3
T - 68 - 26
60 Trifiliar Turns
R25
1.5 k
R24
1 k
R18
1 k
U2
µA78S40
Isen
Dr
C
Ref
E
Vcc
K
Ct
GND
R17
0.47 Ω
C16
470 µF
C15
680
J5
DB - 9 Female
Programming Port
D7
1N4007
R13
22
R14
10 k
C14
100 nF
Q4
2N3906
R15
22 k
R16
15 k
Vcc
R27
47
C19
470 µF
Vcc In
5 V Out
Port 6
Port 7
Port 15
Port 12
Port 8
Port 0
SerOut
SerIn
Program
GND
Port 2
U3
Basic Stamp II
J3
SO - 239
Antenna
SWR Sensor
C8
100 nF
R5
330
Q1
2SC945
R4
4.7 k
C9
330 nF
U1B
LM358
T2
FT - 37 - 61
33T
3T
D5
1N4148
D4
1S426
C7
10 nF
FWD
R2
100 k
R3
10 k
T1
FT - 37 - 61
1T
22T
R1
100
1/2 W
R9
330
Q2
2SC945
R8
4.7 k
C10
330 nF
U1A
LM358
D6
1N4148
D3
1S426
C6
10 nF
REF
R6
100 k
R7
10 k
C5
100 nF
100 V
J2
SO - 239
Transceiver
R10
1 k
C13
1 nF
R12
22 k
Q3
2SC945
+Vcc
F1
1 A
C12
1 nF
C11
100 nF
TS
J4 TS - 50 acc
R11
100
TT
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000. Resistors are 1/4 W; Capacitors are 25V or greater.

replies. It receives forward and reflected voltage samples and has full control over the dc-dc converter that powers the servo. It also takes input from the motor-current sensor.

The circuit inside the shield is the SWR sensor. T1 is a current-mode transformer. On its secondary side is a sample of the current present on the coax line, which is 1/22 of the line current. This current is applied to R1, so that a voltage appears on it having a magnitude of 4.545 V for each ampere on the line.

T2 serves double duty. It feeds the switched 6-V dc from the dc-dc converter into the coax line. It also provides samples of the line voltage, each of which is 1/11 of the line voltage. That is 4.545 V for the 50 V that should be on a 50-Ω line when the SWR is low and 1 A is flowing. The two voltage samples are combined with the current sample in such a way that one adds and the other subtracts. The two combined samples are rectified by D3 and D4, and RF-filtered to provide dc voltages proportional to forward and reflected voltages on the line. This simple SWR sensor is quite accurate for the frequency range needed here, provides sufficiently large sample voltages for the 10-W power level used by most radios for tuning, and can pass 100 W of RF with only 0.5% loss, thus staying cool.

A reactance sensor would be more useful than a SWR sensor, since it could tell unambiguously to which side the antenna should be tuned. However, such a sensor would need to be installed inside the antenna, because of the impedance transformations performed by the impedance matching coil and the unknown length of coax cable between the antenna and the controller. That would require more wires running between the controller and the antenna. For that reason, I chose the SWR sensor, which works through any length of coax. I solved the directional ambiguity problem in software.

The dual operational amplifier U1 is used to interface these sample voltages to the Basic Stamp, which has no analog inputs. This is accomplished by converting voltage to time: With an input voltage of 0-5 V to R2, the transistor Q1 will conduct a controlled current of 0 to roughly 0.1 mA. During rest time, C9 is kept discharged (both sides at 5 V) by the Basic Stamp. To acquire the data, the Stamp switches the pin to input mode, and starts counting the time. C9 charges at a rate determined by the current flowing through Q1, taking a certain time to reach the threshold voltage at which the Stamp senses the change of state and stops counting. This implements a linear analog-to-digital converter with minimal complexity.

The circuit is duplicated for the reflected signal. Diodes D5 and D6 protect the IC during full-power transmissions, when the SWR sensor may produce relatively high voltages during a sudden mismatch. That can happen, for example, when driving through an underpass.

U2 is the core of a dc-dc converter that takes the car's 14 V and converts it to regulated ±6-V outputs. A powdered-iron toroid with a trifilar winding is used, while the µA78S40 provides most of the circuitry, including the power transistor. The output current is actively limited by the converter, eliminating risk of any damage caused by shorting the antenna connector. The outputs are rectified by Schottky diodes; but given the low demands imposed on them, non-Schottky fast diodes could be used without significant disadvantages.

The converter is started and stopped by the Stamp influencing the feedback circuit. In fact, the Stamp could even control the voltage—and thus motor speed—by applying a variable-duty-cycle rectangle wave here, but this possibility is not exploited by the software.

A small, encapsulated reed relay (K1) is used to select either the positive or the negative output to be applied to the coax cable. This relay is always switched cold by shutting down the converter and waiting for the filter capacitors to discharge before changing the relay state. This allows using a very small relay. When applying 5 V to the coil, the relay I found in my junk box drew slightly more current than the Basic Stamp's safe output limit. It has a coil resistance of only 120 Ω; but it switches reliably even at only 1.5 V! So I added R26 in series, limiting the current to 23 mA; this works fine. If you find a 5-V reed relay with at least 200 Ω of coil resistance, you may omit R26. Such relays are quite common.

The circuitry around and including Q4 is a current sensor that will pull port 15 of the Stamp high whenever the dc-dc converter is drawing more than roughly 30 mA. This serves two purposes: The Stamp can quickly detect when the antenna is not installed. As there would be no servo to draw current, the current sensor would not activate, and the Stamp would send an error code to the radio, alerting the operator that he forgot to mount the antenna before tuning it. The other purpose is detecting when the tuning mechanism has reached a limit, because the limit switches interrupt servo current. This information is used in the tuning algorithm.

Q3 provides an open-collector output to drive the data-receive line of the TS-50 transceiver. Many other radios are compatible with this circuit, but the connector style and the software would require modification to suit them.

A DB9 connector allows connection to a PC, both for programming the Basic Stamp and for reading status messages during the antenna testing. This connector is not needed once the unit is ready; it could well be fitted in a temporary fashion.

The antenna circuit is at the top right of Fig 27. This is a simple, base-loaded vertical whip, with an added impedance matching coil (L2). This coil is necessary because short base-fed antennas like this present very low feed-point impedances, as low as 12 Ω on 40 meters. The exact value depends on ground loss, which is not precisely predictable. Without that coil, the best SWR at resonance would be as high as 4:1! Many mobile antennas, both commercial and homemade, exhibit this problem. Some manufacturers choose to provide very lossy loading coils. Since the coil loss resistance adds to the total feed-point resistance, a good match can be obtained by this method at the cost of dismal performance. It's much better to employ a matching coil.

It works on the very simple principle of an L-matching section: An inductance across the feed line and a capacitance in series. This capacitance does not need to exist physically. The loading coil has such high inductance that by removing just a little bit of it, the antenna capacitance is not completely resonated and appears as a little capacitive reactance in series.

As the operating frequency rises, the feed-point resistance of the antenna whip rises too, reaching about 30 Ω on 10 m. This requires a reactance of the matching coil that also varies, increasing with frequency. A very nice coincidence is that a fixed inductance provides just this frequency-proportional rise in reactance! The result is that a fixed inductor will match the antenna impedance very well to 50 Ω over the entire range. The worst SWR I have seen in this antenna has been 1.3:1. It is below 1.2:1 over almost the full range, with a spot-on

Fig 27—(left) A schematic of the complete circuit with the antenna and loading-coil circuits at the upper-right corner.

1:1 throughout 40, 20 and 10 m. This was measured with a digital Daiwa SWR meter and confirmed with a high-precision homemade bridge. With both meters, the reflected power was below detection thresholds.

The controller was assembled on two single-sided glass-epoxy printed circuit boards. One holds all the low-frequency parts. The other holds the SWR sensor; it is installed in a shielded enclosure. Fig 28 shows the copper pattern for both boards, and Fig 29 is a simple component-placement diagram, which should help you to populate the boards, together with the photo (Fig 30).

Notice that C9 and C10 are relatively large. These capacitors must have reasonable temperature stability, so I used plastic-dielectric capacitors. The pin layout of the relay will not necessarily match that of your relay. If so, you will need to modify the board or use some wires to connect it.

The toroid for the dc-dc converter was mounted using a bolt and foam rubber washers (Fig 31). This is quite elegant, but you could as well use hot-melt glue to hold it onto the board. This transformer is wound with thin magnet wire, the exact diameter being uncritical. I used a wire about 0.25 mm thick. The solder side of the board (Fig 32) shows that this is easy to assemble. Nothing fancy here!

The SWR sensor was enclosed in a shield made in cheap, quick-and-dirty fashion from some 0.1-mm copper sheet (Fig 33). Feedthrough capacitors are used for the three dc leads entering the shield. The value of these feedthrough capacitors is uncritical. Mine were unmarked junk-box parts, probably around 1 nF.

The coaxial connectors were first bolted to the external case and the open shield. Then the PC board was soldered to the center pins (Fig 34), followed by assembling and soldering the shield. The entire sensor can be removed from the external case without removing the connectors.

The case was built from two pieces of 1-mm aluminum sheet. The SWR sensor is held in place just by the coax connectors and the soldered shield, while the main board is mounted with four screws and spacers (Fig 35).

After testing and before closing the controller, liberal amounts of hot-melt glue were used to immobilize anything that could otherwise shake and become loose (Fig 36). After all, I'm using this controller in a 4WD

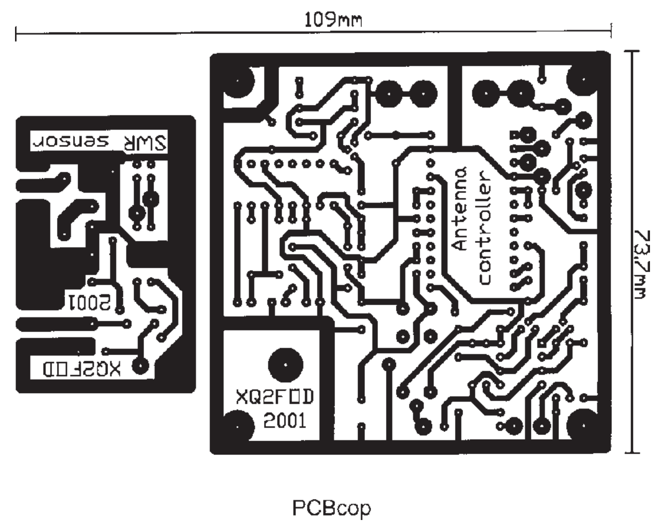


Fig 28—Etching patterns for the two circuit boards. Dark areas represent the copper traces. If you make boards from these patterns, enlarge them to the dimensions shown.

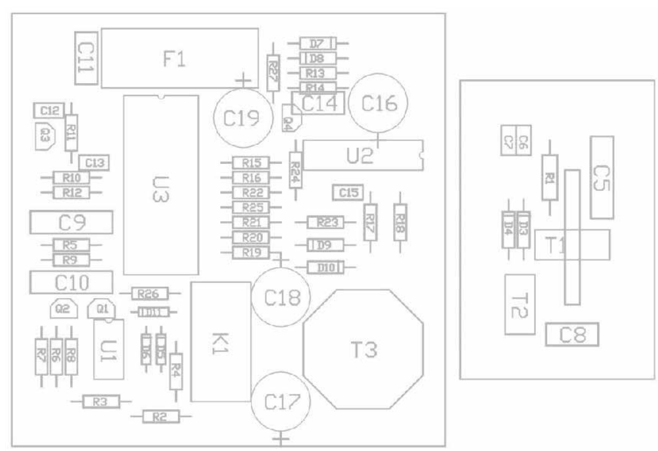


Fig 29—Part-placement diagrams for the circuits.

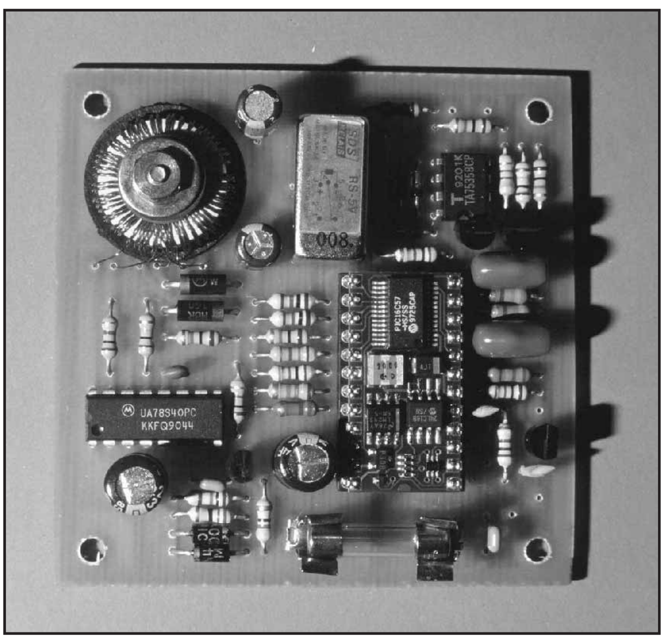


Fig 30—The completed controller on its single-sided glass epoxy printed circuit board.

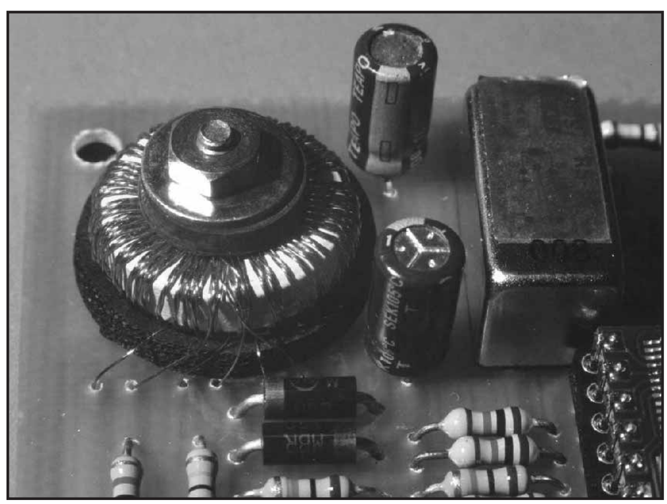
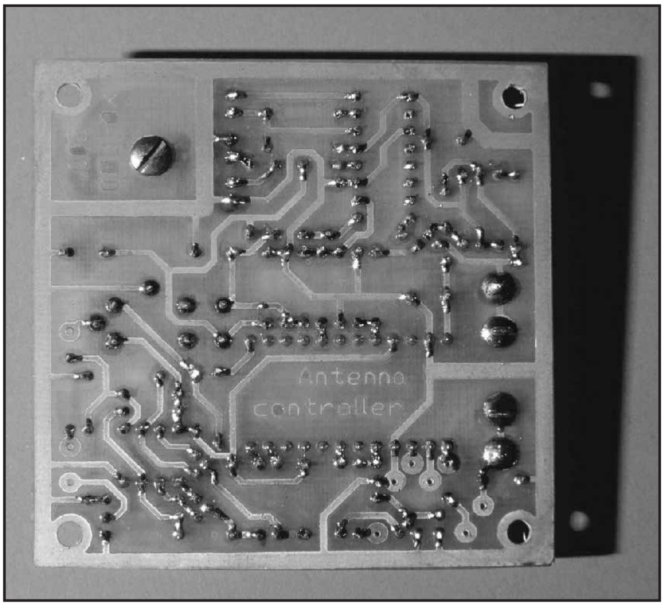


Fig 31—The toroid for the dc-dc converter was mounted using a bolt and foam rubber washers.



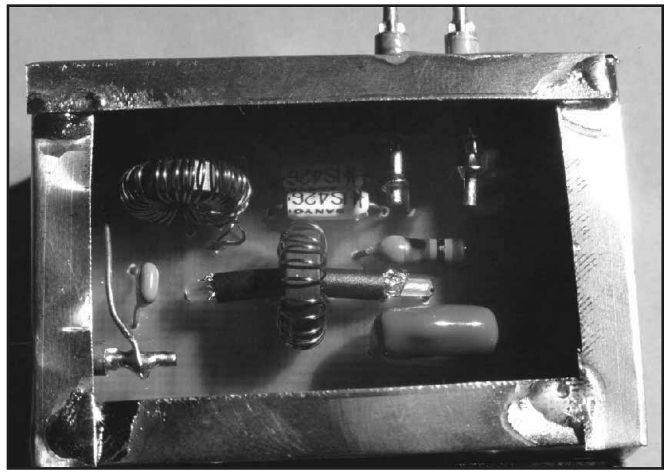


Fig 32—(left) A solder-side view of the controller circuit board.

Fig 33—(above)The SWR-sensor circuit is constructed inside a shield box made from thin (0.1 mm) copper sheeting.

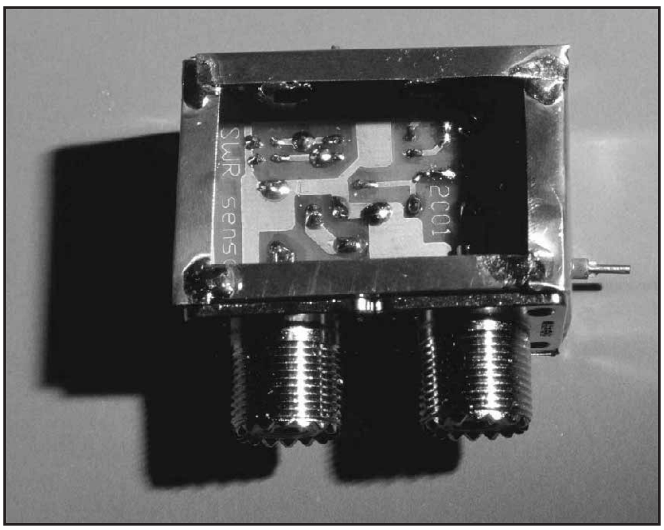


Fig 34—The solder side of the SWR-sensor board, showing how the SO-239 connectors are mounted.

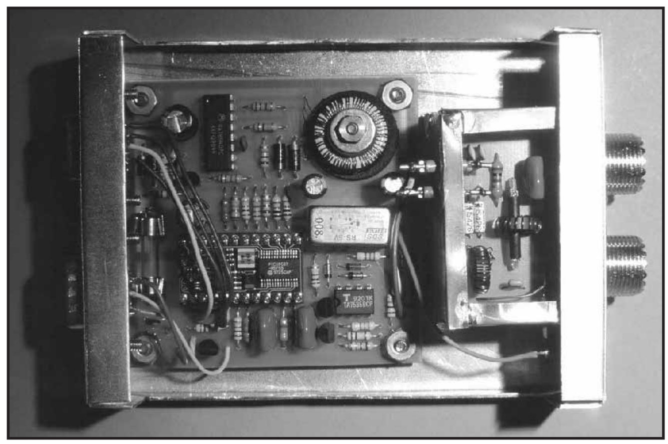


Fig 35—The case was built from two pieces of 1-mm aluminum sheet. The SWR sensor is held in place by the coax connectors and the soldered shield, while the main board is mounted to the case with four screws and spacers.

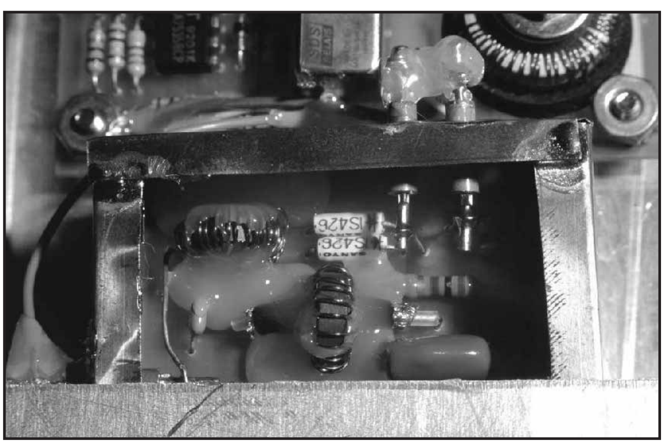


Fig 36—After testing and before closing the controller, liberal amounts of hot-melt glue were used to immobilize anything that could otherwise shake and become loose.

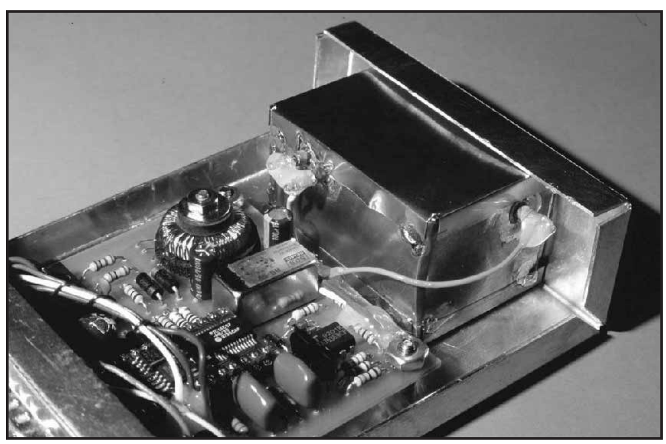


Fig 37—The shield of the SWR sensor is completed by top and bottom covers that are soldered at only a few spots. I used this extensive shielding to prevent any noise from the Basic Stamp or dc-dc converter oscillator getting to the receiver.

vehicle that I drive over all kinds of terrain, so some considerable mechanical stress is unavoidable. Moreover, it's no fun to disassemble half of the car to dig out a failing controller, only to find a broken wire or solder joint! Prevention is the best remedy.

The shield of the SWR sensor is completed by soldering the top and bottom covers in place at only a few spots (Fig 37). These are easy to remove, should that ever become necessary. I used this extensive shielding to prevent any noise from the Basic Stamp or the dc-dc converter oscillator getting into the receiver. I did not try the circuit without the shields, but maybe such care is not necessary. Still, it's cheap insurance against noise problems.

Fig 38 shows the completed controller. One side carries the coax connectors, while the other has the power/data connector and the programming connector. The cover of the box is shaped such that it can be used to mount the controller to the car using sheet metal screws, with ample freedom of where to drill the holes for them. I did not apply any finish to the case. That was unnecessary, since the location chosen for the controller is well protected from water, abrasion and completely out of view.

The Software

The nicest thing about the Basic Stamp, which offsets its relatively high cost, is the ease and speed of programming it. The program code for the Stamp is part of the download package for this project (see Note 1). It can be read and modified by any text editor, such as Windows' *Notepad.* For loading it into the Basic Stamp, however, you need the proper software, which is available free of charge from Parallax (**www.parallaxinc.com/**), the manufacturer of the Stamp.

Here is a description of the software flow. After powering up, the controller waits for the radio to prompt for any connected tuners, and replies to it just like an AT-50 tuner would do. This will light the "AT" indication on the radio's display, and allow further commanding of the controller. It will then wait for such further commands.

When a command arrives, the controller will acknowledge and then execute it. The important commands are "start tune" and "band-set." There is also an "enable tuner" command, which is interpreted just like "start tune," and a "stop tuner" command, which is acknowledged but otherwise quietly ignored, since the tuning portion in this antenna cannot be switched off!

When a tune command is received, the controller first looks at the included band information. If the radio is asking to tune on a band below 40 m, the controller answers with an error code, because these bands are not supported. If the band is 40 m or higher, the controller checks whether the SWR is totally sky-high or if something close to a resonance can be seen. If the antenna is close to resonance, the controller immediately starts the fine-tuning algorithm. Otherwise, it starts the coarse-tuning, or "search mode."

In search mode, the controller first tries to determine whether the coil tap must be moved upward or downward. If the antenna was previously tuned without shutting off the radio, the previous band will be stored in memory and the controller will compare this to the present band, thus easily determining whether it must tune upward or downward. If the previous band is unknown, the controller applies a guess. If the band is 40 m, it will tune downward; and if the band is any other, it will tune upward. This should provide the shortest tuning time, because 40-m resonance is achieved with the tap very close to the bottom, while all other bands need the tap somewhere in the upper half of the coil.

The coarse tuning runs the servo at full speed, while continuously measuring SWR, until it drops below a value of about 5:1. In fact, the internal SWR measuring routine does not calculate the actual SWR, but simply a relative resonance value: 10 means far offresonance (infinite SWR) with higher values indicating lesser SWR. Typical values for the fully tuned antenna are between 20 and 50. The value at which the coarse tuning ends can be preconfigured in the constants' declaration section at the start of the program.

Suppose you had the antenna tuned on 12 m, switched off, switched on again later, changed to 17 m and hit the **AT TUNE** button. The controller would start tuning up the coil, in wrong direction. The servo will drive the tap to the end of the coil without the controller finding a resonance. As soon as the limit switch stops the servo, the controller will sense the current drop and start the servo in the opposite direction, quickly finding resonance. If changing servo direction does not make the current restart, however, the controller will assume that there is a problem and send an error message to the radio, which will then emit its characteristic error beeping.

This situation happens when the antenna was not installed or cannot be tuned, such as if the car was parked with the antenna touching a wet tree. When the resonance has been detected, fine tuning starts. This consists of moving the servo in small steps by controlling the runtime and measuring the resonance level between steps. The controller tries to find and overshoot the best possible value by stepping until the resonance value decreases, then changes direction and repeats this. After this step, its position relative to perfect resonance is known. The controller now moves the servo in very small steps in the correct direction, until it hits precisely the optimal setting. It then sends a code back to the transceiver, indicating that the tuning procedure is complete.

If at any time during the tuning process the radio drops out of the transmit mode or the RF power level

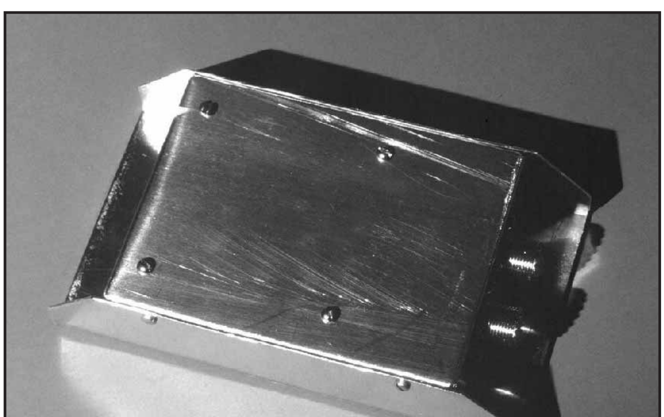


Fig 38—This is the completed controller. One end holds the coax connectors, while the other has the power/data and programming connectors. The box cover is shaped to facilitate mounting the controller in a vehicle by means of sheet metal screws, with ample room to drill the holes for them.

becomes too low or too high for tuning, the controller will stop the process and send an error code to theradio. The time taken for fine tuning is just about five seconds, and the worst-case resonance search still completes in well under one minute. Average band changing time is 15 seconds. While this is slower than some automatic tuners, it's still quite fast compared with some commercial antennas and very much faster than getting out of the car, changing resonators and fine-tuning a whip!

The software source code has enough comments to allow understanding of it. When adapting it to different radios, it may be necessary to change not only the commands and answers, but also the serial-port parameters. If you want to modify to the program, I suggest you download the Basic Stamp II manual from the Parallax site. The hardest part about adapting the controller to another radio is finding out exactly what commands the radio sends and what answers it expects!

Installation in a Car

Fig 39 shows where my controller was installed. My Nissan has lots of unused room between the external skin and the internal plastic trim. I removed a plastic cover close to the rear seat and installed the controller there, using four sheet-metal screws. Then the plastic trim was reinstalled, completely hiding the controller. All cables were routed under plastic trim panels, making a neat and clean installation. Also notice that both antennas (HF and VHF) were installed through the roof above the interior light fixtures. This allows easy access by removing the lights, and hides the antenna installation from the inside.

By the way, do you know how to drill a hole into the roof of a new car? Here is the recipe: First, mark where you need to drill and check several times that it is correctly marked. (After drilling, you cannot go back.) Then install a sheet-metal drill bit of the proper size (12 mm for this antenna) in your electric drill. Place the drill-bit tip against the roof, firmly shut your eyes, collect all your courage, clench your teeth and press the trigger. It will be over very quickly, so it won't hurt for long! Soon the antenna will close the two wounds: one in the car roof and the other in your heart!

Joking aside, I would like to stress that the best way to install a mobile antenna, especially one for HF, is through a hole in the car roof. I have done this on all my cars, and never had any trouble from it. When reselling a car, I simply close the hole with some discrete, properly sized cover. Although I've pointed out the hole to prospective buyers, I have never come across one who saw the hole as a problem.

The TS-50 was installed under the center console (Fig 40). For this to be possible, I removed a plastic panel (I stored it for the day I resell the car) and made a new panel from vinyl-covered aluminum, to close the areas around the radio. Two steel bars holding the console were bent slightly outward and spacers were installed for a neat installation.

A ClearSpeech DSP filter was installed under the TS-50 using double-sided foam tape. While such a filter cannot make signals audible that are deeply buried in noise, it really helps make HF mobile operation less tiring by removing almost all noise. Without the filter, when you drive under power lines there will be deafening noise. With the filter, the signal fades while you are under power lines, but there is no loud noise. Comparing with earlier filter-less times, I find that the DSP filter allows me to do about three times more mobile operating than I did without it.

Speaking about noise, my car is quite RF clean. It uses a carburetor and has no electronic engine management systems. The only electronics are the voltage regulator and a few simple circuits such as the rear-wiper controller. This helps a lot in reducing RF noise! The ignition system is almost noise-free as it comes, so there was no need for improvement. My main noise sources when traveling are roadside power lines, which unfortunately are very common and not within my control.

The VHF transceiver, a TM-241, fits nicely into the hole left after removing the ashtray. I'm a nonsmoking ham and have a lot more use for a radio than an ashtray! I do not use the internal speakers of the radios. Instead, I use the four existing car speakers. I've wired a four-pole relay to control them. When either or both transceivers switch on, the relay connects the left-front speaker to the TM-241, the right-front speaker to the TS-50 (via the DSP filter) and the two rear speakers remain connected to the car radio. When both transceivers are off, all four speakers are connected to the car radio for good music reproduction.

The power for the two transceivers is brought through the firewall by a pair of heavy wires, connected to the battery and properly fused there. These wires split to carry the power connectors for both radios and some additional outlets for a GPS receiver, altimeter, notebook computer and other playthings.

The two microphones are hung on two wires shaped into mike hooks and secured to the gear and transfer levers (see Fig 40). They are always within easy reach. The microphone of the TS-50 has four programmable but-

Fig 40—The author's mobile installation. Notice the microphones hung from wire loops on the gear-shift levers.

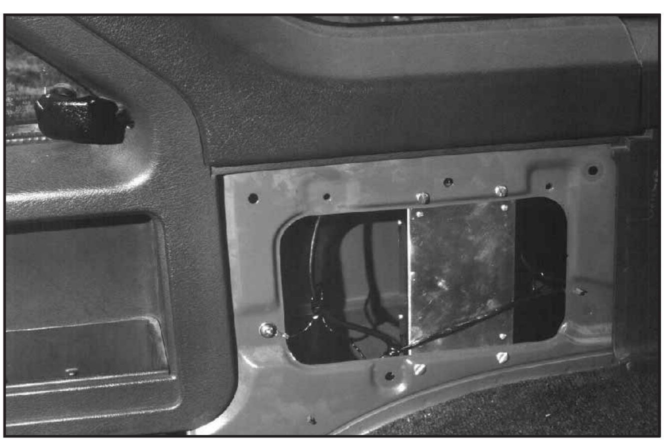


Fig 39—The controller installed under a trim panel in the author's Nissan.

tons. I programmed two of them to step the bands up and down, and another for activating the **AT TUNE** function.

The *modus operandi* is simple: I take the mike, switch to the desired band, use the up/down buttons to tune to a clear frequency close to the one I need and press the **AT TUNE** button. My antenna tunes to the exact frequency and when it has arrived there, the radio beeps shortly and drops back into receive mode. I then go to the exact frequency and make the contact. As simple as that—and on all bands from 40 to 10 meters!

Performance

Some of you may ask why I didn't include coverage of 80 and 160 m in this design. The reason is performance. In any antenna, the radiation efficiency is given by the ratio between the radiation resistance and the total loss resistance. In physically short antennas the radiation resistance is low and becomes very low when the antenna is very short. Loss resistance, on the other hand, is given mostly by ground loss and the equivalent series resistance of the loading coil. Let's see what happens on different bands.

On 10 m, this antenna has a radiation resistance of about 20 Ω. The coil resistance is negligible and ground loss is probably less than 3 Ω, because most ground current is captured by the car body, which is large enough on this wavelength to effectively shield the antenna from the ground. Thus, the efficiency on 10 m will be around 85%, which will be indistinguishable from a "perfect" antenna. If we double the wavelength to 20 m, the radiation resistance drops to around 5 Ω. The loading coil will need to be tuned to a reactance of roughly 500 Ω. Given a Q of 250 (not higher, because part of the coil is shorted!), this would produce about 2 Ω of coil loss. The ground resistance may be around 5 Ω, placing the total efficiency of the antenna at somewhat above 40%. This is about 4 dB down from a perfect antenna. The loss can be noticed, but not too much.

On 40 m, the situation changes a lot. The radiation resistance is down to only 1 Ω! The loading coil needs to provide a reactance of roughly 1400 Ω. The Q is better, approaching 350 here because almost no turns are shorted. Thus, we get a coil loss of 4 Ω. The ground loss is higher on this band because so much more ground current returns via the soil. It will probably be around 6 or 8 Ω, depending on terrain. The result is that the efficiency of the antenna on 40 m is less than 10%! This represents a loss of 10 dB compared to a perfect antenna. Such a loss is easily noticeable, and you will get lower signal reports than from the home station, but a lot of contacts can still be done.

On 80 m, the situation turns dramatic. A whip this size would have a radiation resistance of only 0.2 Ω! Coil loss would be around 10 Ω, even for a very large coil, and ground loss should be expected at about 15 Ω. The result is less than 1% antenna efficiency, which represents a loss of more than 20 dB. On the generally noisy 80-m band, this is such a severe handicap that I decided to stay out altogether.

On 160 m, of course, the situation is much worse: Radiation resistance would be only about 0.07 Ω, coil loss at least 20 Ω, ground loss no less than 30 Ω and the total antenna "gain" would be close to –30 dB! I doubt if any useful work could be done down there. Operating the very low bands from a car requires different approaches, either using oversize whips with top loading, or using loops made from very large copper tubing. Both are mechanically cumbersome.

Comparing this antenna side by side to several commercial ones, installed on other hams' cars, the performance turned out to be equal to the best antennas of comparable length, and surpassing many of the cheaper models available. At the same time, this antenna is lighter and easier to dismount and store than any comparable commercial antenna I have seen. The convenience of automatic tuning to any frequency is matched only by a few commercial antennas. The tuning speed was surpassed by one, but that antenna uses a large tuner box and it was expensive.

Many commercial antennas were very heavy and, being bumper-mounted, had problems of detuning while swaying. My antenna is essentially free of detuning, even if the thin whip sways widely. The reason is simply the roof mounting: The whip can sway a lot without getting significantly closer to the car or the ground.

You may have noticed that no spring is used to mount this antenna. In fact, I need to be careful not to hit any hard object with the loading coil, as this would cause damage. The highly flexible whip is very resistant to impact, however. I have "caressed" trees at highway speeds, and a few times, I forgot to take off the antenna before driving into a garage. In all cases the whip flexed out of harm's way.

During a long trip to remote southern Chile during December 2001 and January 2002, I thoroughly tested this antenna. It withstood heavy rain, high-speed driving and lots of abuse on overgrown forest tracks, while providing many contacts on all bands. The excellent results obtained prompted me to make the design available to the ham community.

Manfred was born in Chile in 1965. He became interested in electronics at the age of 12, first building crystal radios, then audio amplifiers, other audio equipment, communication equipment and so on. In 1980, he got a novice license (the minimum age in Chile for a license was 15 years). He later upgraded to general and then to his present superior class license (equivalent to the US Extra class), as soon as he crossed the age limits for them.

Until 1990, all his ham-radio operating was done exclusively with homemade gear. Today still, a large part of his station is homemade. His main ham radio activities are of technical nature, but he also enjoys operating in voice and digital modes, especially on satellites. He operates fixed and mobile, including aeronautical mobile from a powered paraglider. He also has participated in the building and installation of eight VHF repeaters, various repeater linking systems, and two mountaintop packet nodes. On a historical side, he enjoys restoring and collecting antique radios.

He has written technical articles for several ham radio and general electronics magazines, among them the "13.8 V, 40 A Switching Power Supply" published in QST *and the* ARRL Handbook*.*

He studied electronics at the Universidad Técnica Federico Santa Maria, and since 1989 he has been employed as electronic engineer at the La Silla Astronomical Observatory, which belongs to the European Southern Observatory (visit us at **www.eso.org**). *His job there is building toys for the astronomers, and fixing whatever they break.*

His other hobbies include photography, mountain climbing, model-airplane building and flying (complete with ATV!), free flying and music (mostly classical). He has a Web site about his hobbies, entitled "Homo Ludens" (the playing man) at **www.qsl.net/xq2fod**.

Theory of Intermodulation and Reciprocal Mixing: Practice, Definitions and Measurements in Devices and Systems, Part 2

A expert shows us how to achieve better IMD and IP measurements.

By Ulrich L. Rohde, KA2WEU/DJ2LR/HB9AWE

Synergy Microwave Corporation
201 McLean Blvd
Paterson, NJ 07504
ulrohde@aol.com

Part two of this paper will deal with many practical aspects. We are going to look at the measurements done directly on mixers and analyze the possible pitfalls of such measurements and receiver systems. I will propose a novel interlaced dual-loop AGC system, which drastically improves intermodulation-distortion (IMD) performance in actual use. This is because the receiver system will initially maintain a 40-dB signal-to-noise ratio, and after reaching this, will increase at a reduced rate based on the shared AGC distribution.

Measurement of Mixers

Three elements determine the dynamic range of a receiver: the preamplifier (mostly used at frequencies above 30 MHz, unless electrically small antennas or active antennas are used), mixers and amplifiers. While the measurement principle is the same as for the mixers, we will concentrate on amplifiers for the moment.

For determination of the intercept point (IP) of an (ideal) receiver or a single component (for example, a low-noise amplifier, mixer), an assumption is made that at low-impact levels, the IMD products behave according to a square law (IP_2) or to a cube law (IP_3). They are typically selected to be approximately 1-5 µV or its equivalent in dBm (5 µV = –93 dBm; 1 µV = –107 dBm for 50 Ω). Interfering signals are applied to the device under test at power levels that lead to measurable IMD products. The input *IP* is then calculated according to Eq 19:

$$IP_{\mathrm{n,IN}} = \frac{(P_{\mathrm{OUT}} - P_{\mathrm{IM}n})}{n-1} + P_{\mathrm{IN}} \qquad \text{(Eq 19)}$$

where,
P_{OUT} = power of output signal (dBm)
$P_{\mathrm{IM}n}$ = power of intermodulation product (dBm)
P_{IN} = power of input signal (dBm)
n = order of intermodulation product

The output IP results in

$$IP_{n,\mathrm{OUT}} = IP_{n,\mathrm{IN}} + G \qquad \text{(Eq 20)}$$

where
G = gain of the receiver or device (negative for loss, in dB)

This means that for a passive device, such as a mixer, the output intercept point is reduced. The inverse is also true, meaning that the input intercept point of a passive device is always higher than that of the output.

When measuring receivers, the input signals are converted to an IF or to the audio band and a comparison method is used for determination of the IM products. An in-band test signal is applied to the receiver and the power level of this signal is increased until the signal appears in the audio band so that the signal plus noise is 3 dB above the noise floor. This power level is called P_{NF}. Next, an off-channel two-tone signal is applied, and the power levels of the two tones are adjusted in tandem until the IMD product plus noise produce a level 3 dB above the noise floor. From these measurements, the input intercept point of order n can be calculated as:

$$IP_{n,\,\mathrm{IN}} = \frac{(nP_\mathrm{D} - P_\mathrm{NF})}{n-1} \qquad \text{(Eq 21)}$$

where

P_D = power of input signal producing IM products (in dBm)

P_NF = power of input signal reaching noise floor (in dBm)

n = order of the intermodulation product

The IMD dynamic range (*IMDR*) is the ratio of the level of the two off-channel signals producing an in-channel IMD product to that of a single in-band signal producing the same power. This statement may be confusing because IMDR is the ratio of two powers expressed in decibels, while the rest of the equation is a difference (in dBm).

$$IMDR = P_\mathrm{D} - P_\mathrm{NF} \qquad \text{(Eq 22)}$$

The *IMDR* is related to the input intercept point by:

$$IMDR_n = \frac{\left(IP_{n,\,\mathrm{IN}} - P_\mathrm{NF}\right)\left(n-1\right)}{n} \qquad \text{(Eq 23)}$$

In modern receivers, very high *IP*s are common. Good receivers have a third-order input-intercept point (IIP_3) of +35 dBm and a second-order input intercept point (IIP_2) of +80 dBm. Assuming the noise floor of a receiver is –130 dBm, then the IMDR_3 calculates to 110 dB.

For accurate calculation of the IP_3, we must ensure that the cubical behavior of the IP_3 curve is still valid. The applied power levels must be well below the 1-dB compression point of the receiver. Normally, the 1-dB compression point is 10-15 dB below the IP_3. Using the above example, the power for measuring the *IM* product is –20 dBm, and this is well below the 1-dB compression point of +20 dBm.

The above statements are only correct for single devices such as one mixer or one amplifier. The 3-dB-per-dB law applies only for those single devices. In the case of an RF front end of a receiver, this is not necessarily true. I am not addressing the influence of reciprocal mixing now but just the causes of intermodulation. In the case of receiver front-end switching diodes, as well as IMD products of the first crystal filter, all can occur at the same time. Inside the filter, the ferrite cores will also add to distortion. From a purely scientific view, we will not be able to distinguish what contributes what, but the sum of all products will show up.

Especially when testing a receiver, one never knows exactly where the IMD products occur. Most test setups require a dynamic range of up to 100 dB, spurious free, because (for reasons that will be explained) they have some internal IMD products and level differences for low-level IMD products. Thus, when measuring at a very low level, IMD products do not behave according to 3-dB-per-dB, but by some other funny numbers. ARRL testing has been subject to some comments, as their results do not always follow the 3-dB-per-dB rule. Likewise, the relationship between minimum discernable signal (MDS) and the third-order intercept point to be used for calculation of dynamic range does not provide reasonable answers. Complete receiver systems are just not inherently linear; based on the gain distribution, not all numbers are meaningful. More comments on this will follow.

I am getting ahead of myself, though. As I have stated here, for those measurements required to be at sufficiently high levels for receivers whose IP_3 is between 20 and 30 dBm, I recommend doing the measurements at 2 × –10 dBm at the receiver input. In this case, the dominant source in the chain for IMD products will be active and the 3-dB-per-dB law will work properly. The –10 dBm level may not be valid for all systems, but at least it generates a traceable standard.

Another issue is the use of a spectrum analyzer. Since the year 2000, spectrum analyzers have had a state-of-the-art on-screen resolution of between 100 and 120 dB. The lower level is determined by the noise figure of the spectrum analyzer, typically 20 dB, and the upper level is given by IMD products generated at the first mixer in the spectrum analyzer. Spectrum-analyzer measurements will use single devices and will terminate the device under test with its internal 50-Ω termination. A typical modern spectrum analyzer has an input intercept point of +20 dBm. By adding 30 dB attenuation, the resulting intercept point is 50 dBm and, therefore, all the spurious products will come from the test object or the device under test and not from the analyzer. In addition, reciprocal mixing does not apply here. It would be nice if all receivers had an IF monitoring output after the first IF, in which case the true front-end performance could be measured.

As to the accuracy of measurements, the use of a spectrum analyzer—with a built-in tracking generator for calculation—provides better than 1-dB accuracy. On the other hand, a practical receiver has a noncalibrated S-meter that needs to be calibrated for such tests. Many receivers nowadays don't have analog meters or high-resolution digital outputs with three digits of resolution, but have a bar-graph display. Unless the setting can be selected so a bar just starts, there can be a 6-dB inaccuracy problem, as these bars typically only jump in 6-dB steps. The AGC resolution on those bars makes setting a level for the two interfering tones difficult. One may need to vary those tones by up to 6 dB to get reproducible calculated values.

Measuring IP_3 in Mixers

The quality of a mixer has a great impact on the performance of a receiver overall. In addition to conversion loss and isolation, IP_3 is the key factor in the specification of a mixer. Measuring the IP_3 of a mixer is a task that needs very good measuring equipment and a lot of experience. If it is done without precaution, the results may be inaccurate and differ by tens of decibels from the correct values.

The standard procedure of measuring conversion loss and LO/IF isolation of mixers is to provide an RF signal and an LO signal with two independent signal generators having the required impedance, typically 50 Ω, and high internal isolation. The procedure investigates the power level of the converted output and LO signal at the IF frequency with a spectrum analyzer. For IP_3 measurement, two RF signals are used at adjacent frequencies. The frequency offset between the generators is typically 100 kHz to 1 MHz. Smaller offsets should not be used because the RF stages are limited in processing RF signals and thus IP_3 increases at very low offsets. The signals of the two generators are added via a hybrid coupler or combiner and injected into the RF port of the mixer. Fig 29 shows the spectrum of the input signal to the mixer and the intermodulation products (IM_3) at the frequencies ($2f1–f2$) and ($2f2–f1$), which are generated in the nonlinear mixer and then down-converted by the LO into the IF band. These signals represent the unwanted and interfering signals that limit the dynamic range of the mixer.

According to Eq 19, the input IP_3 of the mixer is given by:

$$IP_{3,\,IN} = \frac{(P_{IF} - P_{IM3})}{2} + P_{IN} \qquad \text{(Eq 24)}$$

where

P_{IF} = power of down-converted IF signal (dBm)

PIM_3 = power of intermodulation product (dBm)

P_{IN} = power of input signals *f1*, *f2* (dBm)

A standard test setup for IP_3 measurements is shown in Fig 30. The signals of two generators are added in a hybrid combiner and fed into the RF port of the mixer. Since most generators have only 15-17 dBm output, the LO signal is amplified to provide the necessary power level, that is, +20 dBm. The ARRL also measures with 20-kHz spacing.

Examination of a Simple Test Setup that Handles Only Medium Values for IP_3

Both generators provide their signals *f1* and *f2* to the hybrid combiner. The isolation between the generators is given by the isolation of the combiner itself plus the output attenuators of the generators, which are used for power-level control. Due to the finite isolation and reflection from poor termination, some energy from each generator appears at the other and nonlinearities in the generator output stages generate IMD in the test signal. The interference contribution of the two generators can be measured at point A in Fig 30: IM_3 products at the frequencies 2*f1*–*f2* or 2*f2*–*f1*. These IM_3 products will be injected into the mixer and degrade the measurement accuracy. This is an ideal case, since it assumes a perfect termination for the IF load, if a load such as a spectrum analyzer is used. The spectrum analyzer is typically operated at 30-40 dB attenuation with a useful dynamic range of 100 dB using 10-Hz resolution bandwidth. In this case, the spectrum analyzer will not show any IMD products.

Example: Assuming a Mixer with 10-dB Conversion Loss and an infinite IP_3

P_{IN} at *f1* and *f2* = 0 dBm; measured IM_3 at point A = –50 dBm; measured down-converted IM_3 product in the IF band = –60 dBm; measured IF output power = –10dBm. Using Eq 24, an input IP_3 of 25 dBm is calculated. Therefore, the test setup itself has an IP_3 of 25 dBm!

If any mixer is now connected at test point A, the injected IM_3 products of the test setup and the IM_3 products generated within the mixer will interfere. What will be the measured result?

If the mixer itself has an IP_3 of about 30 dBm, it cannot be measured with this test setup. Mixers with an IP_3 much lower than 25 dBm can be measured using this test setup with barely sufficient accuracy.

The frequencies of the IM_3 products are 2*f1*–*f2* and 2*f2*–*f1*. The two terms 2*f1* and 2*f2* will be also provided by the generators as harmonics. In the test setup, at point A, the harmonics 2*f1* and 2*f2* can be measured. Normally, harmonics of generators are about 30-40 dB below the fundamental frequency; the higher the output power of the generator, the lower is the suppression of the harmonics. A broadband mixer, such as DUT, converts these harmonics into the IF, which interferes with the desired down-converted signal.

In practice, at least six frequencies: *f1*, *f2*, 2*f1*, 2*f2*, 2*f1*–*f2* and 2*f2*–*f1*—are injected into the mixer instead of only two (*f1* and *f2*). See Fig 31.

Optimizing the Test Setup

The optimized test setup is shown in Fig 32. Interference produced by both generators because of insufficient isolation, and further generating unwanted IM_3 products, can be reduced by inserting attenuators in each signal path. The attenuators also improve the load matching of the combiner, which results in better isolation in the combiner itself, because the combiner achieves certain isolation levels only if the load impedance is correct. Additional isolators can be used to achieve greater isolation. The drawback of the isolators is a reduced bandwidth compared to that available with attenuators. Alternatively, high-linearity

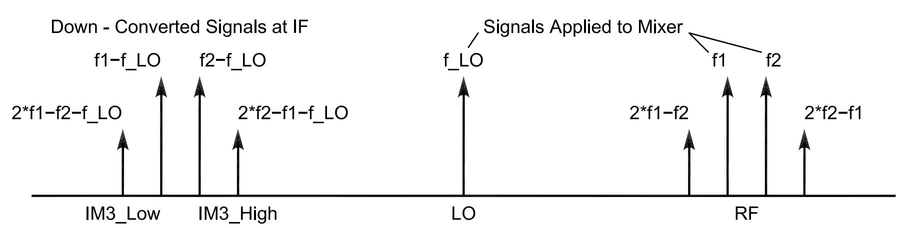


Fig 29—RF signals *f1*, *f2* and f_{LO} are applied to the mixer and are down-converted to the IF together with intermodulation products, IM_3.

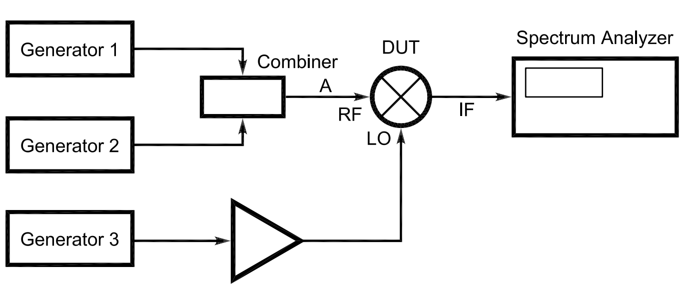


Fig 30—A minimum test setup for IP_3 measurements. (The concept and validation of the mixer test bed was supplied by Guido Baumann, Detlev Hollmann, and Roland Heilig, of the German firm HBH under contract to the author.)

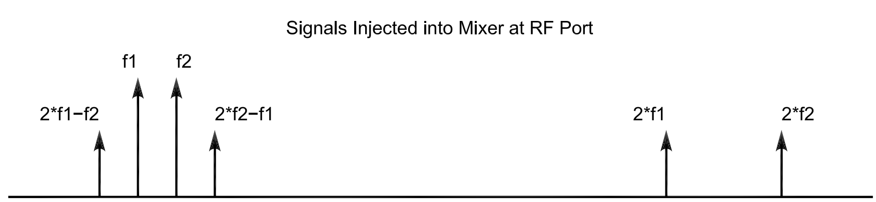


Fig 31—Signals injected into the RF port of the mixer in an IP_3 test setup.

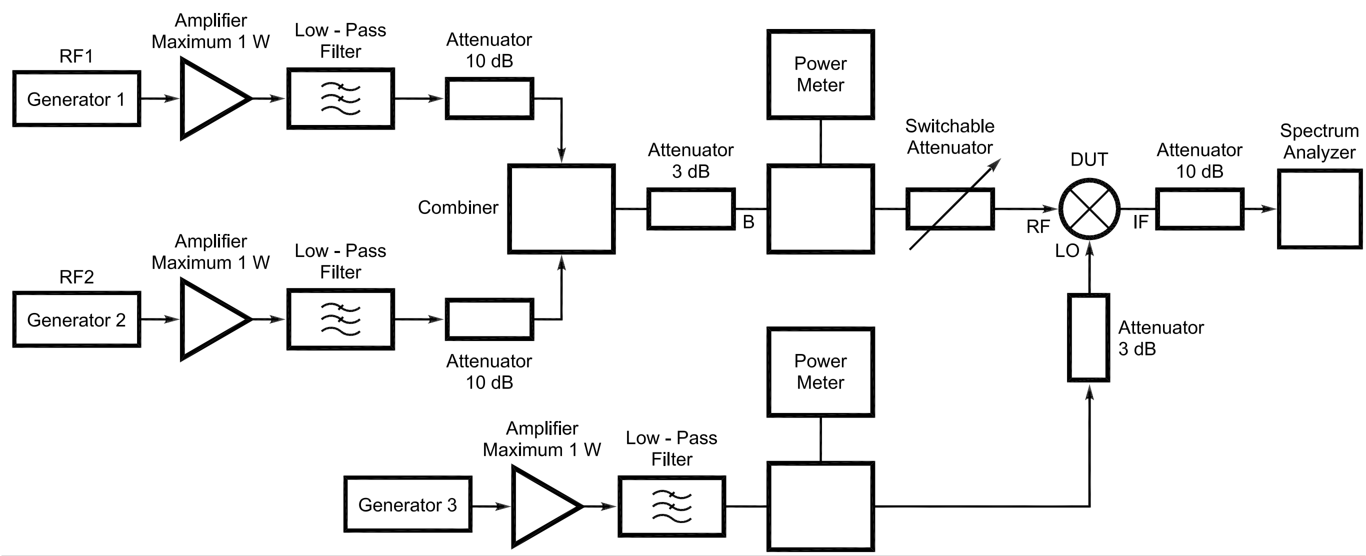


Fig 32—Test setup for high-IP_3 measurements.

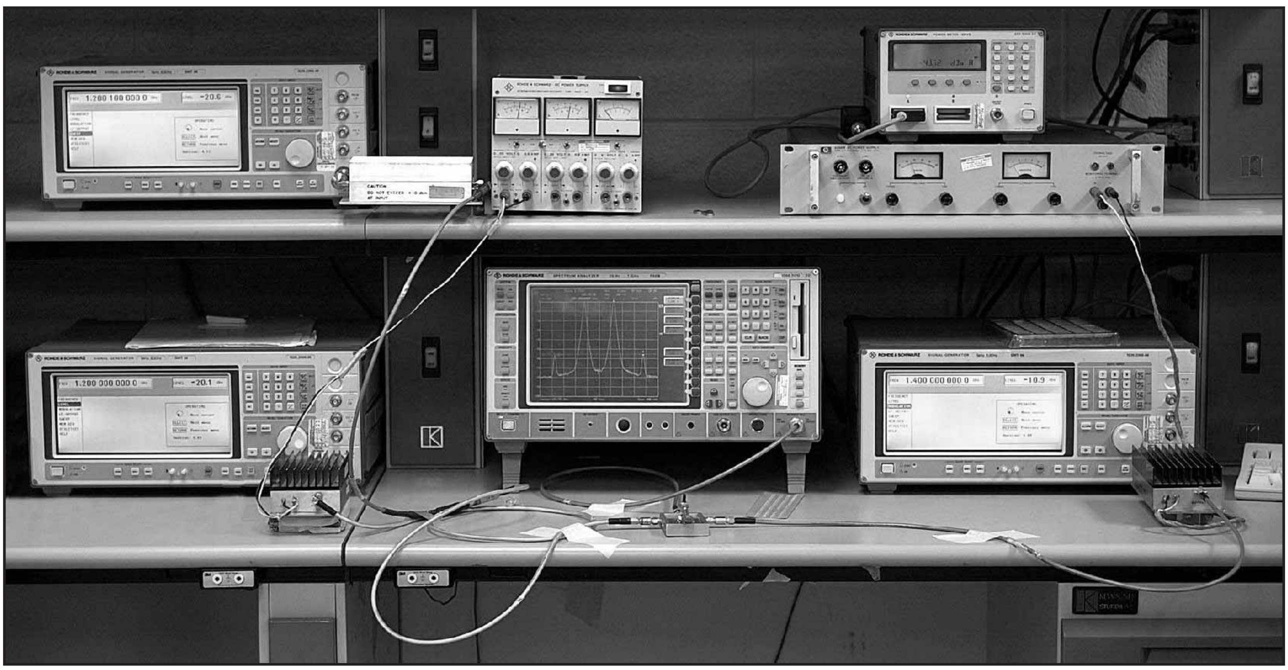


Fig 33—A photo of the setup without harmonic filters.

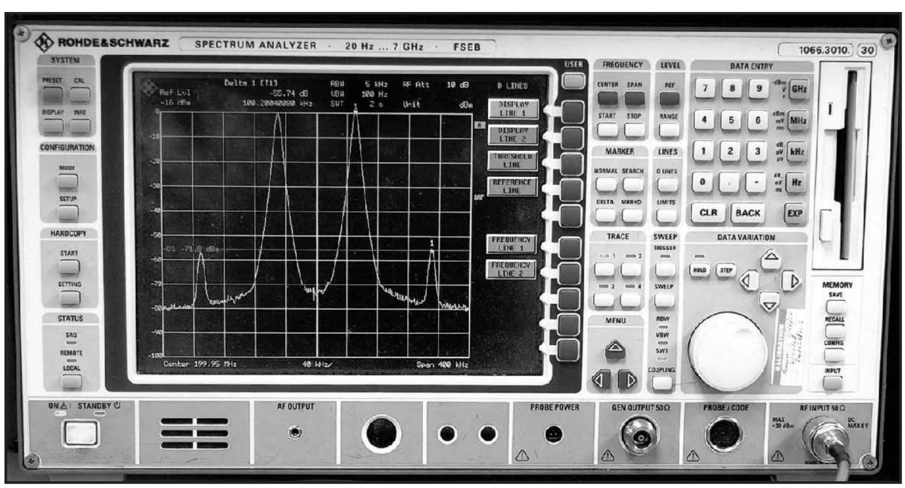


Fig 34—Measurements on the spectrum analyzer.

class-A amplifiers can be used as isolators. They have the advantage of providing the required power levels at the DUT. The isolation between both generators must be as great as possible for an expected IP_3. The IM_3 product at test point B must be at least 10 dB lower than the expected IP_3 product generated by the mixer. For example, to measure a mixer with an IP_3 of 35 dBm, the IM_3 product at point B must be lower than –90 dB for 0 dBm output. Such amplifiers have 20-dB gain, 1-W output power capability and 50-dB reverse isolation.

Fig 33 is a picture of a universal IMD-test setup. The picture shows two signal generators on the left (one on top of the other). They are connected to 1-W power amplifiers via a 10-dB attenuator to increase the isolation. These amplifiers have 20-dB gain and are capable of 1 W output. The device under test is shown at the bottom-center of the picture and is clamped down in a test fixture. The signal generator on the right feeds the input for the LO drive. The spectrum analyzer in the middle is a high performance Rohde & Schwartz FSEB spectrum analyzer operating from 20 Hz to 7 GHz. Its IF stages are DSP-based.

Fig 34 shows a close-in picture of the spectrum analyzer. The two input signals shown are at 0 dBm based on the attenuator setting. The symmetrical sidebands are roughly 56 dB down relative to 0-dB input. This calculates to IP_3 of +28 dBm.

We can reduce the harmonics of each generator by adding a low-pass filter after it. A better solution is to have two narrow band-pass filters shifted in frequency. In this case, both filters provide an additional isolation for the two generators. Practical measurements have shown that a 60-dB reduction of the harmonics has an influence on the IP_3 of about 4 dBm. To cover the complete RF range of the mixer, several low-pass filters with different corner frequencies are necessary.

To reduce the influence of the harmonics resulting from the LO amplifier, an additional low-pass filter after the LO amplifier is necessary. This can be demonstrated from an IP_3 measurement of a FET mixer with and without an additional low-pass filter. Differences up to 4 dB have been measured. For more details, see Fig 35.

For higher isolation, a special hybrid combiner can be used instead of a standard combiner. Some hybrids have a typical isolation of 35 dB. They

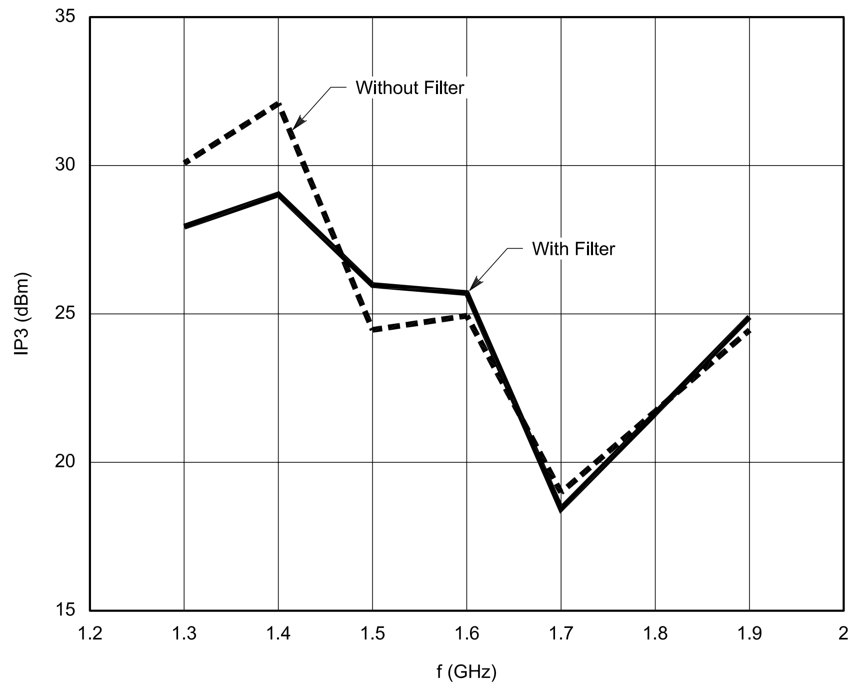


Fig 35—IP_3 measurements with and without a filter following the LO amplifier.

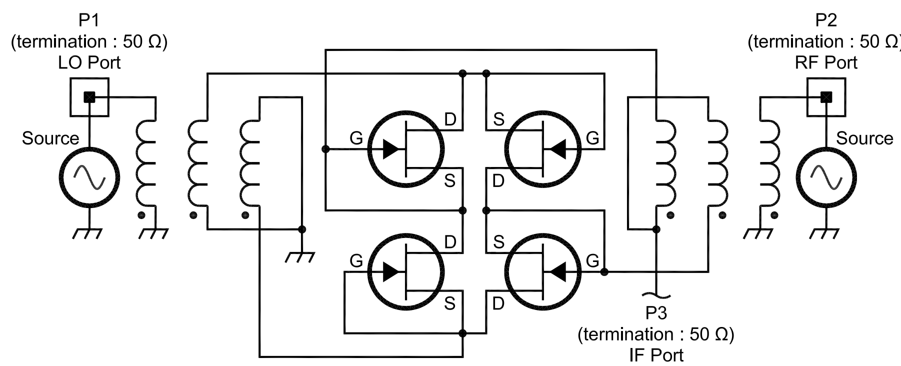


Fig 36—A doubly balanced mixer using GaAs FETs as mixer diodes, gate and source are connected (see References 1 and 2).

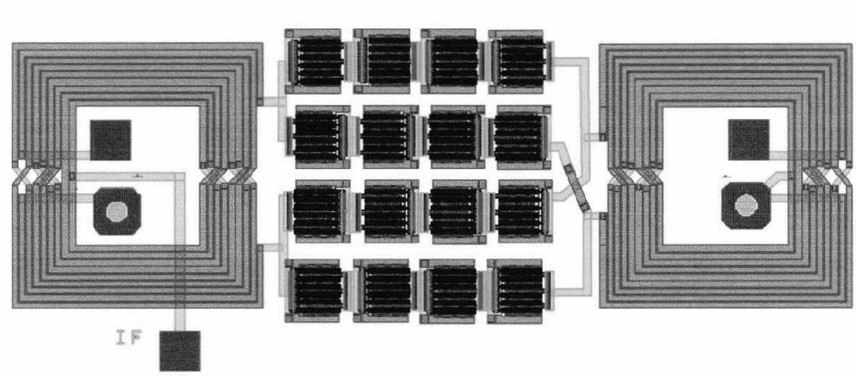


Fig 37—A view of the die for the circuit in Fig 36.

may be built with isolations up to 65 dB. A hybrid is recommended for narrow-band applications. The combiner or hybrid also needs a good load-impedance match at the common port. Therefore, an attenuator must be inserted between the combiner and RF port of the mixer.

Checking the Measured Result

To check the measured result of IP_3, a switchable attenuator is inserted at the input RF port of the mixer (see

Table 3—Comparison of IP_3 Measurements of a Diode Mixer and a FET Mixer

	IP_3 *measured at another facility*	IP_3 *measured by the Author*
Diode Mixer	31.5 dBm (w/o LO filter)	30 dBm (w/o LO filter) 33 dBm (with LO filter)
FET Mixer	34.5 dBm (w/o LO filter)	35.5 dBm (w/o LO filter) 34.5 dBm (with LO filter)

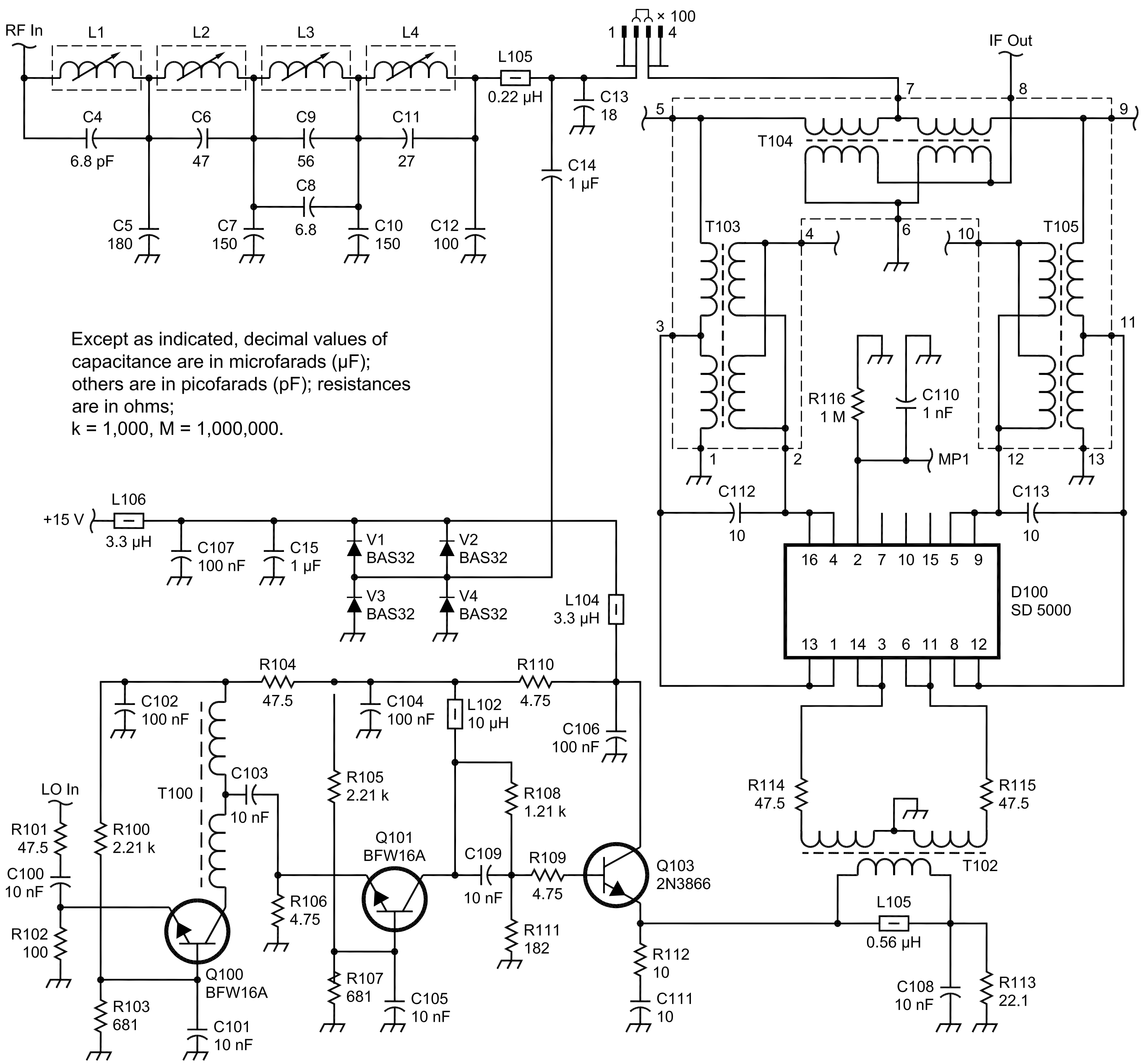


Fig 38—Circuit diagram of a high-performance receiver front end (EK890/895/896). It consists of a low-pass filter at around 33 MHz and a 40-dBm IP_3 switching-type mixer using the SD5000 quad switches. The three-stage amplifier underneath generates the high-voltage swing required for the mixer. To obtain high isolation, the first two amplifiers use a common-base configuration. This prevents any feedback from the mixer into the oscillator itself.

Fig 32). When the attenuation after the combiner is increased by 3 dB, the measured IM_3 product should decrease by 9 dB and the calculated IP_3 should be constant. If the IM_3 product falls less or more than 9 dB, optimize the test setup. Greater decoupling of both generators or greater harmonic suppression is necessary.

When you shift both frequencies *f1* and *f2*, IP_3 must be constant and symmetrical. If not, the diodes are not matched, or signal generators 1 and 2 do not have equal output levels.

With shifts of the frequency difference between *f1* and *f2*, IP_3 must be constant. Example: start with *f1* = 14.0 and *f2* = 14.1 MHz. The IMD products must remain at 14.2 and 13.9 MHz.

When changing the cable length, IP_3 must remain constant. It may happen that nonlinearity is present (cancellation of harmonics) within the test setup, which can result in much greater measured IP_3 than there really is.

To compare IP_3 test setups and measurements, the IP_3 of the diode mixer, as well as an FET mixer, were measured at another facility (see Table 3). Measurements were done at 1.8 GHz. Different high-level mixers available on the market were measured according to the above-described method.

As an example, a diode mixer was intended to have an IP_3 of 30 dBm; but the correct measurement resulted in an IP_3 of only 25 dBm. Another example was a FET mixer with an IP_3 of 38 dBm according to the datasheet. It was measured to have only 31 dBm in my test setup. The question arises: What kind of test setup did those companies use to evaluate their devices?

The mixer internally generates a large number of spurious products. This happens despite the mixer being doubly balanced. Manufacturers typically generate a table of such spurious products. Table 4 shows such a harmonic-spurious table of a doubly balanced mixer.

State-of-the-art mixers for high frequencies have used multiple diodes for high performance. A better or more modern way to use a GaAs FET-based diode ring is shown in Fig 36. Because the diode threshold level is now 1 V, as compared to 0.3 V for hot-carrier diodes, the local oscillator power required is 20 dB higher, and the intercept point is 20 dB higher than a conventional diode mixer ring. Such an array is shown in Fig 37. In low-frequency receivers, the use of silicon-based switching mixers has driven the IP_3 up to typically 36 dBm, and for well-matched cases, up to 42 dBm. A front end based on this doubly balanced mixer, such as the Rohde & Schwartz EK895, is shown in Fig 38. The circuit also gives information about the LO amplifier. An HF/Microwave version of this, called the Star Mixer, is shown in Fig 39. Fig 40 shows a picture of a production version. Because of the high LO drive required, a version with a preamplifier was built and is shown in Fig 41.

Impact of the Receiver Concept on IP_3 Performance

Recent publications, originating from England, have elaborated on dynamic-range requirements. It is true

Table 4—Harmonic Spurious of a Doubly Balanced Mixer

Harmonics		f_{LO}	$2f_{LO}$	$3f_{LO}$	$4f_{LO}$	$5f_{LO}$	$6f_{LO}$	$7f_{LO}$	$8f_{LO}$
$8f_{RF}$	100	100	100	100	100	100	100	100	100
$7f_{RF}$	100	97	102	95	100	100	100	90	100
$6f_{RF}$	100	92	97	95	100	100	95	100	100
$5f_{RF}$	90	84	86	72	92	70	95	70	92
$4f_{RF}$	90	84	97	86	97	90	100	90	92
$3f_{RF}$	75	63	66	72	72	58	86	58	80
$2f_{RF}$	70	72	72	70	82	62	75	75	100
f_{RF}	60	0	35	15	37	37	45	40	50
		60	60	70	72	72	62	70	70

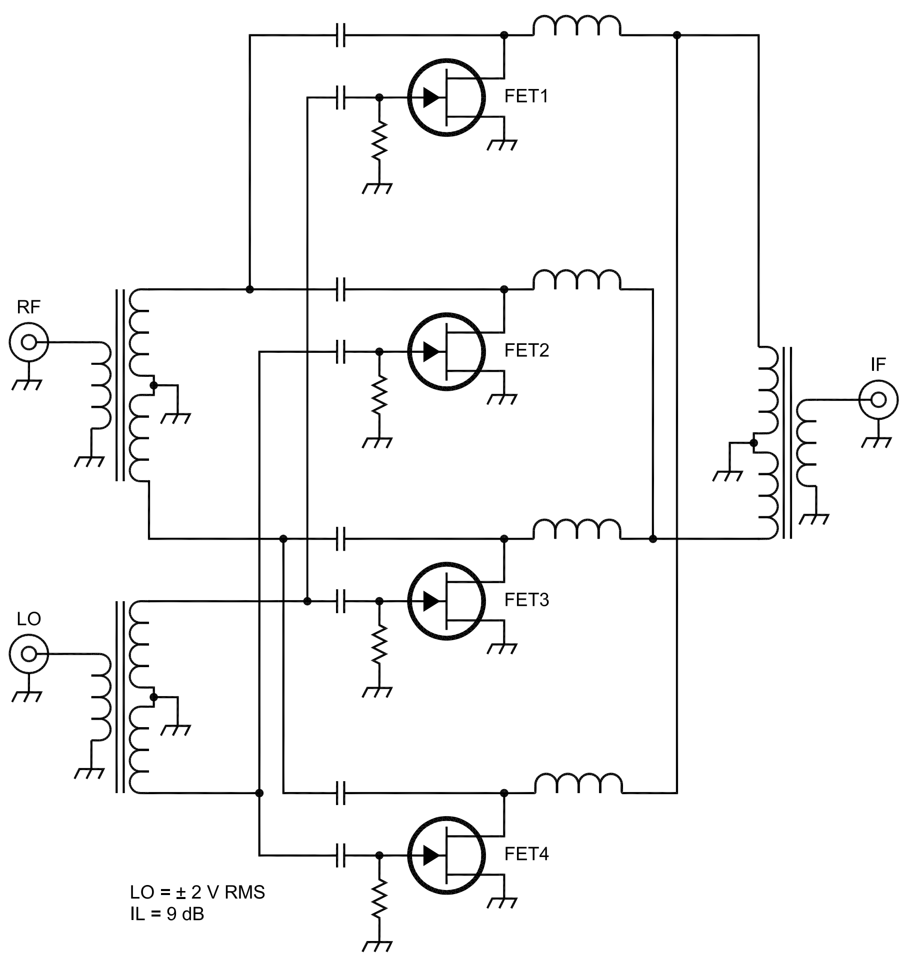


Fig 39—Star configuration, high-performance mixer having a 40-dBm intercept point. It uses GaAs FETs as switches. Depending on the input transformer, the frequency range can be adjusted. The particular one shown operates from 800-6000 MHz.

that about 10 years ago, the propaganda stations worldwide were congesting the airwaves; but things have changed. The real villains are the broadcast stations at 7.2 MHz, 9.6 MHz, 15.2 MHz, 17 MHz and 21.5 MHz. Precise measurements on a Rohde &Schwartz active antenna with constant electrical height show that the 20-, 17- and 15-meter broadcast stations adjacent to the ham bands can generate strong IMD products—specifically third-order. The second-order IMD products, such as 6 + 8 = 14 MHz, are more likely to be suppressed by input selectivity; however, the passband filter for the ham bands generally allows passing of the broadcast stations adjacent to ham bands quite well.

The traditional way around this problem is to build receivers with reasonable input selectivity, no preamplifier, but with high-level mixers. In any case, the preamplifier is an optional item and is switched in by diode switches; in most cases, they are neither PIN diodes nor relays. The best way to switch RF signals in the HF band is to use a common-gate FET switch, as shown in Fig 42.

Fig 43 shows a schematic of a modern receiver. The signal coming from the antenna is fed to a digitally controlled binary-coded attenuator with 60 dB of total range. The IP_3 of this attenuator exceeds 40 dBm and the IP_2 exceeds 80 dBm. It is followed by a low-pass filter set at 10% above the frequency of the highest frequency of reception and then followed by a band-pass filter; in this case, a 6.4-8 MHz filter. If the IP_3 of the low-pass and band-pass filters is above 40 dBm or 80 dBm for second-order products, it makes sense to put an RF attenuator after the second filter because the binary-coded GaAs switches produce some IMD products. An alternative, but much more expensive solution, is the use of a mechanical attenuator that does not add any IMD products. Such an attenuator has an IP_2 and IP_3 of infinity (>80 dBm for IP_3 and >120 dBm for IP_2), but is more expensive than its solid-state replacement.

Following the filters, there is an optional preamplifier with an intercept point of 20 dBm and 10 dB gain. Assuming the first mixer has an IP_3 of 30 dBm, this would reduce the IP_3 of the system down to 20 dBm; this is the same as the preamplifier. If it can be afforded, there is merit in designing a 30-dBm-intercept-point amplifier. This would reduce the practical intercept point from 40 dBm down to 30 dBm. This requires the previously shown high-level switching mixer.

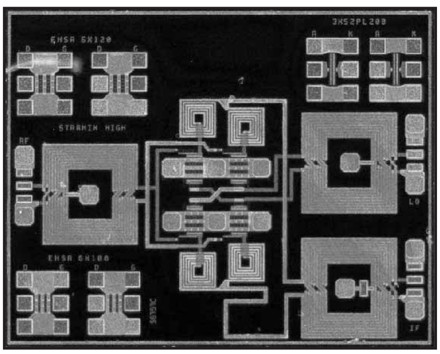


Fig 40— A view of the die for the Star Mixer shown in Fig 39. Notice the three printed inductors.

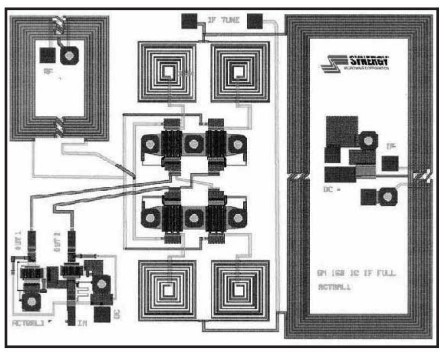


Fig 41—Layout of a proposed Star Mixer die with an LO driver amplifier at the lower left.

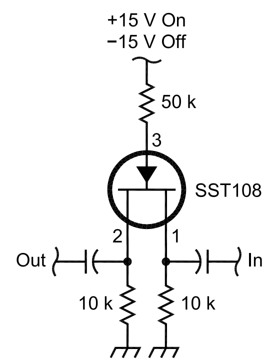


Fig 42—A single-FET switch with 0.1-dB insertion loss and an IP_3 of better than 40 dBm.

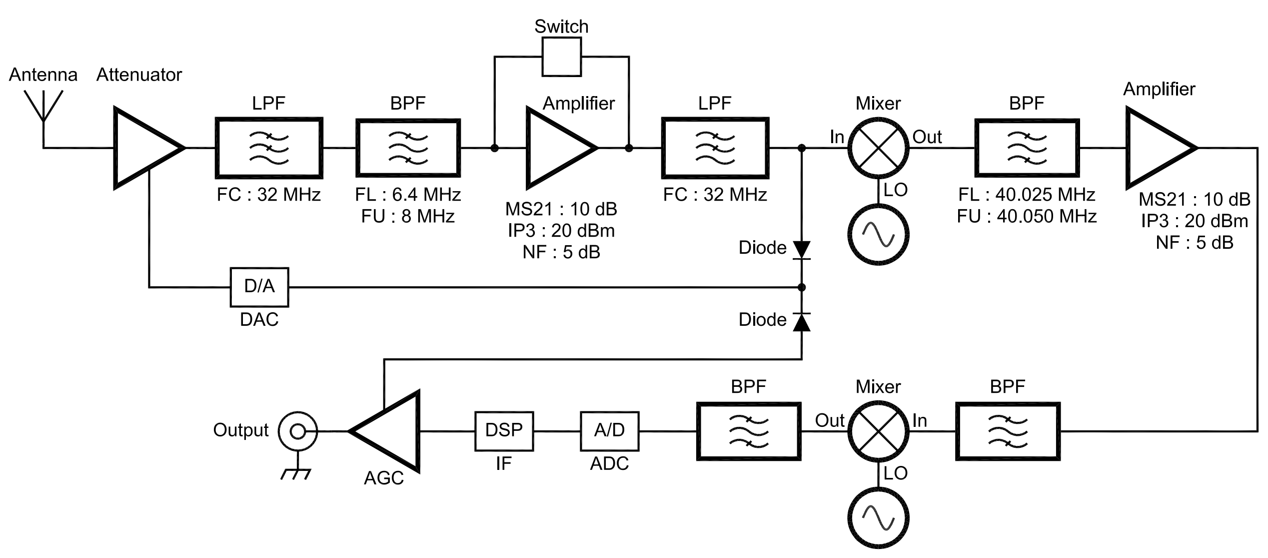


Fig 43—Block diagram of a high-dynamic-range receiver with two independent, interlaced AGC systems.

Dual-Loop AGC

The rest of the chain is conventional until we look at the AGC system. The receiver has two independent, yet connected, AGC loops. For signal levels up to 3 µV, the AGC system of the IF handles the first 30 dB of attenuation. The next 60 dB is a combination of IF and RF AGC; at approximately 1 mV and above, the contribution of the RF gain dominates. This can be seen in Fig 44. Because the AGC now makes heavy use of a pre-attenuator, which operates quasi-continuously, the intercept point now depends on the amount of pre-attenuation.

In the case of a DSP system, the in-band IMD is much less than the first- and second-mixer contributions. Given an intercept point of 20 dBm for the receiver, this is the case with the preamplifier on and listening to a signal of 100 µV based on the AGC in the RF loop. We already operate with 10 dB of RF attenuation. The trick of this method is to maintain a reasonable signal-to-noise ratio of 40 dB and

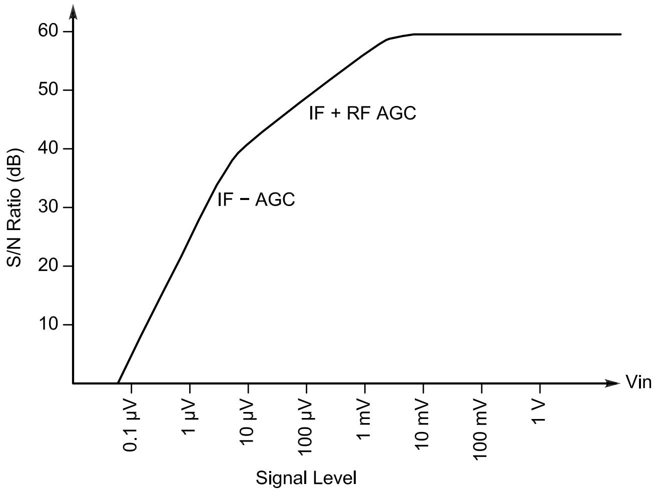


Fig 44—Signal-to-noise ratio (as a function of input voltage) plot for the receiver shown in Fig 43.

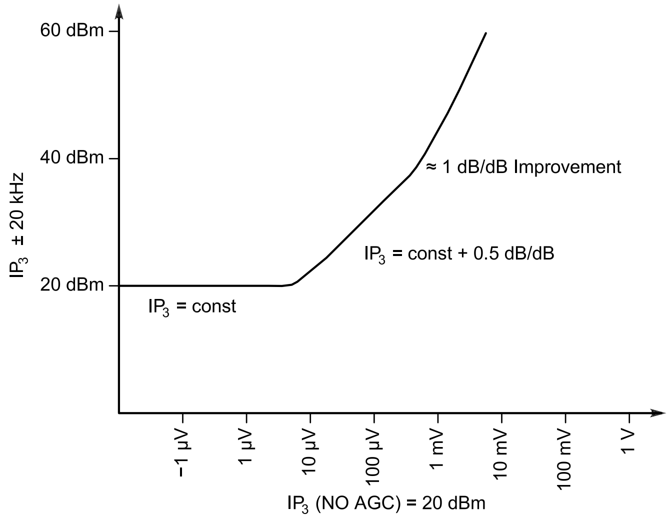


Fig 45—A plot of intercept-point behavior for the receiver system shown in Fig 43. The RF attenuation activated by the AGC voltage improves the third-order intercept point.

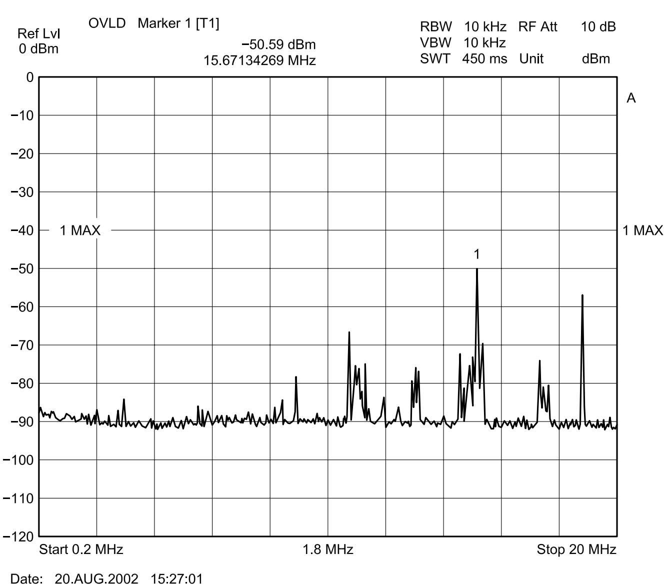


Fig 46—A spectrum-analyzer plot of signals from a calibrated Rohde & Schwartz active antenna covering 100 kHz to 100 MHz. The displayed frequency band is shown with a resolution bandwidth of 10 kHz. By reducing the resolution bandwidth to 1 kHz, 10 dB of more dynamic range can be obtained. The noise figure of the spectrum analyzer is roughly –100 dBm relative to a 1-kHz bandwidth.

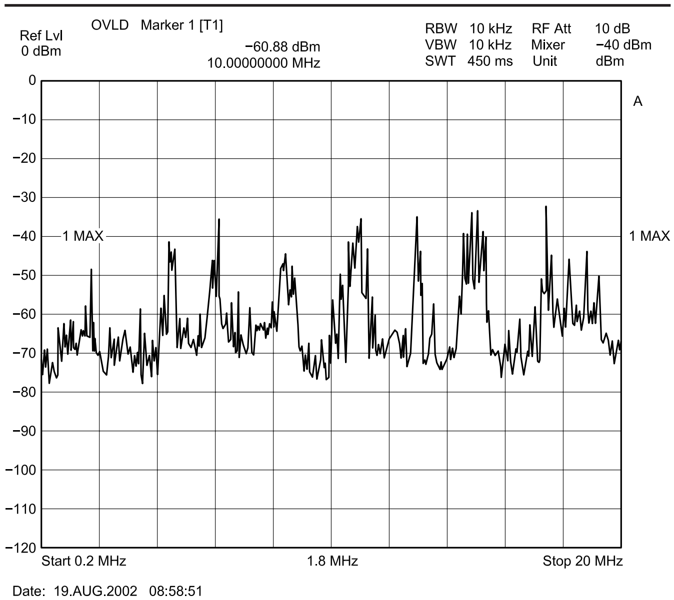


Fig 47—A 24-hour spectrum measurement covering the frequency range from 2 to 20 MHz. It is interesting to see that from 6 MHz to roughly 18 MHz we find a lot of busy bands. By reducing the resolution bandwidth to 1 kHz, the noise floor would drop down about 90 dBm and the actual shortwave dynamic range, meaning peak signal divided by noise floor, is 90 dBm – 35 dBm = 55 dBm. In this level range, a signal-to-noise ratio of the receiver of more than 55 dBm would be useless. This is why it was decided to limit the AGC signal-to-noise ratio to 60 dB, resulting in a sufficient safety margin.

use as much RF attenuation as possible. If an input signal would be about 1 mV, or 60 dB μV, equivalent to S9 + 26 dB, the intercept point of the receiver would increase over 40 dBm. Alternatively, the dual AGC system monitors the input to the first mixer. If the input at this point increases to 1 mV for any given interference outside the IF bandwidth, the pre-attenuator will become active. These conditions can occur at night on 40 meters. In this case, the noise level on 40 meters is equivalent to 10 μV into 50 Ω. Since the receiver sensitivity is 0.3 μV for a 10-dB signal-to-noise ratio (as seen in Fig 45) the receiver can afford at least 20 dB preattenuation, in which case the intercept point with the amplifier will increase to 40 dBm. Without the preamplifier on, it will increase to 60 dBm. At those levels, the input filters are likely to be the dominant source of intermodulation, unless good precautions have been taken.

In the past, many authors—including myself—have done signal evaluation on the shortwave bands, including ham bands. Previous spectrum analyzers, such as the HP-141, did not have low-phase-noise oscillators to really evaluate the spectrum. In addition, the shape factor of IF filters also was not sufficient to provide the necessary resolution. Only the modern (year 2000) spectrum analyzers built in extremely low-phase-noise, fractional-N synthesizers and DSP-based IF stages that provided enough resolution. The useful dynamic range of the spectrum analyzer needs to be more than 100 dB with a safety margin of 120 dB. Figs 46 and 47 show peak average measurements done with an active antenna and an appropriate spectrum analyzer. It is amazing how many holes are found between the stations, which the average receiver would not indicate.

Example: There have been discussions about measurements in general and the League's measurements in particular. For a point of reference, I have decided to do a set of measurements, in parallel with the League, and revisit one of my modified ham transceivers, the FT-890 by Yaesu. The test setup is the same as shown previously, and the first test object was the Rohde & Schwartz XK2100L and the EK895. Both receivers have the same front end but different IF combinations. The transceiver XK2100 is a dual-conversion system from a 45-MHz first IF to a 25-kHz second IF. The EK895 has three IFs; the middle one being 1.44 MHz. The reason for this is that the IF was once analog and was replaced by a DSP system. The Yaesu FT-890, one of my favorite inexpensive radios, was modified by replacing all the input filter switching diodes with MI204 PIN diodes. First, the system is calibrated for linearity.

The XK/EK system was measured with two tones set at 0 dBm (1 mW) at the antenna terminal and the two-tone IMD products at 82 dB down, or at –82 dBm, using the formula

$$IP_3 = P_{IN} + \frac{P_{OUT} - IM_3 \text{ products}}{2} = 41 \qquad \text{(Eq 25)}$$

where, IM_3 products is in decibels and P_{IN}, P_{OUT} and IP_3 are in dBm.

Both the XK and the EK systems were measured this way to have an intercept point of 41 dBm. The intercept point remains constant over the HF band (1.5-30 MHz) with a tolerance of ±1 dB. The actual specification for this was better than 36 dBm. These values are about the best on the market. When measuring at significantly lower levels, the IMD products of the test system appear to have an intercept point of +26 dBm. It is, therefore, essential to use an appropriate high level signal for the test.

The FT-890 uses a push/pull junction-FET mixer, and such a high intercept point cannot be expected. With the preamplifier on, two tones of –20 dBm, exactly, were provided to the receiver. The output from the hybrid coupler was set at 0 dBm followed by a 20-dB switchable attenuator. Two sets of measure-

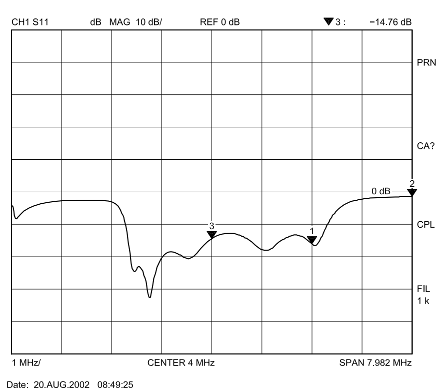


Fig 48—Input return loss of the Yaesu FT-890 for the 80-meter band.

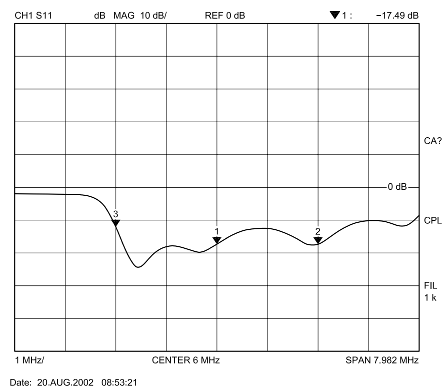


Fig 49—Input return loss of the Yaesu FT-890 for the 40-meter band.

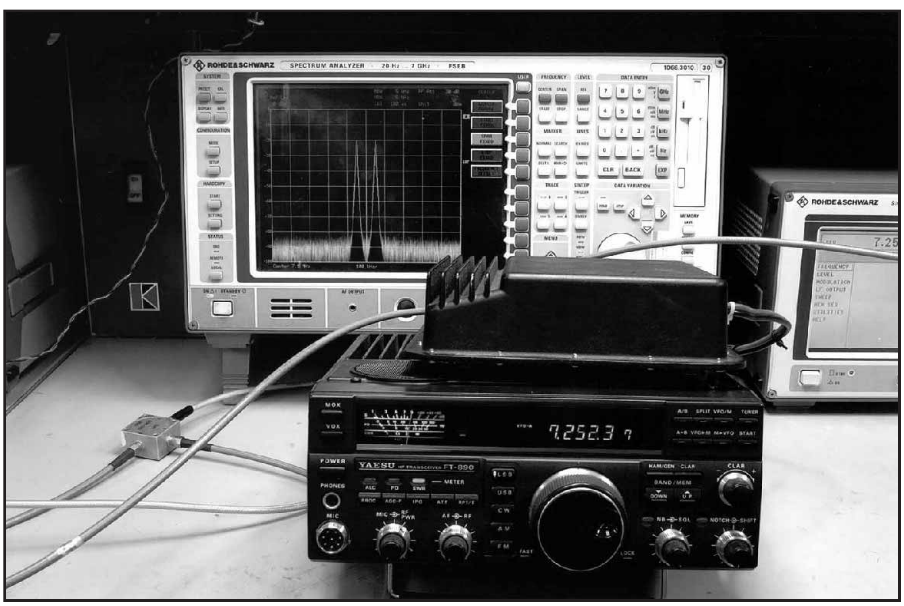


Fig 50—The test setup for the Yaesu FT-890. The S-meter indicates such spurious products as described.

ments were done: one at 7.15 MHz/ 7.05 MHz and one at 14.15 MHz/ 14.05 MHz. With the preamplifier on, the spurious product—100 kHz higher or 7.25 MHz or 14.25 MHz—appears at a level of –84 dBm or approximately 14 µV. Using the above-stated equation, we obtained ((84 – 20) ÷ 2) – 20 = 32 – 20 = 12 dBm. This with the amplifier on! The corresponding values, without a preamplifier, were 2 × –6 dBm and the products at –70 dBm or 71 µV. The resulting number is ((70 – 6) ÷ 2) – 6 = 32 – 6 = 26 dBm. For comparison, here is data for the IC-746 at 25-kHz and 100-kHz spacing. The IP_3 with preamplifier on is ((70 – 22) ÷ 2) – 22 = 24 – 22 = 2 dBm. IP_3 with the preamplifier off is ((70 – 10) ÷ 2) – 10 = 60/2 – 10 = 30 – 10 = 20 dBm. IP_3 products were –70 dB down. Input power was –22 dBm and –10 dBm for the two cases.

Most companies in the ham business during recent years have followed my recommendation and incorporated PIN diodes prior to the mixer. The values for the input level at –6 dBm and –20 dBm are levels that Amateur Radio transceivers should be able to accept.

Having a step attenuator in front of the system allows us to reduce the level by some decibels, like 3 dB, in which case the IMD products need to go down by 9 dB. This is a necessary test to make sure that the measurements are in the linear region of the receiver.

To actually do those measurements, the S-meter reading was calibrated after seeing the IMD products at slightly more than S9. The limitation of the test setup can be dependent on the return losses of the receiver input. Fig 48 shows the input return losses at 4 MHz to be about 14.76 dB and Fig 49 shows the return loss at 6 MHz to be +17.49 dB. By definition, the maximum isolation for the hybrid coupler is insertion loss minus return loss. In our case at 40 meters, that is 6 + 17.45 dB = 23.45 dB. Fig 50 shows the actual test setup with the Yaesu FT-890.

The measured 10-dB signal-to-noise ratio with a 2.4 kHz bandwidth was –130 dBm with preamplifier on and –115 dBm with preamplifier off. The conventional definition of dynamic range would not be –30 dBm + 115 or 85 dB in 2.4 MHz bandwidth, while with the preamplifier on it is –22 dBm + 130 or 108 dB in 2.4 MHz bandwidth.

In my opinion, however, the conventional definitions are incorrect. One really should take a three-way power divider and add a third channel that is 3 kHz away from the IMD products. Such strong carriers generate some blocking problems from reciprocal mixing and the noise floor will now be 3 dB or higher. (In accordance with international conventions, the blocking and reciprocal mixing are the same effect only different expressions are used in different countries.) The 10-dB signal-to-noise ratio should be measured again 3 kHz away from the IMD products, and the actual large-signal 10-dB signal-to-noise ratio is found. This is the correct number when listening to stations that are strong enough to cause reciprocal-mixing noise. The current literature does not consider this effect.

Conclusion

In this paper, I have tried to give a comprehensive overview of the theory of nonlinearities—specifically, IMD products—and show practical requirements for a high-performance test setup. Several examples for mixers were given and IMD measurements were done on a set of military high-performance radios, as well as a popular Amateur Radio transceiver.

References

1. U. L. Rohde and D. P. Newkirk, *RF/Microwave Circuit Design for Wireless Applications* (New York: John Wiley & Sons, Apr 2000, ISBN 0-471-29818-2), pp 93-95.
2. U. L. Rohde and J. Whitaker, *Communications Receivers: DSP, Software Radios, and Design,* 3rd Edition, (New York: McGraw-Hill, Dec 2000, ISBN 0-071-36121-9).
3. D. Smith, KF6DX, "Improved Dynamic-Range Testing," *QEX*, Jul/Aug 2002, pp 46-52.
4. U. L. Rohde, "How Many Signals Does a Receiver See?" *ham radio* magazine, Jul 1977, p 58.
5. P. E. Chadwick, G3RZP, "HF Receiver Dynamic Range: How Much Do We Need?" *QEX*, May/Jun 2002, pp 36-41.
6. U. Graf, DK45X, "Empfänger-Intermodulation (1)—Theorie und Praxis," *CQ DL*, June 2002 (Part 1, Vol 6; Part 2, Vol 7; Part 3, Vol 8).
7. U. L. Rohde, "Testing and Calculating Intermodulation Distortion in Receivers," *QEX*, Jul/Aug 1994, pp 3-4.
8. U. L. Rohde, "Recent Advances in Shortwave Receiver Design," *QST*, Nov 1992, pp 45-54.
9. U. L. Rohde, "Key Components of Modern Receiver Design, Part 1," *QST*, May 1994, pp 29-32; Part 2, Jun pp 27-31; Part 3, Jul pp 42-45.
10. U. L. Rohde, "Key Components of Modern Receiver Design: A Second Look," *QST*, Dec 1994, pp 38-42, 44.
11. U. L. Rohde, "Requirements For High Performance RF/UHF ICs And Possible Solutions," Keynote Address - RFIC Symposium, MTT, June 2002.

A picture is worth a thousand words...

With the all-new

ANTENNA MODEL™

wire antenna analysis program for Windows you get true 3D far field patterns that are far more informative than conventional 2D patterns or wire-frame pseudo-3D patterns.

Describe the antenna to the program in an easy-to-use spreadsheet-style format, and then with one mouse-click the program shows you the antenna pattern, front/back ratio, front/rear ratio, input impedance, efficiency, SWR, and more.

An optional Symbols window with formula evaluation capability can do your computations for you. A Match Wizard designs Gamma, T, or Hairpin matches for Yagi antennas. A Clamp Wizard calculates the equivalent diameter of Yagi element clamps. Yagi Optimization finds Yagi dimensions that satisfy performance objectives you specify. Major antenna properties can be graphed as a function of frequency.

There is no built-in segment limit. Your models can be as large and complicated as your system permits.

ANTENNA MODEL is only $85US. This includes a Web site download and a permanent backup copy on CD-ROM. Visit our Web site for more information about ANTENNA MODEL.

Teri Software
P.O. Box 277
Lincoln, TX 78948

www.antennamodel.com

e-mail sales@antennamodel.com
phone 979-542-7952

Understanding Switching Power Supplies, Part 2

Come learn design techniques for the real world.

By Ray Mack, WD5IFS

In the first part of this article,[1] we looked at basic circuit configurations and operation. In this installment, we will look at designing power supplies with real-world transformers and coils. We will look in detail at the datasheet for ferrite material 77 for transformers and at various powdered-iron materials for smoothing chokes. We will look at material 77 from Fair-Rite Corporation because Amidon Inc carries transformer cores in this mix, and they have long supplied the Amateur Radio community with reasonable ordering requirements. Amidon also carries other mixes, but mix 77 was most available when I checked their inventory. We will also look at the relative merits of materials from other suppliers.

The examples in this installment will have requirements that appear as if by magic. The information on how to decide turns ratios, inductance values, ripple current, etc will come in a later installment.

Properties of Real Magnetics

Our discussion of basic circuits assumed ideal inductors and transformers. Real inductors have the properties of inductance, resistance and capacitance plus losses due to the core material. The inductor section of the *ARRL Handbook* gives a good description of the losses in magnetic cores. The primary properties we must consider when choosing a core are permeability, magnetic losses (hysteresis), maximum flux density (saturation) and core temperature.

Ferrite is a ceramic composed of magnetic metal oxides mixed with iron oxide. The most popular magnetic materials are manganese and zinc or nickel and zinc. The magnetic materials are mixed with an organic binder and fired in a kiln to make a ceramic in the same manner that ceramic dishes are made. Since they are ceramics, ferrites can be produced in many shapes simply by varying the mold shape. They can also be machined after firing to produce smooth surfaces and precise dimensions. Ferrite cores for power use are typically made of manganese-zinc compounds for higher permeability. Eddy-current loss in ferrite is quite low in the normal operating range of the material because of the insulating property of the

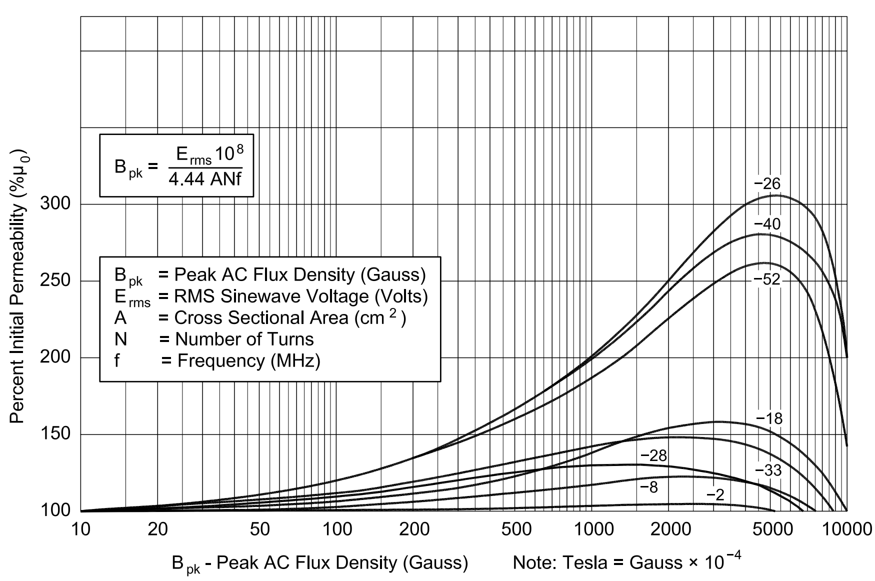


Fig 1—Change in initial permeability versus ac flux density in powdered-iron cores.

[1] R. Mack, "Understanding Switching Power Supplies, Part 1," *QEX*, Sept/Oct 2002, pp 30-35.

17060 Conway Springs Ct
Austin, TX 78717-2989
wd5ifs@arrl.org

oxides and the binder. The resistance of the material is quite high compared to metals. Eddy-current loss increases as frequency increases.

Powder cores are made by grinding iron or other alloys into fine particles and coating them with an insulating material. Then they are die-pressed and baked. Toroids or rods are common shapes, but some manufacturers can produce other shapes, like E cores. The size of the powder limits the upper frequency limit of powdered cores because of eddy-current losses. The Amidon RF toroids we are familiar with are all powder cores.

Hysteresis (ac flux density) losses are the magnetic equivalent of dielectric losses. Both losses result from the interaction between the electrons of the material and the external field. The electrons absorb part of the field as field intensity increases, but they release part of the energy as heat, rather than returning the energy to the field. As frequency or ac flux density rise, the associated losses in a magnetic material also rise. Flux density losses can be determined by examining the area inside the B-H curve for the operating conditions. Hysteresis losses limit maximum flux density for frequencies above 20 kHz. Below 20 kHz, the maximum flux density is usually limited by the saturation flux density.

The material of choice for transformer cores in modern switching supplies is a high frequency ferrite. Choke cores can be either powdered iron or ferrite.

Each of the manufacturers has a Web site with large amounts of useful data on their products and in-depth application notes. You will find a list of Web sites at the end of this article. Each manufacturer provides a set of nomographs, figures and procedures for deciding which core to use. Unfortunately, the procedure is different from one manufacturer to another. The procedures presented later are useful with the bare minimum of information about a set of cores. If you are going to do a rigorous design, it is best to get the full application information from the manufacturer you intend to use, so the design process is streamlined.

Many catalogs and application notes refer to "soft ferrites." This has nothing to do with the mechanical hardness of the material, but instead refers to the magnetic B-H curve of the material. Soft ferrites have a rather low value of residual magnetism when the magnetizing force is removed, whereas hard ferrites have residual magnetism that is nearly identical to the saturation flux density. In addition, soft ferrites have low coercive-force values. That means it takes very little magnetic force to reduce the residual magnetism to zero. Ceramic permanent magnets are "hard ferrites."

Choosing a Choke Core

Forward converters and buck regulators require a choke inductor to smooth the rectangular pulses to a dc value. The smoothing choke in a high-current supply will have a large dc bias current flowing through the windings, which creates a very large magnetic field in the core. An air gap is necessary in the core to ensure that the core does not saturate. If we choose to use a ferrite core, it must have a gap when the core pieces are assembled. This gap allows flux leakage and can be a problem for interference and control of the effective permeability as the gap gets large. A powder core has an inherent distributed gap caused by the insulating material around the magnetic particles. Most designers prefer to use powdered core toroids for smoothing chokes because of cost, control of permeability and ease of assembly. The limiting factors for choke cores are saturation and temperature rise.

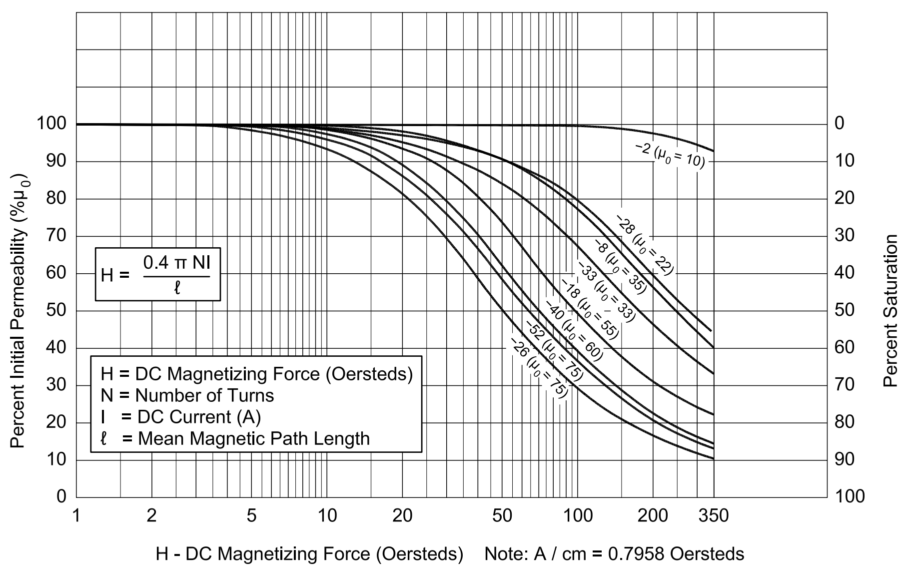


Fig 2—Decrease of permeability due to dc bias in powdered-iron cores.

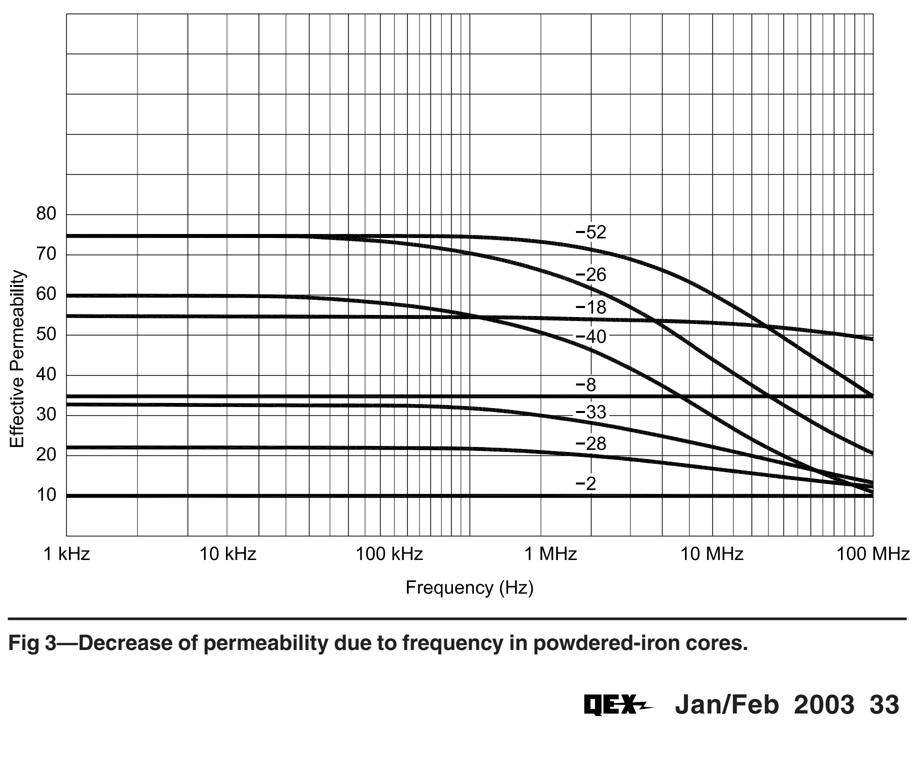


Fig 3—Decrease of permeability due to frequency in powdered-iron cores.

The permeability of magnetic materials is not constant. It varies as a function of ac flux density, frequency, dc bias current and temperature. Fig 1 gives a graph of the change in initial permeability for various levels of ac flux in various Micrometals powdered core materials. The dc bias current that is present in continuous mode operation decreases the permeability of the material as shown in Fig 2. Fig 3 shows the change in permeability versus frequency. Micrometals powder cores show a minor, linear change of permeability with temperature. Mix-26 has a temperature coefficient of 825 ppm/C° and -52 has 525 ppm/C°. Fig 4 shows the B-H curves for four Micrometals materials including -26 and -52.

Mix -26 (yellow with white) and -52 (green with blue) are the least expensive of the powdered core materials. They also exhibit the largest change in permeability due to environmental factors. Mix -52 is recommended for frequencies above 100 kHz and mix -26 for frequencies below 100 kHz. These two materials are the most likely ones to find in computer power supplies when scrounging for parts. If your design requires a large output current range (such as a SSB transmitter supply) and constant inductance is required, then mixes -18 and -8 are the preferred materials. These have much more stable characteristics with changes in flux and frequency. The trade-off is that permeability is lower (55 and 35) and cost is approximately doubled for identically sized cores. A larger core is also required, so the final cost of the core can increase by four times or more. Unfortunately, neither Amidon nor Micrometals has all of their application information on the Web. Micrometals will provide their application catalog on request. It is very useful if you need to pick the optimum core. I have included the bare minimum of data from Micrometals to allow you to follow the examples.

As long as the ripple current is

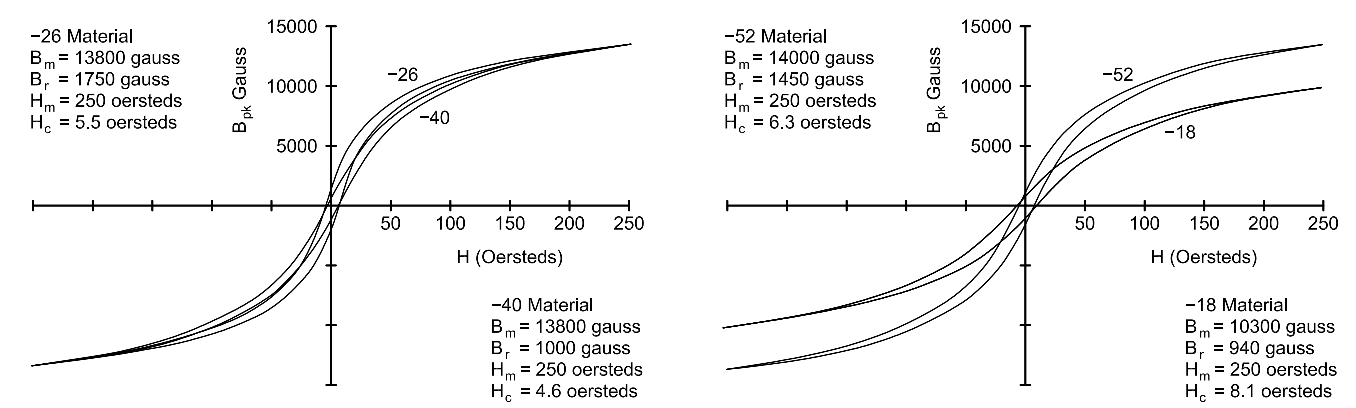


Fig 4—B-H characteristics of powdered-iron cores.

Table 1—Wire Table

Wire Size (AWG)	*Resistance Ω/meter**	*Wire OD (cm) (Heavy Build)*	***Wire Area*** *Circular Mils*	*cm²* *(×0.001)*	***Current Capacity, A (listed by columns of A/cm²)*** *200*	*400*	*600*	*800*
8	0.00207	0.334	18,000	91.2	16.5	33.0	49.5	66.0
9	0.00259	0.298	14,350	72.7	13.1	26.2	39.3	52.4
10	0.00328	0.267	11,500	58.2	10.4	20.8	31.2	41.6
11	0.00413	0.238	9160	46.4	8.23	16.4	24.6	32.8
12	0.00522	0.213	7310	37.0	6.53	13.1	19.6	26.1
13	0.00656	0.1902	5850	29.6	5.18	10.4	15.5	20.8
14	0.00827	0.1714	4680	23.7	4.11	8.22	12.3	16.4
15	0.01043	0.1529	3760	19.1	3.26	6.52	9.78	13.0
16	0.01319	0.1369	3000	15.2	2.58	5.16	7.74	10.3
17	0.01657	0.1224	2420	12.2	2.05	4.10	6.15	8.20
18	0.0210	0.1095	1940	9.83	1.62	3.25	4.88	6.50
19	0.0264	0.0980	1560	7.91	1.29	2.58	3.87	5.16
20	0.0332	0.0879	1250	6.34	1.02	2.05	3.08	4.10
21	0.0420	0.0785	1000	5.07	0.812	1.63	2.44	3.25
22	0.0531	0.0701	810	4.11	0.640	1.28	1.92	2.56
23	0.0666	0.0632	650	3.29	0.511	1.02	1.53	2.04
24	0.0843	0.0566	525	2.66	0.404	0.808	1.21	1.62
25	0.1063	0.0505	425	2.15	0.320	0.641	0.962	1.28
26	0.1345	0.0452	340	1.72	0.253	0.506	0.759	1.01
27	0.1686	0.0409	270	1.37	0.202	0.403	0.604	0.806

*Multiply table value by 0.305 for Ω/ft

approximately 1% of the total dc current, the only temperature rise will be due to copper losses. The temperature rise will be greater when ripple current rises because of hysteresis losses in the core. You must also consider the increase in inductance that will occur as permeability increases when ripple current increases to 10% or 25% of total current.

Table 1 is a wire table that describes the characteristics of various wire gauges. The columns of interest are the resistance/meter and the current at various current densities. A current density of 600 A/cm^2 corresponds to roughly 40°C rise and 400 A/cm^2 corresponds to roughly 20°C rise. Table 2 indicates the maximum number of turns that will fit in a single layer on various Amidon toroids. Table 3 lists the magnetic path length for various core sizes. We also need to account for the change in initial permeability due to dc flux. The formula for percent change in permeability for Micrometals materials has the form:

$$\%\mu = \left(\frac{A + c\mathrm{H} + e\mathrm{H}^2}{1 + \mathrm{bH} + \mathrm{dH}^2}\right)^{\frac{1}{2}} \qquad \text{(Eq 1)}$$

Table 4 lists the values of these coefficients for the materials of interest.

Let's do an example of a choke design for a forward converter. The design requires an inductance of 15 µH and has a maximum dc current of 20 A with 200 mA maximum ripple. We choose to limit the temperature rise to 40°C so we can use the current from the 600 A/cm^2 column. We assume we found a T106-26 core in an old PC power supply. All parameters not found in tables here can be found in the Component Data chapter of the *ARRL Handbook*. A_l for this core is 900 µH/100 turns. First, we calculate the number of turns required:

$$N = 100\left(\frac{L}{A_l}\right)^{\frac{1}{2}}$$

$$= 100\left(\frac{15}{900}\right)^{\frac{1}{2}} \qquad \text{(Eq 1A)}$$

$$= 100(0.0167)^{\frac{1}{2}} = 13 \text{ turns}$$

From Table 2 we see that 13 turns of AWG #12 will fit this core. From Table 1 we see that #12 wire will allow 19.6 A. This is close enough to the design goal of 20 A. Now we can verify that the magnetic parameters meet

Table 2—Turns versus Wire Size and Core Size

Powdered-Iron Toroidal Cores: Dimensions

Red E Cores—500 kHz to 30 MHz (µ = 10)

No.	*OD (in)*	*ID (in)*	*H (in)*
T-200-2	2.00	1.25	0.55
T-94-2	0.94	0.56	0.31
T-80-2	0.80	0.50	0.25
T-68-2	0.68	0.37	0.19
T-50-2	0.50	0.30	0.19
T-37-2	0.37	0.21	0.12
T-25-2	0.25	0.12	0.09
T-12-2	0.125	0.06	0.05

Black W Cores—30 MHz to 200 MHz (µ=6)

No.	*OD (In)*	*ID (In)*	*H (In)*
T-50-10	0.50	0.30	0.19
T-37-10	0.37	0.21	0.12
T-25-10	0.25	0.12	0.09
T-12-10	0.125	0.06	0.05

Yellow SF Cores—10 MHz to 90 MHz (µ=8)

No.	*OD (In)*	*ID (In)*	*H (In)*
T-94-6	0.94	0.56	0.31
T-80-6	0.80	0.50	0.25
T-68-6	0.68	0.37	0.19
T-50-6	0.50	0.30	0.19
T-26-6	0.25	0.12	0.09
T-12-6	0.125	0.06	0.05

Number of Turns vs Wire Size and Core Size

Approximate maximum number of turns—single layer wound—enameled wire.

Wire Size	*T-200*	*T-130*	*T-106*	*T-94*	*T-80*	*T-68*	*T-50*	*T-37*	*T-25*	*T-12*
10	33	20	12	12	10	6	4	1		
12	43	25	16	16	14	9	6	3		
14	54	32	21	21	18	13	8	5	1	
16	69	41	28	28	24	17	13	7	2	
18	88	53	37	37	32	23	18	10	4	1
20	111	67	47	47	41	29	23	14	6	1
22	140	86	60	60	53	38	30	19	9	2
24	177	109	77	77	67	49	39	25	13	4
26	223	137	97	97	85	63	50	33	17	7
28	281	173	123	123	108	80	64	42	23	9
30	355	217	154	154	136	101	81	54	29	13
32	439	272	194	194	171	127	103	68	38	17
34	557	346	247	247	218	162	132	88	49	23
36	683	424	304	304	268	199	162	108	62	30
38	875	544	389	389	344	256	209	140	80	39
40	1103	687	492	492	434	324	264	178	102	51

Actual number of turns may differ from above figures according to winding techniques, especially when using the larger size wires. Chart prepared by Michel J. Gordon, Jr., WB9FHC
Courtesy of Amidon Assoc.

Table 3—Magnetic Path Length of Cores

Core	T16	T20	T25	T37	T50	T68	T80	T94	T106	T130	T200
Path Length (cm)	0.93	1.15	1.50	2.31	3.19	4.23	5.14	5.97	6.49	8.28	13.0

the maximum for the core. Eq 2 gives the magnetizing force applied to a core.

$$H = \frac{(0.4\pi N I)}{l} \qquad \text{(Eq 2)}$$
$$= \frac{(0.4\pi 13 \bullet 20)}{6.49} = 50.3 \text{ Oersteds}$$

We verify that 50 Oersteds is below the saturation point using Fig 4. We can calculate the change in permeability due to dc bias from Eq 1:

$$\%\mu = \left(\frac{10090 + 13.1H + 0.0212H^2}{1 + 0.00505H + 0.00117H^2}\right)^{\frac{1}{2}}$$
$$= \left(\frac{10090 + 658.9 + 53.6}{1 + 0.259 + 2.96}\right)^{\frac{1}{2}}$$
$$= 51\%$$

Now we can adjust the number of turns based on the reduced permeability:

$$N = 100\left(\frac{L}{(A_l \times 0.51)}\right)^{\frac{1}{2}} \qquad \text{(Eq 2B)}$$
$$= 100\left(\frac{15}{(900 \times 0.51)}\right)^{\frac{1}{2}}$$
$$= 18 \text{ turns}$$

We cannot fit 18 turns of #12 on this core in a single layer, so we have several options. The first is to use the next larger core (T130), so we can fit the proper number of turns of #12. The next option is to put an additional two turns of #12 over the single layer. This will cause a slightly higher temperature rise but we probably have enough margin to do so. The next option is to adjust the inductance value so that we can fit enough wire. In this case we can fit 16 turns of #12, which yields an inductance of 11.8 µH. The last option is to allow greater temperature rise in the inductor and use #14 wire.

Choosing a Boost Converter Core

Ripple current is usually much

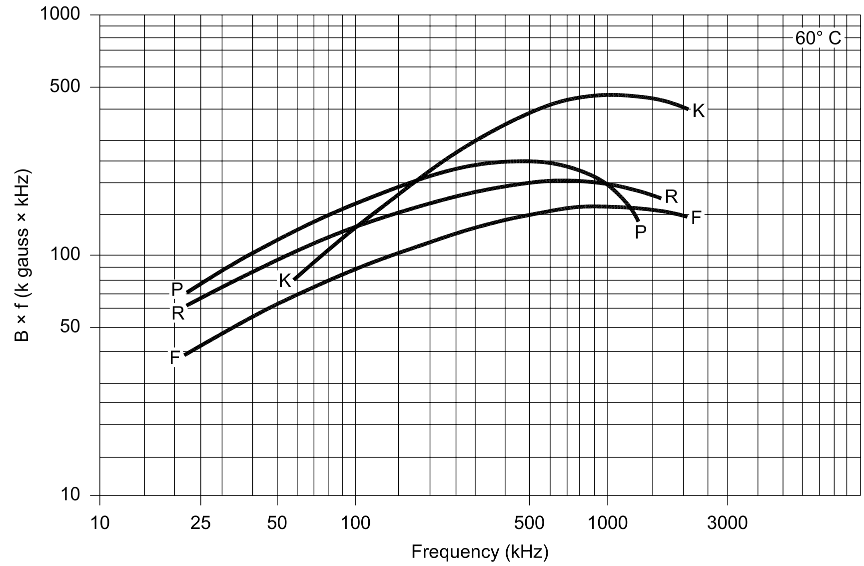


Fig 5—Power handling versus frequency for ferrite cores. The curve letters are Magnetics ferrite designations. They don't have direct correspondence to Fair-Rite numbers.

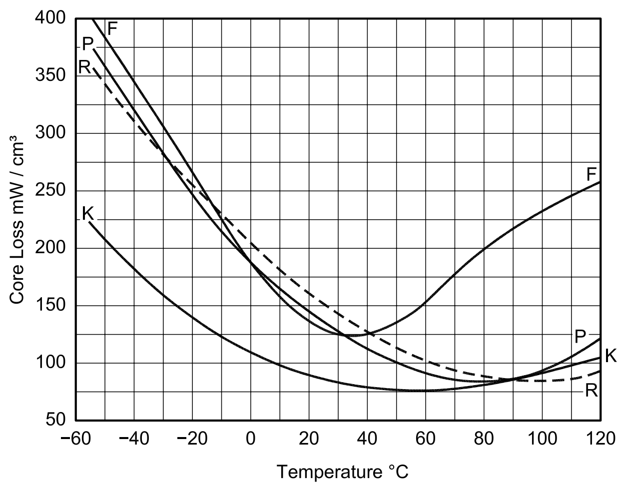


Fig 6—Core loss versus temperature for ferrite cores. Curves F, P and R are at 100 kHz with 1 k Gauss; curve K is at 500 kHz with 500 Gauss. The curve letters are Magnetics ferrite designations. They don't have direct correspondence to Fair-Rite numbers.

Table 4—Coefficients for Eq 1 Change of Permeability due to dc Flux

Material Coefficients

	a	*b*	*c*	*d*	*e*
2	10000	–4.99e–3	–49.5	9.16e–6	0.0865
8	10090	4.26e–3	30.9	7.68e–5	–0.0119
18	9990	8.36e–4	14.4	3.92e–4	0.0853
26	10090	5.05e–3	13.1	1.17e–3	0.0212
52	10240	6.71e–3	24.7	7.75e–4	–0.0105

Table 5—Coefficients for Eqs 2 and 3 Change of Permeability due to ac Flux

Material Coefficients

	a	*b*	*c*	*d*	*e*
2	9970	5.77e–4	7.29	–8.96e–8	–1.18e–3
8	9990	4.52e–4	11.4	8.82e–9	–8.29e–4
18	10270	1.01e–4	12.3	2.70e–8	–8.43e–4
26	10600	7.21e–5	37.8	–7.74e–9	–3.56e–3
52	92.0	0.0134	2.77	–3.66e–6	

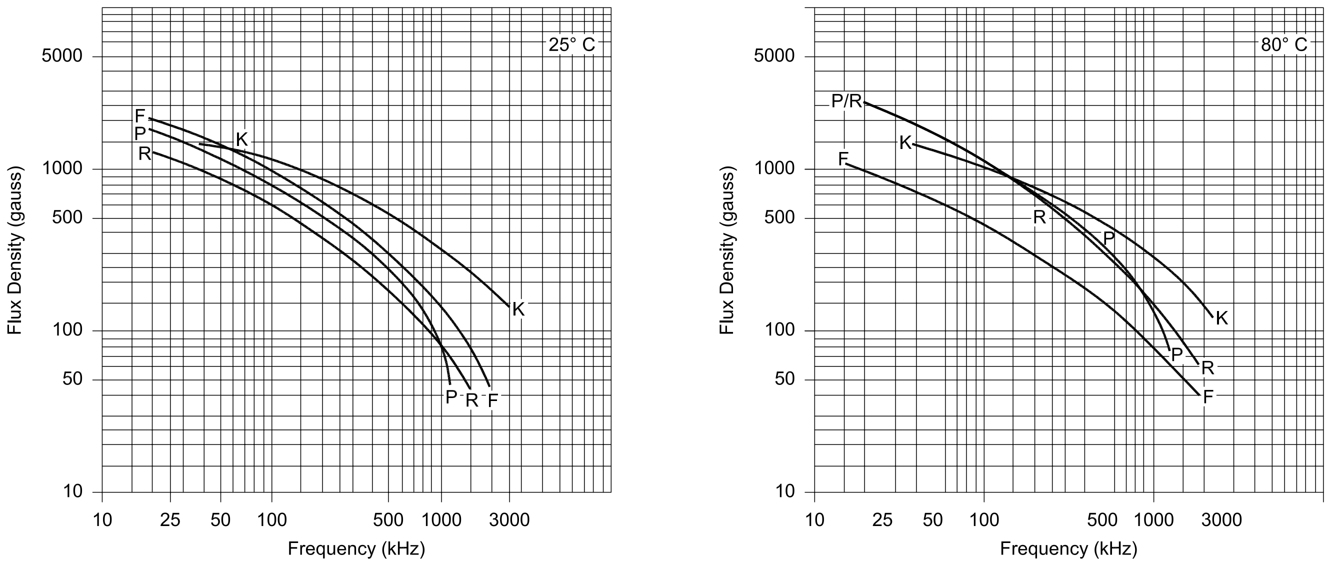


Fig 7—Maximum flux density versus frequency for ferrite cores at 25°C and 80°C. The curve letters are Magnetics ferrite designations. They don't have direct correspondence to Fair-Rite numbers.

higher in boost converter designs than in buck converters, so we must account for permeability changes due to ac flux as well as dc flux and core losses due to ac flux. The formula for percent permeability change due to ac flux for Micrometals materials has the form:

$$\%\mu = \left(\frac{a + cB + eB^2}{1 + bB + dB^2}\right)^{\frac{1}{2}} \qquad \text{(Eq 3)}$$

Table 5 lists the values of these coefficients for the materials of interest.

Material -52 uses the formula:

$$\%\mu = a + bB + cB^{\frac{1}{2}} + dB^2 \qquad \text{(Eq 4)}$$

Table 5 also lists these values.

We need to determine the value of B (the ac flux in the core). For a rectangular wave the formula is:

$$B = \frac{E\,t\,10^8}{2\,A\,N} = \frac{L\,\Delta I\,10^8}{2\,A\,N} \qquad \text{(Eq 5)}$$

Where E is the peak voltage across the inductor, t is the on time, A is the magnetic cross section of the core, N is the number of turns, L is the inductance, and ΔI is the peak-to-peak ripple current. $(L \times \Delta I)/N$ is equivalent to $(E \times t)/N$ or volt-seconds/turn. Table 6 gives the volt-microsecond/turn rating of each core size for a 15°C rise due to ac flux core loss for material-26.

Let's look at an example where the average inductor current is 5.0 A, the peak-to-peak ripple current is 2.5 A and the frequency is 100 kHz. Our design requires an inductor with a minimum inductance of 15 µH and a maximum temperature rise of 40°C. Let's start with the same T106-26 from

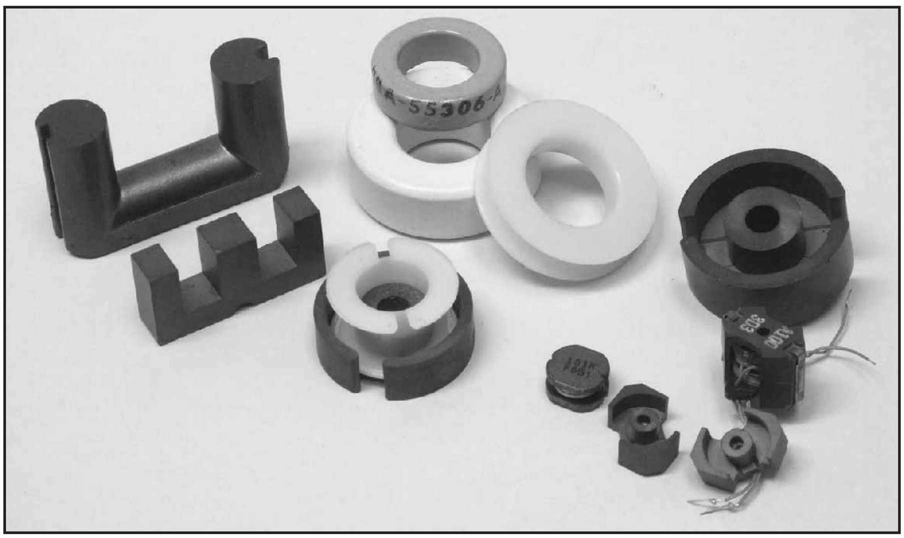


Fig 8—Photo of various ferrite cores.

Table 6—E t /N Ratings of Powder cores for 15 C° Rise Mix 26

Core	*Frequency*			
	50 kHz	*100 kHz*	*250 kHz*	*500 kHz*
T37	0.86	0.54	0.29	0.18
T50	1.30	0.81	0.44	0.28
T68	1.80	1.10	0.62	0.39
T80	2.20	1.40	0.75	0.47
T94	3.10	1.90	1.00	0.66
T106	4.70	3.00	1.60	1.00
T130	5.00	3.20	1.70	1.10
T157	6.90	4.30	2.30	1.50
T184	10.0	6.50	3.50	2.20
T400	17.0	11.0	5.90	3.70

our previous example. We again calculate 13 turns required for 15 µH based on initial permeability. Now we calculate the reduction in inductance due to dc flux:

$$H - \frac{(0.4\pi \; N\,I)}{l} = \frac{(0.4\pi \; 13 \bullet 5)}{6.49} = 12.6 \text{ O}$$

(Eq 5A)

We verify that 12.6 Oersteds is below the saturation point using Fig 4.

We can calculate the change in permeability due to dc bias from Eq 1:

$$\%\mu = \left(\frac{10090 + 13.1H + 0.212H^2}{1 + 0.00505H + 0.00117H^2}\right)^{\frac{1}{2}}$$

$$= \left(\frac{10090 + 165 + 3.37}{1 + 0.636 + 0.186}\right)^{\frac{1}{2}}$$

$$= 75\%$$

(Eq 5B)

Now we can adjust the number of turns based on the reduced permeability:

$$N = 100\left(\frac{L}{A_1 \times 0.75}\right)^{\frac{1}{2}}$$

$$= 100\left(\frac{15}{900 \times 0.75}\right)^{\frac{1}{2}}$$

$$= 15 \text{ turns}$$

(Eq 5C)

Next we calculate the volt-microseconds/turn:

$$\frac{Et}{N} = \frac{L \times \Delta I}{N}$$

$$= \frac{15\ \mu\text{H} \times 2.5\ \text{A}}{15}$$

$$= 2.5$$

(Eq 5D)

From Table 6 we see that 2.5 volt-microseconds/turn will give us slightly less than 15°C rise in temperature. This allows us at least 25°C rise due to copper loss. We look at Table 1 in the 400 A/cm² column to find that we need #19 AWG wire. Since #18 and #20 are the most available standard sizes, we can choose #18 to be safe. Since we are slightly under budget on core loss and copper loss, we can probably use #20 if it is convenient.

If the ac flux density is high, the permeability can be affected as well. We can check the ac flux density to see if there is a noticeable effect:

$$B = \frac{15\ \mu\text{H} \times 2.5\ \text{A} \times 10^8}{2 \times 0.659 \times 15} = 190 \text{ Gauss}$$

(Eq 5E)

From Fig 2, we see that this level of ac flux will increase the permeability by 30%. This gives us plenty of margin to stay above the minimum inductance.

Transformer Properties

Fig 5 shows the power-handling capability of various ferrite materials compared to operating frequency. The graph plots a figure of merit (*B* times frequency) for a constant power loss of 300 mW/cm³. It is important to verify the agreement of graphs between vendors. Magnetics uses 300 mW/cm³ as in Fig 5, whereas Ferroxcube uses 500 mW/cm³. Fig 6 gives a comparison of core loss versus temperature for various materials. Notice that the test conditions are not identical for all materials. Fig 7 shows how maximum flux density must be decreased as frequency increases for various Magnetics materials and correlates the flux density to temperature rise. It is possible to use lower-frequency compounds above their optimum frequency (where a curve in Fig 5 starts to turn downward) as long as you account for the greater losses and temperature rise. Temperature is related to surface area; increasing surface area gives more heat dissipation for a given loss. Core loss is related to core volume; a greater volume generates more heat for a given flux density. The problem is that volume increases as the cube of the size and surface area increases as the square of the size. It is interesting that the lowest power loss density occurs at temperatures above 25°C; so elevated temperature operation actually increases efficiency.

Fig 8 shows a variety of core shapes and bobbins available for use in inductors and transformers. Cores usually have bobbins available for winding the coils so the final component can be assembled from three or four parts.

Toroid and pot cores provide the largest amount of magnetic shielding. The magnetic field is almost entirely confined to the core. Any amount of magnetic field that is present outside the core represents potential for lessened efficiency. These core types are useful where there is not a lot of heating of the wire due to current flow or heating of the material due to magnetic losses. Generally, these cores are used at power levels below 10 W.

The remaining useful cores are variations of the **E** shape and the **U** shape. These cores have very large open areas that allow the heat generated inside the windings to be dissipated to the surrounding air. This space also makes the windings less efficient due to stray magnetic fields outside the core material. A disadvantage of "E" cores is that the primary and secondary are in close proximity, which makes voltage isolation for safety problematic. A standard **E** core has a square center post. An **EC** core has a rounded center post, which provides for 11% shorter wire for a

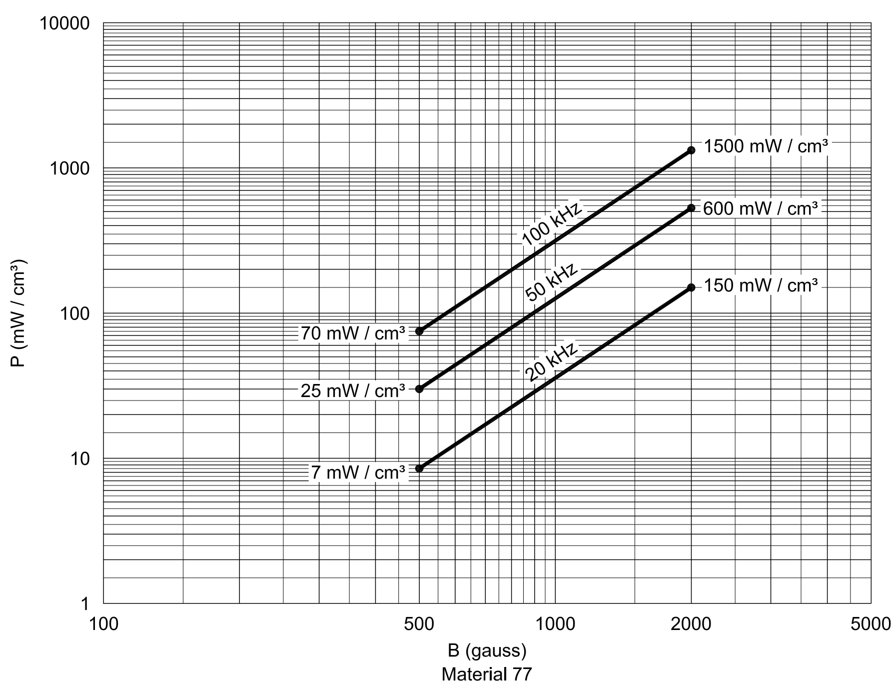


Fig 9—Power-loss density versus flux density for material 77 at various frequencies.

winding. It also means that there is less mechanical stress on the windings because the bend radius is larger. **U**-shaped cores can be used to advantage where high-voltage isolation is required. This shape allows the primary to be wound on a bobbin on one leg and the secondary windings to be wound on the opposite leg. **U** cores are used almost exclusively in television horizontal-output transformers because of the isolation provided.

We must consider the wire for transformer use, because wire can be a source of significant loss. When we looked at choke cores, the wire was a single layer for good thermal characteristics. Transformers are typically wound in layers, with the primary closest to the core and the secondary coils wound on top of the primary. A good rule of thumb (suggested by Magnetics) requires wire sized for 200 to 250 A/cm^2 to minimize the heating caused by wire. As wire size gets larger and frequency gets higher, skin effect losses rise and can be as significant as bulk-resistance losses. Commercial transformers may use Litz wire, copper ribbon or copper strip to reduce the effects of frequency. MWS Wire Industries makes a line of square cross section magnet wire. This wire is more expensive and less readily available, but the square shape is advantageous for making compact individual layers that stack well. MWS also manufactures ribbon and strip. They classify ribbon as having a width up to 0.100 inch and strip as having a width of 0.125 to 2 inches. The skin effect is greatly reduced if the conductor is flattened into a thin strip. Many commercial power supplies use strip conductors for the low-voltage, very-high-current windings such as a 20-A, 5-V winding in a PC power supply. These windings are typically only 2 to 5 turns.

Polyester tape is usually wrapped around each layer of the primary winding of an off-line transformer to provide safety isolation between layers of a winding and between windings. Tape that is 1 mil thick is rated for 5500 V of isolation up to 130°C. Tape 2-mils thick is rated for 7000 V. If higher temperature is required, use Kapton polyimide tape, which is rated to 155°C and 7500 V. Copper strip is usually not insulated, so the strip is insulated with polyester tape.

The limiting factor for ferrites at frequencies above 20 kHz is temperature rise due to core losses. Below 20-kHz operation, saturation flux density is the limiting factor. Modern supplies operate at a minimum of 20 kHz and more likely at 100 kHz, or higher. Old articles about choosing a core will refer to the large power capacity of cores with bipolar (ac) drive, as in the half-bridge and the full-bridge converter topologies. This is not true for modern high-frequency designs because the limiting factor is no longer the saturation flux for the core, but rather the heat rise in the core due to hysteresis losses. AC drive circuits offer significantly more margin between the maximum flux and the saturation flux than unipolar drive circuits, but even unipolar drive designs are unlikely to approach saturation flux densities before things start to melt inside the transformer. It is still good design practice to verify the margin for saturation flux under worst-case conditions.

Choosing a Forward-Converter Transformer Core

Our example will be a transformer designed for a single switch forward converter operating at 100 kHz with a turns ratio of 5.7:1, input voltage of 310 V and total power of 100 W (12.6 V at 8.0 A). We will choose an Amidon Associates core since they are easy to obtain. From the Amidon Web site, we see that the EA-77-500 core is rated for 100 W. Since our turns ratio is not an integer, we choose 17 primary turns and 3 secondary turns (5.67:1).

The physical dimensions (in cm) for this core (½ core) are:

W	*L*	*T*	*Window W*	*Window L*
4.13	1.65	1.27	1.03	0.792

This gives a volume for two core halves of 13.2 cm^3.

Eq 6 gives the flux density:

$$B = \frac{E}{4 \times A \times N \times F \times 10^{-8}} \qquad \text{(Eq 6)}$$

E = RMS voltage $V_{P\text{-}P}/2$ for a square wave

A = core magnetic area in cm^2 (from data sheet)

N= Number of primary turns

F= frequency in Hz

For our core

$$B = \frac{155}{4 \times 1.60 \times 17 \times 10^5 \times 10^{-8}} = 1424 \text{ Ga} \qquad \text{(Eq 6A)}$$

From Fig 9, we see that this turns ratio will yield 600 mW/cm^3. This value will produce a very great temperature rise. We can reduce the flux density by increasing the primary turns. Next we try a 34-turn primary and 6-turn secondary, which yields 712 Gauss. Now we have 150 mW/cm^3, which will give approximately 50°C of temperature rise. Total core loss is 13.2 cm^3 × 150 mW/cm^3 = 1.98 W. We have a core temperature of 75°C before we factor wire losses. We need to minimize wire losses if we use this flux density. It appears that we need even more turns in order to lower flux density and reduce temperature rise. A 51-turn primary will reduce the flux density to 475 Gauss and produce a power loss of 40 mW/cm^3. This will produce a minimal temperature rise with a core loss of 500 mW. Looking at Table 1 for the wire, we see that we need #18 AWG wire for the primary and #10 AWG for the secondary (using 200 A/cm^2).

Choosing a Flyback-Converter "Transformer" Core

We will design a transformer for a 60-W flyback converter that will provide 5 V at 5 A and ±12.0 V @ 1.5 A. The design requires an inductance of 4.5 mH with an input voltage of 310 V and operates at 20 kHz. We will look again at a material-77 **E** core from Amidon. We will need a gapped core to ensure that the core does not saturate. We have two options for a gapped core: We can order a gapped core as a special item from Amidon, or we can assemble the core halves using plastic shim stock. Since the input voltage is 310 V, we will need 195 mA average current through the inductor to produce 60 W, but this design has a peak-current requirement that results in ΔI of 2 A. We start with the same EA-77-500 core as in our forward-converter example.

First, we need to find the inductance correction factor caused by the gap from Eq 7:

$$k = 1 + \left(\mu_i \frac{G}{l_e} \right) \qquad \text{(Eq 7)}$$

where

μ_i = initial permeability from data sheet

G = gap length in mm

l_e = effective path length of the core

If we use 0.5-mm shims, the actual gap is 1 mm because there is a gap across the center post and in each parallel leg; k is then

$$1 + \frac{2000 \times 1}{76.7} = 27 \qquad \text{(Eq 8)}$$

Now we can calculate the number of turns required for the inductance:

$$N = 1000 \times \left(\frac{L \times k}{A_l} \right)^{\frac{1}{2}}$$

$$= 1000 \left(\frac{4.5 \times 27}{4470} \right)^{\frac{1}{2}} \quad \text{(Eq 9)}$$

$$= 0.165 \times 1000 = 165 \text{ turns}$$

Notice that Amidon and Micro-metals list A_l in µH/100 turns for powder cores, and Amidon lists A_l in mH/1000 turns for ferrites.

Next, we need to use Eq 5 to determine the peak ac flux density in the core.

$$B = \frac{L \times \Delta I \times 10^8}{2 \times A \times N}$$

$$= \frac{0.0045 \times 2.0 \times 10^8}{2 \times 1.60 \times 165} \quad \text{(Eq 10)}$$

$$= 1700 \text{ Gauss}$$

From Fig 9, we see that the ac flux core loss for this core will be about 100 mW/cm^3. Total core loss is 13.2cm^3 × 100 mW/cm^3 = 1.32 W. From Table 1, we can use either #28 or #26 AWG wire for the primary, #12 wire for the 5-V secondary and #18 wire for the 12-V secondary windings.

Sources of Information

I have listed addresses and Web sites for each of the manufacturers mentioned. In particular, Amidon and Magnetics were very helpful in giving information and permission to use their data. Amidon has only limited information on their Web site. You can find additional information on Amidon materials at the ByteMark Web site. ByteMark is a distributor and value-added coil-winding service for Amidon and other materials. They also distribute an interesting niche product for "flat transformer" technology.

Amidon Inc, 240 Briggs Ave, Costa Mesa, CA 92626; **www.amidon-inductive.com**.

Fair-Rite, Box J, 1 Commercial Row, Wallkill, NJ 12589; **www.fair-rite.com**.

ByteMark, 1510 E Edinger Ave #B, Santa Ana, CA 92705; **www.bytemark.com**.

Magnetics, Box 391, Butler, PA 16003; **www.mag-inc.com**.

Ferroxcube USA Inc, 12375 B Pine Springs, El Paso, TX 79928; **www.ferroxcube.com**

Errata from Part 1

I made a graphical error in cutting and pasting to make Figs 11 and 12 for the half-bridge and full-bridge circuits. The secondary side in both drawings should have had a full-wave rectifier identical to that in Fig 10. The half-wave rectifier circuit shown will work, but it will be very inefficient. The phase that does not drive the output will only reset the flux in the transformer without supplying current to the load. A full-wave rectifier will deliver the proper amount of current during both phases.

In addition, it was not obvious that the initial condition of the circuit of Fig 4 is with the electronically controlled switch open. The switch does not start closing until the control circuit stabilizes. During the stabilization time, the capacitor and inductor charge up with initial voltage and current.

We Design And Manufacture To Meet Your Requirements
*Protoype or Production Quantities

800-522-2253

This Number May Not Save Your Life...

But it could make it a lot easier! Especially when it comes to ordering non-standard connectors.

RF/MICROWAVE CONNECTORS, CABLES AND ASSEMBLIES

- Specials our specialty. Virtually any SMA, N, TNC, HN, LC, RP, BNC, SMB, or SMC delivered in 2-4 weeks.
- Cross reference library to all major manufacturers.
- Experts in supplying "hard to get" RF connectors.
- Our adapters can satisfy virtually any combination of requirements between series.
- Extensive inventory of passive RF/Microwave components including attenuators, terminations and dividers.
- No minimum order.

NEMAL
Cable & Connectors for the Electronics Industry

NEMAL ELECTRONICS INTERNATIONAL, INC.
12240 N.E. 14TH AVENUE
NORTH MIAMI, FL 33161
TEL: 305-899-0900 • FAX: 305-895-8178
E-MAIL: INFO@NEMAL.COM
BRASIL: (011) 5535-2368

URL: WWW.NEMAL.COM

Linrad: *New Possibilities for the Communications Experimenter, Part 2*

From the Analog World into the Digital: How do we get the desired signal from RF to the sound card?

By Leif Åsbrink, SM5BSZ

Linrad is a very flexible computer program that can work with more or less any digital data stream that can be moved into the computer. This is not unique to *Linrad*. Once the signal is present in a digital format, the flexibility of digital systems should make it reasonably straightforward to interface it to any software-defined radio. Bringing the radio signal into the digital world can be done in many different ways. The hardware descriptions below and the discussion of their advantages and disadvantages are an introduction to *Linrad*, but everything is fully applicable to any other software-defined receiver.

Jaders Prastgard 3265
63505 Eskilstuna, Sweden
leif.asbrink@mbox300.swipnet.se

The Filter Method

Part 1 of this series pointed out that frequency conversion and A/D conversion are closely related. One way of understanding that is to think about what would happen if the output of an A/D converter were routed to a D/A converter without any digital processing at all in between.

Half the frequency of the sampling clock is called the Nyquist frequency. It limits the highest frequency that can be represented by the digital data. The A/D converter is like a frequency mixer in which the LO is a signal at the Nyquist frequency and at its overtones. As in any other mixer, an output at a frequency f may be caused by an input at the frequency $(LO - f)$, $(LO + f)$, $(2LO - f)$, $(2LO + f)$, $(3LO - f)$ and so forth. In the analog world, one of the pair $(LO - f)$ and $(LO + f)$ is usually the desired signal while the other of this pair is the mirror-image frequency. The remaining false responses are denoted *spurs*, and they are weak if the analog mixer is good. In the digital world, all the unwanted responses are called *aliases* and many of them may be equally strong.

It is well-known that RF filters are needed in an analog radio to suppress the mirror image and spurs caused by the first LO. For an A/D converter, the corresponding filters are known as *anti-aliasing* filters.

If the RF filter were tuned to a response caused by an LO overtone, as is sometimes done in microwave receivers, the analog mixer would be called a harmonic mixer. If the filters in front of an A/D converter selected one of the responses above the Nyquist frequency, the A/D would be *under-*

sampling. Under-sampling is done to convert signals above half the maximum sampling speed for a particular A/D converter. It only works if the analog bandwidth of the A/D is high enough, typically several times higher than the maximum frequency for the sampling clock. Such A/D converters are known as radio A/Ds.

It is up to the system designer to decide what frequency range is desired and how much undesired signals must be suppressed. The problem is the same for an A/D converter as for an analog frequency mixer.

In the case where the lowest response, dc to the Nyquist frequency, is selected for an A/D, one needs a steep IF filter to suppress the image frequency that occurs in the mixing process from RF to audio. No filter is perfectly steep, so one will typically lose 15% of the possible digital bandwidth or more. Using an IF filter so that low frequencies reaching the A/D only come from IF signals at one side of the last LO (the BFO) is called the *filter method* because a qualified filter is necessary. The anti-aliasing filter needs a similar steepness, but since it is at a low frequency, it is comparatively simple. When using the filter method, one typically uses the IF-filter to improve alias suppression also.

The Phasing Method

The phasing method or direct-conversion method does not need any qualified filters. It is similar to transmitters wherein the phasing method can be used to generate SSB without crystal filters. By generating the complete (complex) baseband signal from the RF (or IF) signal, one feeds the computer with enough information to decide—for each signal—from which side of the BFO it comes. To do this, one needs two A/D channels. Each of them needs an anti-aliasing filter, but that is a simple filter—often included in the A/D converter board.

In the phasing method, two mixers in quadrature (the relative phases differ by 90°) generate an I/Q pair. A signal 1 kHz above the LO will produce a signal of 1 kHz with equal amplitude in both I and Q and their relative phase will be 90°. The computer program will combine them and correctly place the signal 1 kHz above zero. The analog circuitry suffers from tolerance problems that create a mirror-image spur at –1 kHz with an amplitude that is typically 30 to 40 dB below the main signal. It is possible to balance out this mirror image with analog phase and amplitude controls, but it can be dealt with equally well in the digital processing.

Desired Bandwidth Selects Technology

When selecting hardware to convert from RF to digital data, the most important aspect is the bandwidth. For many purposes, it is enough to move a bandwidth of 2 kHz into the computer. That is extremely easy: A conventional SSB radio and a simple audio-interface board will do the job. The dynamic range within the 2-kHz passband will be very poor, but the tolerance to strong signals outside the passband is at least as good as the analog radio specifications.

If more bandwidth is desired, say 10 kHz or so, one needs a solution with much better linearity than the product-detector and audio-amplifier circuitry of a normal receiver. The easiest way of doing this is to use the filter method with over-sampling. One uses an IF filter to select a passband that comes out as frequencies between something like 0.45 and 0.85 times the Nyquist frequency. The main problem of audio nonlinearities and overtones then won't be present and one can use simple circuitry with good results. With a 96-kHz audio board, one can easily get a bandwidth of about 20 kHz this way. The digital signal will be over-sampled but the extra load on a reasonably modern computer is not a problem. It is possible to get 20-kHz bandwidth using the filter method while sampling at 44.1 kHz, if a special crystal filter is used.

With PC sound cards, it is possible to get bandwidths up to about 95 kHz. If a radio A/D were used, the bandwidth limitation would come from the processing capabilities of the digital system and from the analog filter in front of the A/D converter. An AD6644 sampling at 65 MHz, preceded by a crystal filter with 200-kHz bandwidth at 10.7 MHz, 70 MHz or any other frequency, would make an excellent IF-to-digital converter. As far as I know, there is no standard for interfacing a radio A/D to a PC. The rest of this article shows what we can do with the sound cards that are around now.

The crucial factor is the desired bandwidth. For normal usage as a radio receiver, Table 1 gives an idea how dynamic range requirements increase with bandwidth.

It is extremely easy to make an RF-to-audio converter with a dynamic range of at least 40 dB, independent of the bandwidth. If *Linrad* were used to monitor microwave bands, low-cost solutions for the hardware would be appropriate. Occasional occurrences of strong local signals causing a few spurs on the screen then would not be a problem.

Below is a discussion of some circuits I have used to move radio signals into a PC via sound cards. Any of these solutions will work fine with *Linrad* if your local radio environment is compatible with the dynamic-range limitations of the particular solution.

The cost of increasing the bandwidth while maintaining adequate dynamic range is substantial. It is great fun to monitor nearly 100 kHz on the screen but for practical operation, chasing DX stations in a contest or just ordinary chatting, a bandwidth of about 2 kHz might be fully adequate for CW.

The main reason for bringing large bandwidths into the computer is that it allows the computer to do efficient noise blanking in a way that analog systems can never do. In a location where a better noise blanker than the one included in your current SSB receiver is not needed, *Linrad* will work happily with the output from your SSB receiver and allow you to dig out weak CW signals from the noise.

If power-line noise is not a problem, a 10-kHz bandwidth may be enough. With very difficult power-line noise problems, even 100 kHz is marginal because the interference pulse rate can become very high.

Table 1

Bandwidth (kHz)	*Dynamic Range (dB)*	*Comment*
2	40	Analog hardware stops strong signals.
10	70	Offending signal quality often makes better performance useless.
20	80	Having to keep really strong signals outside the passband is not a serious limitation at modest bandwidths. Is often necessary anyway because of offending signal quality.
100	100	With a large bandwidth one has to allow the local strong stations within the passband.

2-kHz Bandwidth: A Conventional SSB Receiver

It may be a good idea to insert a high-pass filter between the radio and the sound card to reduce 50/60-Hz hum. The audio signal that is routed to the sound card is actually an IF signal and it is a good idea to shift it upwards by a few kilohertz as one usually does when working high-speed meteor scatter. Placing the passband in the range 3-5 kHz rather than 0.3-2.3 kHz by a shift of the BFO greatly improves dynamic range. The dominating nonlinearity is usually IM_2 in the product detector and audio amplifiers, and it will be completely eliminated this way. The lowest possible IM_2 product would be 6 kHz, and that is above the passband. By setting the sound card to sample at 15 kHz or above, all overtones of 3 to 5 kHz are stopped by the anti-aliasing filter or by the digital processing. A shifted audio range also helps to get rid of overtones of the power-line frequency, which often go as high as the tenth harmonic and above when the volume is high and the RF gain is low. Those are the settings for a good dynamic range into the computer. A resistive voltage divider at the computer input is necessary to set the level low enough for the sound card. The volume usually cannot be set low enough without loss of dynamic range because of the noise in the AF amplifier after the volume control.

10-kHz Bandwidth: The Filter Method with Oversampling

Bob Larkin, W7PUA, has designed the DSP-10 to use the Analog Devices EZ-Kit for all the signal processing while the PC is used for control only. The analog hardware of that radio would work well for use with a sound card and PC replacing the EZ-Kit. The schematics (described in *QST,* Sept 1999, pp 33-41) include a crystal filter with the –6-dB points at 19.6591 and 19.6708 MHz. With the last LO at 19.680 MHz, the audio passband is placed from 9.2-20.9 kHz and that fits well to a standard sound card sampling at 48 kHz. Alias signals originate in audio frequencies above 27.1 kHz, and such signals are well suppressed by the anti-aliasing filter of the sound card. Such high audio frequencies originate in IF signals below 19.6529 MHz, so the IF filter gives an additional suppression of about 25 dB. Very strong signals at 19.670 MHz produce audio at 10 kHz with an overtone spur at 20 kHz that will be falsely interpreted as a signal at 19.660 MHz, just at the opposite end of the passband.

This is not a problem: One can simply avoid using the highest frequencies within the passband to receive weak signals. It is also possible to reduce the level of this spur by shifting the last LO downwards from 19.680 kHz until the alias spur starts to be visible at a similar IF level as the IM_2 spur. I have not tried this solution myself, but I see no reason why it should not be perfectly adequate as long as a 10-kHz bandwidth is enough.

20-kHz Bandwidth: The Filter Method without Oversampling.

My first DSP radio in a PC used this method. At that time, standard audio boards were limited to a maximum sampling speed of 44.1 kHz. I needed as much bandwidth as possible for two reasons. First, I suffer from terrible power-line noise now and then. The analog noise blankers do not work when the moon is low, and my local EME friends become very strong, so I needed something clever in the computer. Second, seeing a 20-kHz bandwidth rather than only 10 kHz is an advantage in EME contesting.

To get 20-kHz bandwidth with a sampling rate of 44.1 kHz, requires extremely good crystal filters to avoid spurious responses. How I did that is described in detail at **ham.te.hik.se/~sm5bsz/pcdsp/pcif.htm**. The filter contains 27 surplus crystals for each RF channel, and it is designed for easy tuning rather than low component count. It is flat over 20 kHz and attenuates by more than 100 dB at 2 kHz or more outside the passband. With a more-conventional crystal filter, one must accept somewhat stronger spurious responses at the passband ends, and one must reduce the passband slightly. When using much more than one octave for the audio signal, the mixer becomes a very critical part of the system. Most frequency mixers suffer from second-order distortions. The second harmonic of the IF signal mixes with the second harmonic of the LO to produce an audio signal at twice the frequency of the desired signal. The strengths of such products are proportional to the square of the IF signal level. If the desired signal were reduced by 10 dB, IM_2 would go down by 20 dB only. To get low IM_2 levels from a mixer, one must operate very far below the point of saturation or use some kind of feedback or other technique to reduce IM_2 at high signal levels.

For my first PC-based system, I found Schottky-diode mixers to have far too much IM_2 to be useful. A very simple CMOS mixer (as shown in Fig 1) is good enough despite the rather high noise level of the TL074 op amp. For IFs below 15 MHz or so, this solution offers lower cost and better performance than a Schottky-diode mixer.

40-kHz Bandwidth: A Low-Cost Direct-Conversion Receiver

Using two standard mixers and a crystal oscillator, one can get a

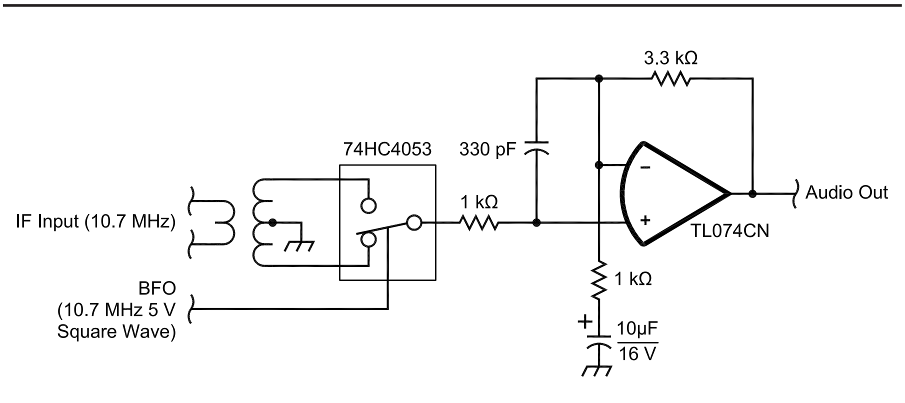


Fig 1—A simple CMOS mixer. The input transformer is not critical and it is typically loaded by a 50 Ω resistor to define the impedance seen by earlier stages.

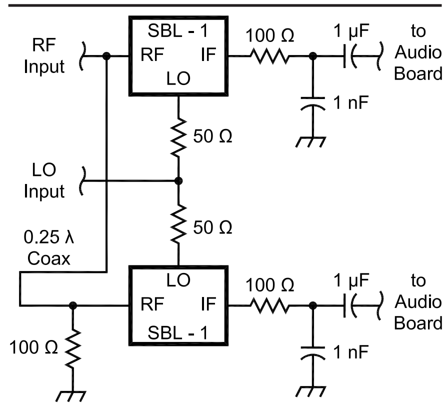


Fig 2—A very easy way to get 40 kHz of the radio spectrum into a PC.

surprisingly good radio receiver, as shown in Fig 2. No anti-aliasing filters are needed; the ones built into the sound card are sufficient if dynamic-range requirements are modest. Without any amplifier at all, the noise figure is far from acceptable; gain on the order of 40 dB needs to be added. Part of the gain is preferably added at baseband. With a very-low-noise amplifier following each mixer, the sensitivity can be made good enough for use without any RF amplifier. Since the LO power level is +10 dBm and one might expect leakage to the RF port about 30 dB below this level, one should never use this circuitry without an RF amplifier. The RF amplifier should also provide some selectivity because a Schottky-diode mixer is sensitive to signals at overtones of the LO frequency. On the HF bands, quite a lot of overtone attenuation is needed. The response at three times the LO frequency is typically only 12 dB below the response at the fundamental. For further details about this simple receiver and performance data, check **ham.te.hik.se/~sm5bsz/linuxdsp/hware/sbl1.htm**.

90-kHz Bandwidth: Another Low-Cost Direct-Conversion Receiver.

Fig 3 shows a low-noise, low-cost audio amplifier that is suitable to insert between a Schottky-diode mixer and the input of an audio board. The input transistor operates at a relatively high current, so it produces a very low noise figure at a source impedance around 50 Ω. The transistor should be rated at 1 A with an F_t of at least 100 MHz. BC489A, BF452 and many others are fine. The op amp that follows the transistor provides high gain, which is fed back to make the voltage swing small at the collector of the input transistor. This way, the second-order distortion is made very low. The second op amp provides 10 dB more gain and stabilizes the operating point of the input transistor through a low-pass filter.

With a low-noise amplifier following a Schottky-diode mixer, one easily runs into problems with the stability and purity of the local oscillator. At 144 MHz, even an IC-202 is not good enough, and that's the purest commercial transmitter I know. It is possible to lower the noise floor of an IC-202 by about 10 dB: First, replace the 470-Ω emitter resistor of the LO with a 150-Ω resistor in series with an RF choke. Then, decouple the base of the first frequency-multiplier stage for low frequencies by installing a 1-µF capacitor in series with an RF choke.

Modified like this, it is just about good enough to be used as the LO in a direct-conversion 144-MHz receiver, but only if a low-level Schottky-diode mixer were used. A high-level mixer needs about 15 dB more of LO power and would therefore be much more susceptible to LO sideband noise. Besides the desired mixing of the signal at the LO port with the signal at the RF port, a mixer will mix the signal at the LO port with itself and its own overtones. As a result, a Schottky-diode mixer will detect both AM and FM undesired modulation that may be present on the LO signal. With more LO power, the undesired AM/FM detection will produce more audio noise at the mixer port. One way of understanding it is like this: If a high-level mixer were used, about 15 dB more LO power would be needed. Both the LO carrier and the sidebands that surround it would then be 15 dB stronger so the mixing product, audio noise, would be 30 dB stronger. It is difficult, if not impossible, to make a local oscillator that is good enough for a high level Schottky mixer in a direct-conversion radio on 144 MHz.

The amplifier of Fig 3 works fine for signals up to about 75 kHz. If higher frequencies were allowed to enter the amplifier, the input transistor would go into nonlinear operation because the feedback is not fast enough. A low-pass filter is required to protect the amplifier from strong signals at high frequencies. For further details about this reasonably good 144-MHz receiver, including performance data, check **ham.te.hik.se/~sm5bsz/linuxdsp/hware/optiq.htm**.

95-kHz Bandwidth: A No-Compromise Design

An analysis of the previous design made it clear that conversion to baseband should not be made directly from VHF frequencies. It is much better to use an intermediate frequency at which a very good local oscillator can be made easily. I have started to design dedicated hardware for *Linrad*, or for any other SDR software that would interface to 96-kHz sound cards. I'm building this hardware primarily for my own use, so the units are rather big and have substantial power consumption, none of which matters at all to me personally. The first unit is a

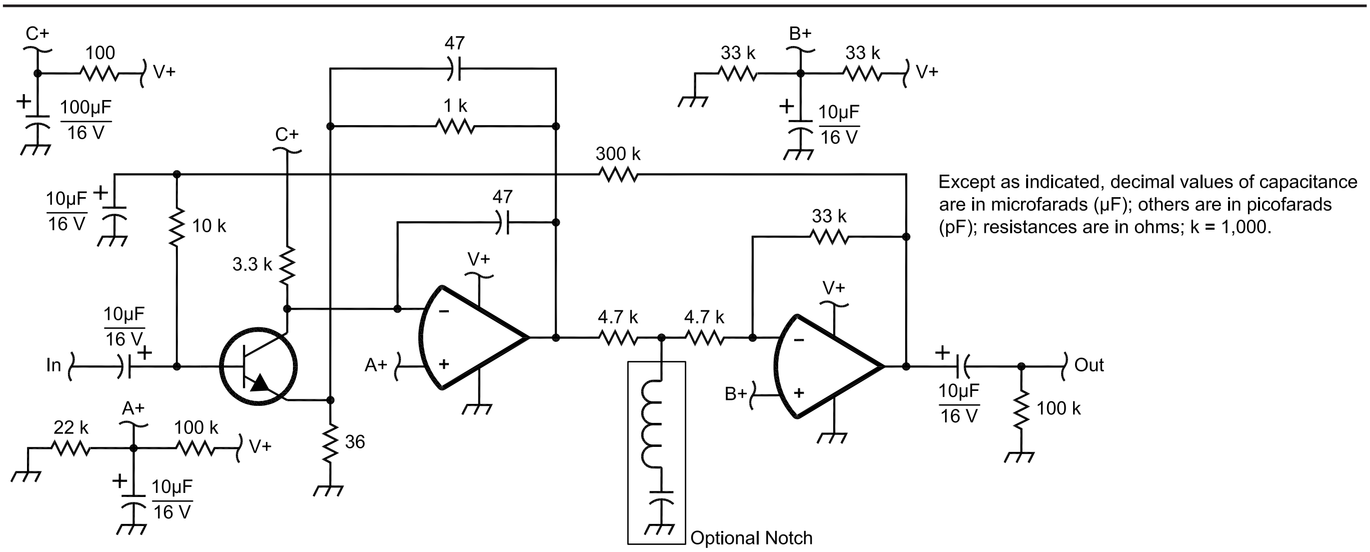


Fig 3—A low-cost audio amplifier with low noise and good linearity. An AD797 with two resistors can replace the entire circuit, but it is expensive and may be hard to find. This circuit gives very good performance with junk-box-grade components. A, B and C are common to I and Q and to both channels in a two-channel system.

2.5-MHz-to-baseband converter with a usable bandwidth of 95 kHz. This unit will be available from Svenska Antennspecialisten **www.antennspecialisten.com**. There will also be some other units to convert to 2.5 MHz from the different amateur bands via higher IF frequencies at 70 MHz and at 10.7 MHz. All the units will be described in detail with links from **ham.te.hik.se/~sm5bsz/linuxdsp/optrx.htm**.

To avoid any problems with sideband noise on the local oscillator, the last IF is placed as low as 2.5 MHz. The LO is obtained by dividing a 10-MHz oscillator by 4. Refer to Fig 4 for the circuit diagram.

A crystal oscillator can be viewed as an amplifier that has positive feedback through a filter. A 10-MHz crystal has a resistance of about 4 Ω at series resonance. When a crystal is used to connect the output of an amplifier to its input, both the input impedance and the output impedance of the amplifier must be well below 4 Ω, not to degrade (increase) the bandwidth. The voltage at the collector of the oscillator transistor is divided by about 50 times by the essentially capacitive voltage divider. This means that the source impedance feeding the crystal is about 2000 times smaller than the impedance at the collector, or about 1 Ω. The input impedance of a grounded base BF240 is very low, so this oscillator has the potential of being very good in terms of sideband noise.

The frequency divider, a 74AC74, must be fed from a low-impedance source. The clock input of this chip is not an amplifier with extremely low noise figure, and it helps to make the source impedance low to avoid phase jitter caused by the equivalent noise current at the input. It is also a good idea to feed the clock input with a high signal level. In this circuit, the pk-pk voltage at the clock input is 4.5 V, and the impedance is about 150 Ω, which is low enough at 10 MHz to do the frequency mixing from 2.5 MHz to baseband. An extremely linear mixer is needed to preserve the very good linearity and high dynamic range of 24-bit, 96-kHz sound cards. The four-phase AF feedback mixer shown in Fig 5 has an extremely low level of IM_2. Fig 6 shows a two-tone test with this unit. Two equally strong signals, both at –22 dBm, are fed into the 2.5-MHz-to-baseband converter connected to a modified Delta44 board. The two signals combine for a peak amplitude of –16 dBm, which is 1.8 dB below saturation for the A/D converters.

The two signals are at 2.4993 and at 2.5092 MHz and show up at 47.3 and 57.2 kHz, respectively, in the *Linrad* spectrum at a level of about 120 dB. The strongest spurs are at the mirror frequencies. In normal operation when *Linrad* is properly calibrated, these spurs are pushed down from about 85 to 45 dB in software.

IM_2 occurs at 46.6, 49.5, 66.4 and 38.8 kHz. The amplitude and phase relation of the IM products does not fit what is expected for a baseband signal, so each IM_2 spur is split into two frequencies. The level of the IM_2 products is about 37 dB, 83 dB below the main signals. TheIM_2 products originate in the Delta44.

IM_3 occurs at 37.4 and 67.1 kHz at a level of 32 dB, 88 dB below the main signals. Since the input signal is –22 dBm, we may use the numbers to extract IP_3, which turns out to be +22 dBm. Unlike a conventional radio that typically does not misbehave too badly until the level is somewhere around 15 dB below IP_3, the 2.5-MHz unit makes the Delta44 saturate at –14 dBm; saturation is fatal to sensitivity. A DSP radio is not well characterized by IP_3 numbers. The level of the intermodulation products at the point of saturation, given in decibels below the main signals, is the best way to specify system linearity as determined by a two-tone test. The 2.5-MHz receiver has IM_2 at –81 dBc and IM_3 at –84 dBc. To fully characterize the in-band properties of this system, one must add that the noise floor is at –146 dBc/Hz at a frequency separation of 5 kHz. It is –147 dBc/Hz when no strong signal is present. Knowing that saturation is at –14 dBm, one finds that 147 dBc/Hz corresponds to –161 dBm/Hz, which is 13 dB above the noise from a room-temperature resistor (the theoretical limit). The noise figure is thus measured to be 13 dB by use of the *Linrad* S-meter. A conventional noise-figure meter gives a similar result.

At frequency separations below about 25 kHz, the 2.5-MHz-to-baseband converter competes well with any modern transceiver because of the reciprocal mixing problems associated with variable-frequency oscillators in analog receivers. But at frequency separations above 25 kHz with the offending signal within the *Linrad* passband, analog receivers may be better because they do not saturate like A/D converters. If the offending signal is more than 47 kHz from the center frequency, the 2.5-MHz-to-baseband converter does not saturate until at about +5 dBm because of the very sharp anti-alias filter following the mixer. The saturation level grows with the frequency separation to about +15 dBm at 100 kHz. This is an impressive 176 dBc/Hz above the noise floor. This kind of level for saturation is required to keep IM_3 low for signals outside the visible passband. The local oscillator is good enough to not

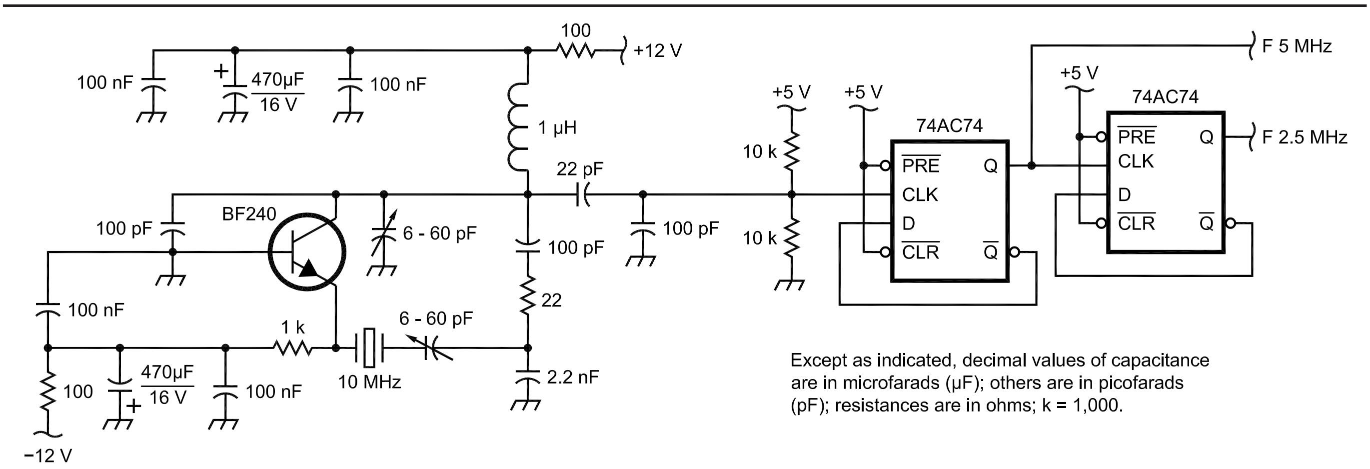


Fig 4—A low-noise crystal oscillator to drive a four-phase CMOS mixer.

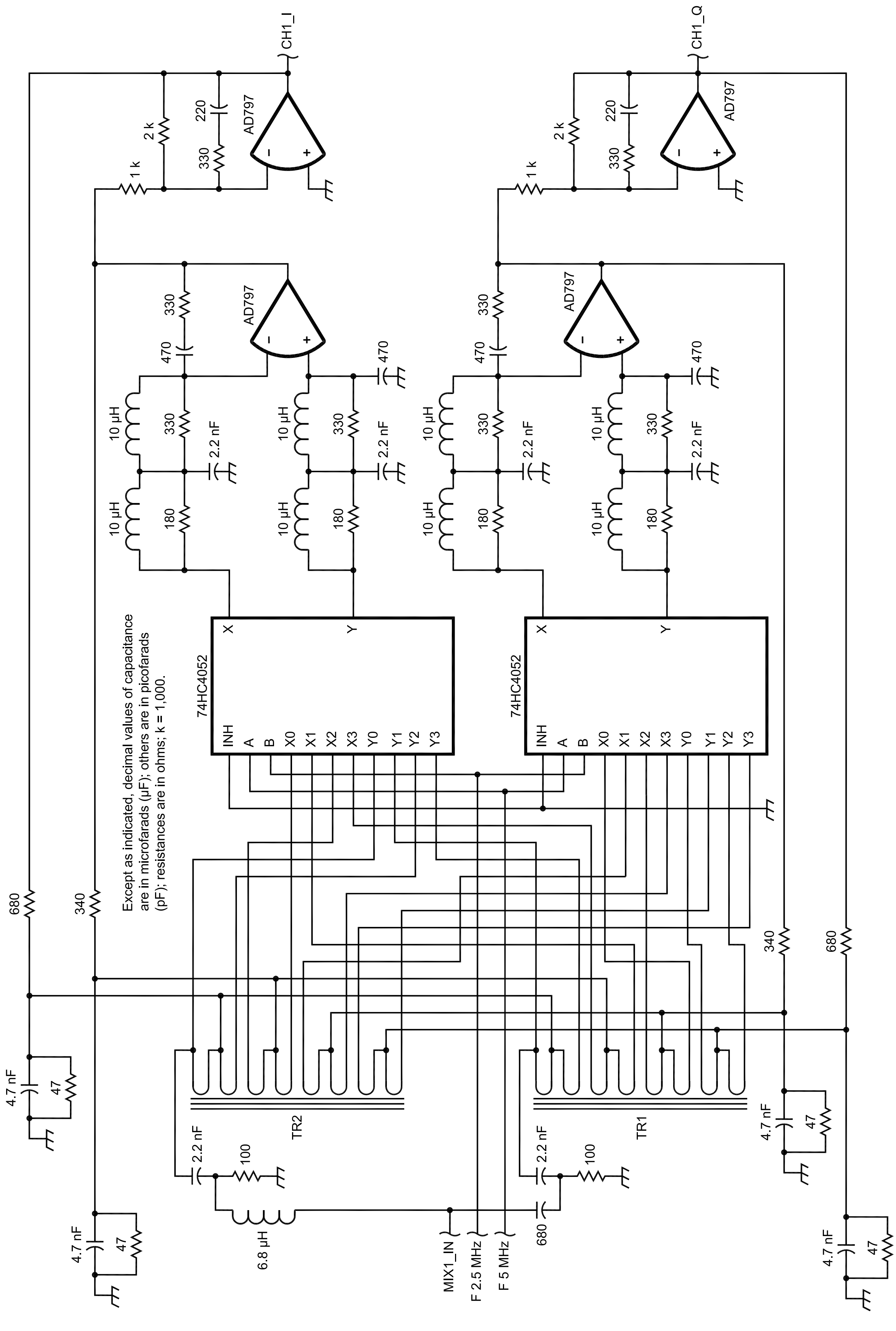

CH1_I
CH1_Q
AD797
2 k
220
330
1 k
470
10 µH
2.2 nF
180
74HC4052
INH
A
B
X0
X1
X2
X3
Y0
Y1
Y2
Y3
X
Y
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
680
340
4.7 nF
47
TR2
TR1
100
6.8 µH
MIX1_IN
F 2.5 MHz
F 5 MHz

Fig 5—(left) The four-phase AF feedback mixer.

degrade the noise floor by reciprocal mixing even for a +15dBm signal at 100 kHz separation.

The dynamic-range numbers given here are for a modified Delta44. The Delta44 has a design error, in that the analog ground of the board is grounded to the computer backplane through the screws that hold the D-sub connector. Some voltages are insufficiently decoupled. With larger capacitors and with the ground loop removed, the noise floor is improved by about 3 dB. In the case where Delta44 is intended for software-radio use only, one can improve by one more decibel. This is done by routing the signal directly to the A/D chip through a voltage divider, thereby excluding the noise from the input amplifier of the Delta44. For details about modifying the Delta44 board, look at **ham.te.hik.se/~sm5bsz/linuxdsp/hware/delta44.htm**.

Problems Associated with Large Bandwidths

The need to allow strong undesired signals within the passband creates several different problems that must be solved when a wide bandwidth system is designed. One must make the system linear up to signal levels as high as possible. At the same time, one wants the noise floor as low as possible. These two requirements are sometimes in conflict. If, for example, a Schottky-diode mixer were used, one can improve linearity by selecting a more expensive mixer that operates with a higher level of LO power. With more LO power, such a mixer becomes a more sensitive AM and FM detector that produces audio noise from the noise sidebands associated with the LO. The sideband noise of the LO may then become the limiting factor for the noise floor, in which case the performance actually is better with a low-cost, low-level mixer. Here is a list of problems associated with wide-bandwidth conversion to audio:

Ground Loops: The mains 50/60Hz and its overtones may flow in the analog ground to the computer sound card. At modest bandwidths, this problem is easily cured by shifting the spectrum upwards as mentioned above. Having a single grounding point (star configuration) for the computer and the rig will help to make the ground-loop currents small. For large bandwidths, when full audio response is required well below 1 kHz, it is often necessary to break up the ground reference. One way is to use a differential amplifier as shown in Fig 7. Any voltage difference between mixer ground and sound-card ground will be compensated for automatically. Another possibility is to break up the ground at the mixer, having separate HF and AF grounds as shown in Fig 8. If the mixer does not have separate ground returns for IF and HF signals, as does the SBL-1, one can isolate RF and LO from the common ground of the mixer by use of small coupling capacitors, typically 10 nF or less.

If the op amp is operated from a single-rail supply, R1 and R2 are made equal. Otherwise R2 is made very large, to make sure the voltage across the electrolytic capacitors does not go the wrong way. R1 should be at least 1 kΩ to keep a high impedance at 50/60Hz between sound-card ground and RF ground.

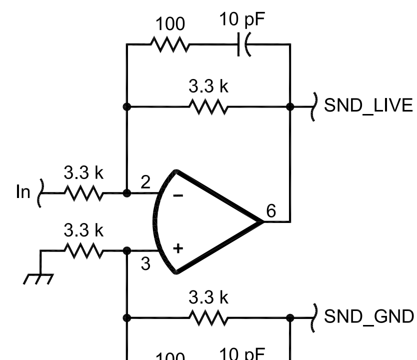
Mixer Linearity: At large signal levels, a mixer saturates. The sine wave that is the ideal response to a pure carrier at some offset from the LO becomes distorted; its overtones constitute spurious responses. A Schottky-diode mixer should not be allowed to give more than about 1% of its saturated output level to make these spurs reasonable in amplitude. FET-switch mixers may produce second harmonics because the current through the switch may have a large component of overtones to the RF frequency. This is particularly true if the switch has capacitors at both sides, an LC filter at RF and a capacitor to ground at the audio side.

RF Amplifier Linearity: An RF amplifier will produce overtones when forced to give large output signals. These overtones mix with the overtones of the LO that circulate in a saturated mixer and produce audio responses that behave exactly as mixer nonlinearities. A good filter that suppresses the harmonics of the desired RF is needed to reduce this problem.

AF Amplifier Linearity: It is obvious that the audio amplifier between the mixer and the sound card must be extremely linear. Distortion creates spurs. It is important to make sure very little RF signal reaches the audio-amplifier input. Less than a 100 mV of RF may be enough to severely degrade linearity.

AF Amplifier Noise: The audio amplifier should not add much noise. Feedback is necessary in order to get good linearity, but it is not necessarily a good idea to arrange the feedback to present a matched load to the mixer output impedance. The AF feedback mixer in Fig 5, as one example, does not load the mixer output at all. In this circuit, two mixers are used in antiphase to make use of both of the op-amp inputs. This way, the signal is increased by 6 dB while the noise is unaffected. The audio frequency impedance at the op-amp inputs is very low. This way, the thermal noise associated with resistive components at room temperature is minimized. The feedback resistor is only 47 Ω to ground. It does contribute some noise, but it

Fig 7—(below) A differential amplifier converts the voltage difference between input and ground to a voltage difference between SND_LIVE and SND_GND. Any voltage difference between ground and SND_GND will not appear at the output.



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k = 1,000.

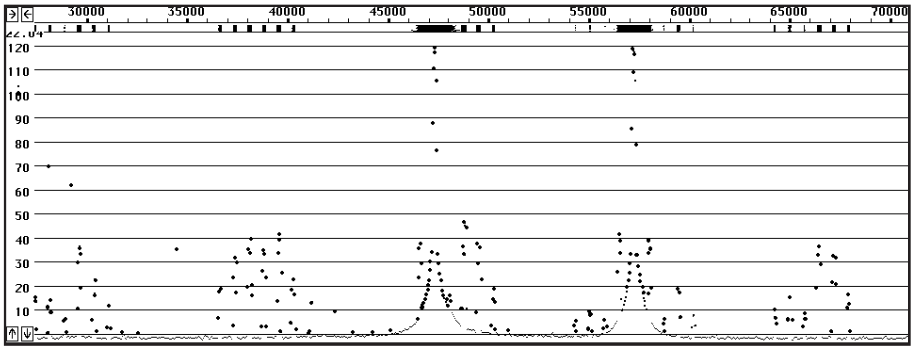


Fig 6—The main spectrum display of *Linrad* when two signals, both of level –22dBm are fed into the 2.5-MHz-to-baseband converter. The A/D is a modified Delta44 board.

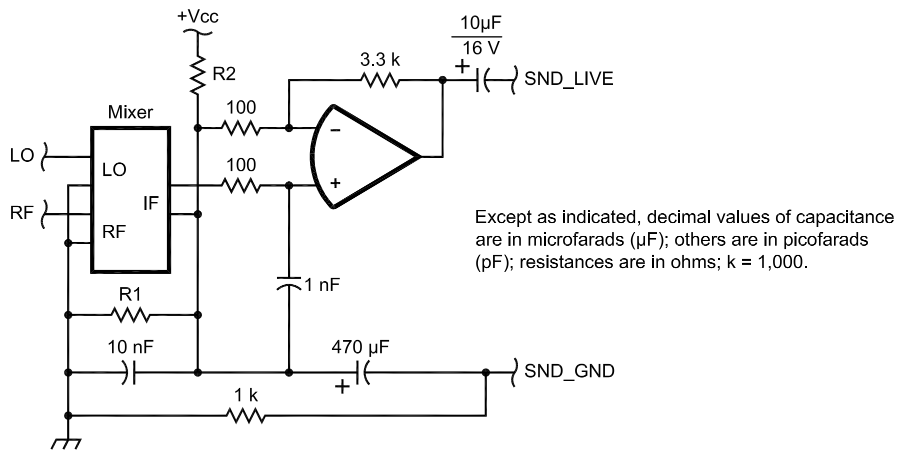


Fig 8—By making the impedance between SND_GND and ground high for audio frequencies, the current through the SND_GND wire will be small. Essentially, all the voltage induced by stray magnetic fields will develop across the 1 kΩ resistor.

cannot be made smaller because the AD797 cannot drive lower impedances than those used here, and more gain would lead to amplifier saturation.

LO Amplitude Noise Modulation: The noise sidebands of a local oscillator are usually referred to as LO phase noise. Amplitude noise is also a problem, and it can easily cause much bigger problems than the phase modulation. When the supply voltage is not extremely well filtered, there will be some AM present on the LO. A Schottky-diode mixer is sensitive to AM on the LO. AM sidebands at –80 dBc typically generate 1 µV of audio at the IF port. A 7812 regulator decoupled by the standard 1-µF capacitor does not produce a sufficiently stable dc voltage. It is necessary to use an additional RC link with something like 10 Ω and 470 µF. Since mixers to audio frequencies are usually very sensitive to AM on the LO, one must make sure that the LO is not modulated by mechanical vibrations. AM sidebands on the LO will also cause noise sidebands on strong signals because of reciprocal mixing, but this is not a major problem since the mixer suppresses AM to some extent in the reciprocal mixing process.

LO Phase Noise: Phase noise on the local oscillator causes noise sidebands on strong signals by reciprocal mixing. If the RF bandwidth is large enough to allow strong undesired signals, reciprocal mixing is a problem regardless of receiver technology. Phase noise on the LO also causes some audio noise on the IF port even when there is no signal present at the RF port. A Schottky mixer detects FM to some extent.

Amplifier AM: All RF amplifiers may introduce AM if the supply voltage is not carefully decoupled. The resulting noise sidebands on strong signals may be difficult to distinguish from noise sidebands caused by reciprocal mixing of LO noise sidebands.

Most of the problems on the list above disappear if a radio A/D is used to sample RF signals directly. I hope there will soon be hardware and drivers available to amateurs at affordable costs. Such systems will require good roofing filters to avoid very large voltages at the A/D input. Maybe a fixed-frequency filter (like 70 MHz) will be the popular solution rather than separate filters for each HF band. High-performance converters from HF to 70 MHz are not difficult. For highest performance, crystal oscillators may be used.

Summary

This article has concerned itself with details in design of the hardware required to move radio signals into the computer by use of sound cards. In the next segment, I shall discuss *Linrad* specifically: how to install it and get it running, what the basic architecture is and how it can be modified by setting various parameters.

Leif was born in 1944 and first licensed in 1961. He holds a PhD in physics and worked with research on molecular physics for 15 years at the Royal Institute of Technology in Stockholm. Since 1981, he has been running his own company. He is the inventor of the intermodulation EAS system now owned by Checkpoint. Leif is a board member of Svenska Antennspecialisten, but is essentially retired and working full time with Linrad *development.*

International Digital Audio Broadcasting Standards: Voice Coding and Amateur Radio Applications

Here's a digital-voice standard for broadcast and Amateur Radio—read all about it!

By Cédric Demeure and Pierre-André Laurent

This paper presents a new digital system for use in terrestrial audio broadcasting in the frequency bands below 30 MHz and its potential use for Amateur Radio. The system is based upon a COFDM modem; it was elaborated as a derivation of the DRM modem. DRM (Digital Radio Mondiale) is a worldwide consortium proposing through ITU a new standard for digital radio broadcasting for frequencies below 30 MHz. It is now the only digital standard adopted by ITU for HF digital radio broadcasting. A specially adapted version that fits inside a 3-kHz channel was derived for Amateur Radio. This paper presents the existing demonstrator to be tested on an across-the-Atlantic link, together with the associated modem and voice coding techniques.

Thales Communications SA
66 Rue du Fossé Blanc
92231 Gennevilliers Cedex
France

A new coherent COFDM (coded orthogonal frequency-division multiplexed) modem for use in high-quality audio broadcasting in the frequency bands below 30 MHz is proposed by DRM as a candidate to replace the current AM modulation that is notoriously perturbed by multipath, Doppler shift and fading. The main characteristics of this system are to be compatible with existing channelization within these bands and with current transmitter technology (5-kHz spacing). Therefore, it should be a better alternative than satellite broadcasting at higher frequencies because it is cheaper, offers better propagation and indoor coverage. It also opens the door for much better quality by offering net bit rates compatible with up-to-date audio compression techniques like AAC-MPEG 4 (near FM quality).

Such a broadcasting scheme may even be implemented using existing power transmitters in a transmission mode named *Simulcast*. This mode consists in transmitting a program in a given frequency channel by using an AM compatible analogue waveform and the new digital signal simultaneously.

This broadcasting system was initially elaborated within the NADIB consortium (European Eureka 1559 project: *NArrow Band DIgital Broadcasting*) and was adopted and enhanced within the DRM consortium as a candidate for international normalization at ITU. Real transmission on a standard 100-kW, shortwave AM transmitter was demonstrated at the

1997 IBC conference in Amsterdam. Since then, numerous field trials have been performed. The latest series (May/June 2002) involved transmissions from Canada, the Caribbean and Europe to Australia and New Zealand.

Thales, with its two affiliates Thales Communications SA and Thales Broadcast and Multimedia SA, was within the initial group of companies to work on the standard.[1] Going for digital broadcasting allows the introduction of new services such as Program Associated Data (PAD), RDS type alternative frequency switch (AFS) management, single frequency network (SFN), simultaneous data channels for various services such as picture transmission, and so forth. This paper describes work performed by Thales Communications France to adapt this standard and the associated modem to 3-kHz Amateur Radio needs.

DRM Standards

DRM stands for Digital Radio Mondiale, where the spelling of the last word is on purpose to show its international nature. DRM is a worldwide consortium established in Geneva, Switzerland, to promote a unique LF, MF and HF audio-broadcasting standard. See **www.drm.org** for more details. This site contains two reference ITU papers on the subject.[2, 3]

DRM's members include: broadcasters and broadcasting associations; network operators; research institutes; component, receiver and transmitter manufacturers; regulatory and standardization authorities.

There are presently more than 80 members in DRM, corresponding to 27 countries. In the USA for example, members include: Harris Broadcast, the International Broadcasting Bureau, Continental Electronics Corporation, Sangean America Inc, Technology for Communications International. Some other active members include: the British Broadcasting Corporation (BBC), Sony, Bosch, Thales, RFI, DW, JVC, Telefunken and so on.

The standardization process involves mainly the ITU (International Telecommunication Union) based in Geneva, the European body ETSI (European Telecommunications Standard Institute) and the ISO (International Standard Office). Other bodies such as IEC and ARIB are also involved.

Key features of the standard include:

- A worldwide standard to allow for unique replacement of the AM format
- Compatibility with existing channelization: The DRM signal is designed to fit in with the existing AM broadcast band plan, based on signals with 9 or 10-kHz channelization. It has modes requiring as little as 4.5- or 5-kHz bandwidth, plus modes that can take advantage of wider bandwidths, such as 18 or 20 kHz.
- Better audio quality: The aim is to obtain near-FM quality within a much narrower frequency bandwidth. The improvement upon analogue AM is immediately noticeable. DRM can be used for a range of audio content, including multi-lingual speech and music.
- Simple-to-use receivers, especially when it comes to HF programming (frequent frequency changes due to propagation conditions)
- Low-cost equipment to quickly reach the mass market
- Text messaging similar to RDS at FM for simple PAD (Program Associated Data) transmission
- Data applications: an open path for new applications of this medium with a large geographical coverage
- Future enhancements: a clear path for new ideas with an open standard

The DRM standard has been designed taking into account a number of technical constraints, among which are:

- Short access time for the receiver: The listener shall not wait for more than a few seconds before getting access to the desired program, and shall obtain radio broadcasting information even faster.
- Maximum quality (objective and subjective): In the allowed transmission bandwidth, a maximum useful bit rate must be conveyed. This implies a high-spectral-efficiency modulation scheme—more than 2 bit/s/Hz.
- Robustness against distortions (multipath, Doppler, noise): This is mandatory, especially in the shortwave bands (which are often severely affected by propagation disturbances and interference) or in medium waves during the transition between day and night.
- Flexibility: According to the current broadcasting frequency band, the frequency separation of different transmissions, the bandwidth of the transmitter and the total available bandwidth can be adjusted to the needs of the broadcasters. In the same way, the required protection level is not the same in LW, MW and SW and in a given band. It can vary according to the time of day. Moreover, the system should include operating modes that can be used in the transition phase, where simultaneous broadcasting (*simulcast*) of compatible AM is required. Notice that a change in any parameter needs no intervention from the listener, since the receiver is remote controlled by the transmitter.
- Minimum disturbance of AM users on the same band: This leads to the design of a signal which, as seen by an AM receiver, must be as noise-like as possible. The frequency spectrum of the transmitted signal must also be as compact as possible to minimize jamming in the adjacent channels.
- Low complexity: This is essential for having receivers of low complexity and low power consumption, especially in countries where batteries are rare and expensive.
- Graceful degradation: If desired, graceful degradation may be obtained by the use of hierarchical coding. This option, although compatible with the system design, is not described in this paper.

DRM benefits for the listeners are the following:

- FM-like sound quality with the AM reach
- Improved reception quality
- Flexible use of radio, whenever and wherever you want it
- No change to existing listening habits: same frequencies, same listening conditions (fixed, portable and mobile radio), same listening environment (indoors, in cities, in dense forests)
- Low-cost receiver, low energy consumption
- Easy tuning with selection by frequency, station name or programming
- More diverse program content, using the full capabilities of new digital features
- Wide receiver range with more and better features

Radios that will give you programs with associated text information, station name, record title, singer's name.

DRM System Description

With the limited bit rate available, it is important to strike the right balance between flexibility and efficiency while protecting each bit of information to an appropriate degree. A distinction is therefore made between main payload data and the various types of data that the receiver needs to help it find and decode the desired program.

The main payload is called the *Main Service Channel (MSC)*. Two subsidiary channels are also provided namely the *Fast Access Channel (FAC)* and the *Service Description Channel (SDC)*. These two are key to ensure simplicity of receiver operation and are therefore designed to be reliably received in adverse conditions, with different forward error-correction schemes from the MSC.

The FAC is intended to be decoded quickly by the receiver on first acquiring the signal (at switch-on, or during scanning). It carries a minimum of

[1]Notes appear on page 56.

constantly repeated data that might be essential at this stage: informing the receiver what bandwidth option is in use, what modulation is used for the SDC and MSC, which length of interleaver is used for the MSC, and so forth.

The FAC is naturally concentrated in the narrowest frequency bandwidth available (namely 4.5 kHz) so that the receiver needs only to receive and decode this bandwidth to know exactly what is transmitted. (See Fig 1.)

The SDC contains more data, also sent repeatedly but in a longer cycle to maintain efficiency. It contains an identification of the services available in the MSC, together with further information to instruct the receiver how to decode each service. Here, lists of alternative frequencies and frequency schedules would be transmitted if appropriate.

Finally, the bulk of the signal conveys the MSC. With the limited bit-rate available within one 9- or 10-kHz channel, this would normally be used to carry one audio program, together with a modest stream of data. Nevertheless, there is a degree of flexibility, so the MSC may contain between one and four streams of data. Streams and services are distinguished as follows.

An audio service consists of one stream carrying audio, and optionally one stream carrying data. A data service consists of one stream carrying data.

The proposed system is based upon a multi-carrier modulation. It can be seen as a regular juxtaposition in the frequency domain of k elementary narrow-band subcarriers, each conveying a bit rate of d/k if d is the bit rate of the overall system. This choice comes from the robustness of such a waveform when a high spectral efficiency is necessary, while at the same time severe propagation conditions must be endured. Similar choices were made for higher-frequency terrestrial radio broadcasting (DAB, see Note 2) and TV broadcasting (DVB-T, see Note 3) in Europe, or more recently for the indoor high-data-rate wireless communications standard IEEE 802.11a at 5 GHz.

This leads to an optimum occupancy of the available bandwidth since the frequency spectrum of the signal is (almost) rectangular. This is possible because the signal that is conveyed by each carrier is orthogonal to the others. This property enables the signals conveyed by the different carriers to be separable, even if the narrow-band carrier spectra overlap. The adjustment of the system bandwidth is obtained via the modification of the number k of subcarriers, without any other change, especially at the receiver hardware and software level.

These definitions are used hereafter: The system conveys symbols that are located at known instants and frequencies. A carrier is the set of symbols that are located at the same frequency. A COFDM symbol is the set of symbols that are synchronously transmitted on all the used carriers. Hence, the number of symbols in a COFDM symbol is the number of used carriers. The COFDM symbols are grouped in a complete transmission frame that appears periodically with the same format. That is, the same repartition of reference and useful symbols. The notion of a transmission super-frame is also used to denote the group of transmission frames that starts with the SDC special group of COFDM symbols and contains exactly three transmission frames. As decoding the SDC is necessary to start decoding the whole bit stream, one can see that a standard delay to start reception is on the order of a super-frame.

Short access time is obtained by means of a few dedicated carriers that the receiver looks for in a first step for fast frequency synchronization. In addition, another set of carriers always contains the same information at a given instant—this corresponds to a constant known waveform/pattern which is used for pattern synchronization.

Maximum quality is obtained by the

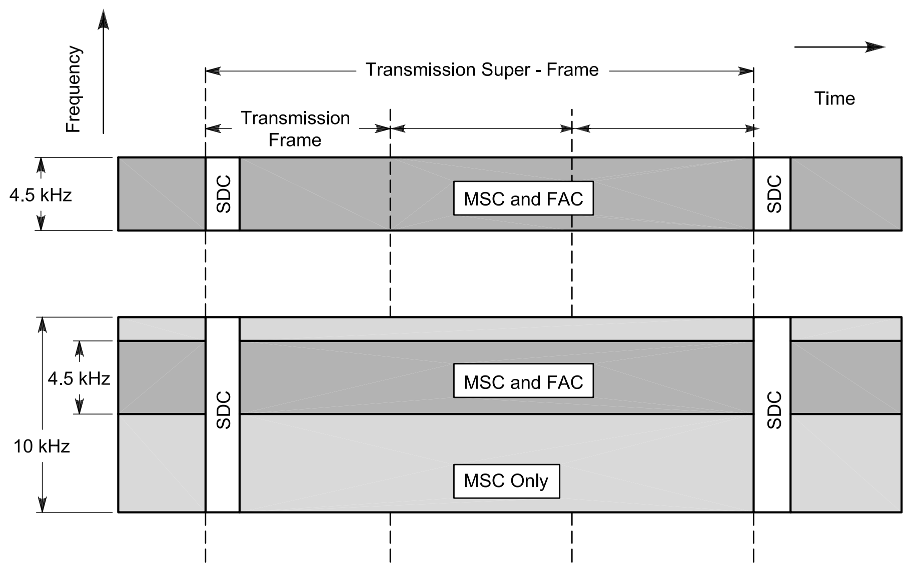


Fig 1—DRM time-frequency structure showing the various channel positions for 4.5 and 10-kHz bandwidth.

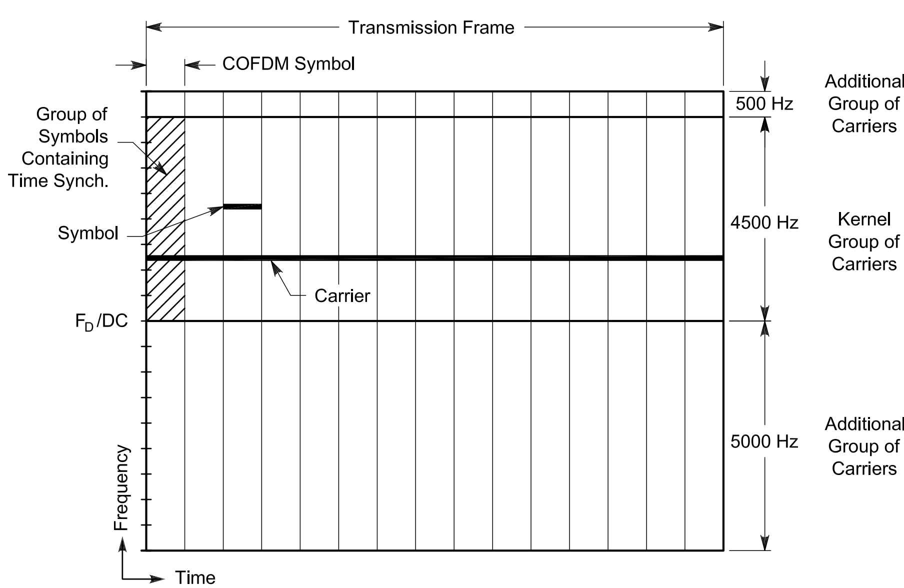


Fig 2—Symbol, frame and carrier definitions (example of 10-kHz overall bandwidth).

use of multilevel quadrature amplitude modulation (QAM). The proposed QAMs have levels of 4, 8, 16 or 64, the number of states being chosen as a function of the desired level of robustness. For this type of modulation, coherent demodulation is required; at each instant and at each frequency, it is necessary to estimate the complex gain of the transmission channel. To evaluate the channel response, some of the symbols—at predefined frequencies and instants—are sent with predefined amplitude and phase references so that the gain of the channel can be evaluated at any instant (using time interpolation) and any frequency (using frequency interpolation).

Robustness is achieved by the use of multilevel coding (MLC). An association of convolutional encoding and interleaving leading to an optimization of the transmission efficiency.[4, 5] In conjunction with coding, time and frequency interleaving are used, which have the effect of spreading perturbations (frequency-selective fading, flat fading, interference) on distant symbols, so that decoding is more efficient.

"Parametrability" is obtained by an incremental design of the broadcast signal. The signal contains a standard kernel group of carriers (occupying 4.5 kHz) that is common to all versions. The required bandwidth/bit rate is obtained by adding additional groups of subcarriers on either side of the kernel group. (See Fig 3.)

During the analog-to-digital transition phase, a small number of additional groups may be used. The unused spectrum will be occupied by an AM-compatible waveform, which can be received by classical AM receivers without any modification.

There is flexibility because the bit stream is divided into a main stream for conveying standard audio or audio plus still pictures or data and a system data stream with a much lower bit rate conveying additional data, such as program-associated data (PAD similar to RDS). With up-to-date audio encoders/decoders, the ratio of audio to pictures may be instantaneously variable in the main stream, and the significance of the data bit stream can vary as desired.

There is minimum disturbance to traditional AM users because the digital waveform has a flat spectrum so that, as received by an AM receiver, the digital signal sounds almost like white Gaussian noise. A pulse shaping of the transmitted symbols (that is, time windowing of the COFDM symbol) and/or additional output signal filtering can further decrease the disturbance of AM receivers in the adjacent channels.

Low complexity is inherent in the system. Since the signal can be seen as a number of elementary carriers uniformly spaced in frequency, the main digital signal processing is done by means of several fast Fourier transforms or inverse fast Fourier transforms (FFTs, IFFTs) which are known for their very efficient implementations. At the receiver, the FFT is equivalent to a large filter bank, each filter selecting only one subcarrier. The complexity level is only proportional to the occupied bandwidth, and it is independent of the channel quality.

Finally, ease of use of the system is obtained by automatic remote control of the receiver. Dedicated, highly protected symbols convey all the necessary configuration parameters: Any change in the transmission characteristics (bandwidth, coding, interleaving) is automatically taken into account by the receiver. These symbols are grouped in the FAC and the SDC channels.

DRM currently contains 4 modes:

- Mode A (ground wave): Gaussian channels, with minor fading adapted to LW and MW during daytime.
- Mode B (sky wave): Time and frequency-selective channels, with longer delay spread for SW and MW nighttime.
- Mode C (robust): Time and frequency-selective channels, with greater Doppler spread for bad SW channels.
- Mode D (extreme): Very robust mode, but with a reduced net bit rate (approximately half of the bit rate of mode A).

The first three modes are expected to cover most applications.

Mode A has a guard interval of $2^2/_3$ ms, together with a carrier spacing of $41^2/_3$ Hz. This is described as intended for "Gaussian channels, with minor fading," and is thus particularly suitable for local or national coverage at LF/MF, although it may also be useful in some longer-distance applications. The guard interval is sufficient for SFN operation.

Mode B can be described as intended for "Time and frequency selective channels, with longer delay spread." It has a guard interval of $5^1/_3$ ms, with a carrier spacing of $46^7/_8$ Hz and a higher density of pilots. The longer guard interval is intended to cope with greater multipath spread, as can be caused during multimode, multihop sky-wave propagation, while the greater carrier spacing and pilot density give greater tolerance to Doppler spread.

Mode C is devoted to "Fast varying time and frequency selective channels, with longer delay spread." It has a guard interval of $5^1/_3$ ms, with a carrier spacing of $68^2/_{11}$ Hz and the highest density of pilots. The long guard-interval is intended to cope with greater multipath spread, as can be caused during multimode, multihop sky-wave propagation, while the greatest carrier spacing and pilot density gives maximum tolerance to Doppler spread.

Each symbol is modulated independently from its neighbors by choosing a given point in a predefined constellation according to the value on an information word (2-6 bits). The reference symbols are transmitted with a predefined amplitude and phase. A reference symbol is used either for synchronization purposes or for channel complex-response estimation. A useful symbol must be demodulated and decoded in order to recover the original information it conveys.

The subcarriers are located at offsets from the center frequency that are multiples of $\Delta f = 1/T_u$, that is $41^2/_3$ Hz (system for ground wave) or $46^7/_8$ Hz (system for sky wave) or $68^2/_{11}$ (robust). By convention, in the complex baseband representation, the *k*th carrier is at an

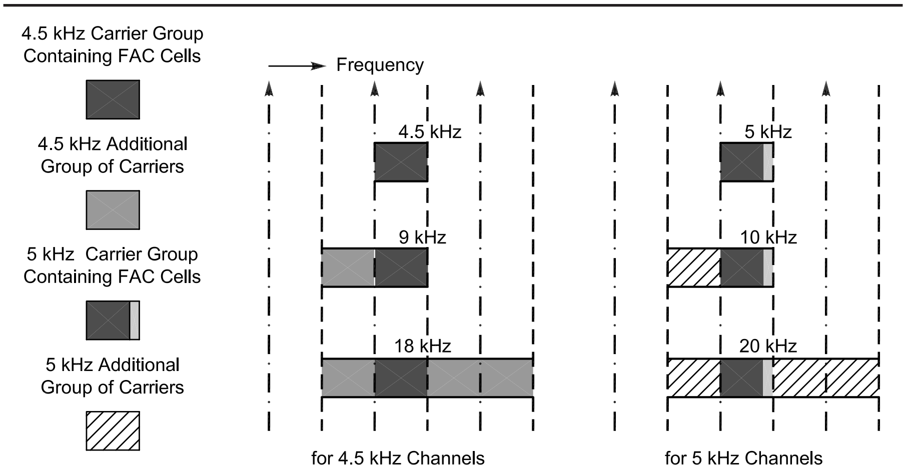


Fig 3—Various bandwidths possible.

offset of $k\Delta f$ from the dc component; k can be positive (to the "right" of the reference carrier) or negative (to the "left" of the reference carrier). The 0th carrier is dc. In the first two parameter sets the transmission frame contains 15 COFDM symbols; there are 20 COFDM symbols in the last one, leading to a constant common duration of 400 ms.

As already mentioned above, the carriers are grouped in a kernel group of carriers, which is common to all transmission modes. The group conveys all the reference symbols necessary for time and frequency synchronization, as well as the FAC symbols that describe the current mode. The kernel group is immediately above the carrier at frequency F_0 and does not contain it. Its bandwidth is exactly 4.5 kHz between the zeros of its frequency spectrum.

Additional groups of subcarriers (possibly none, in the 4.5-kHz version), the number of which is defined according to the desired bit rate and available bandwidth. An additional group does not contain any synchronization or FAC symbol. If there are additional groups below the carrier, their number is always such that the frequency spectrum of the signal is symmetrical around the carrier. The bandwidth of each additional group is exactly 1.5 kHz between the zeros of its frequency spectrum.

The source coding requirements are directly derived from the channel capacity. The capacity available for audio within a single 9- or 10-kHz channel is distinctly limited—at 20 to 25 kbps and perhaps as little as 10 kbps for some extremely unfavorable HF paths. This clearly represents a serious source-coding challenge for DRM, which expects to deliver good audio quality for both speech and music.

DRM has incorporated work and key technologies already done in the development of source coding elsewhere and fine-tuned it to this particular application. For coding most broadcast program material, an audio coder is needed to cope with the arbitrary mix of speech, music and incidental background sounds. For this purpose, DRM uses advanced audio coding (AAC) from the ISO MPEG-4 standard, supplemented by spectral band replication (SBR).

The SBR technique synthesizes the sounds that fall within the highest frequency octave. Sounds in this range are usually either:

1. noise-like (sibilance, percussion instruments such as shakers, brushed cymbals and so on), or

2. periodic and related to what

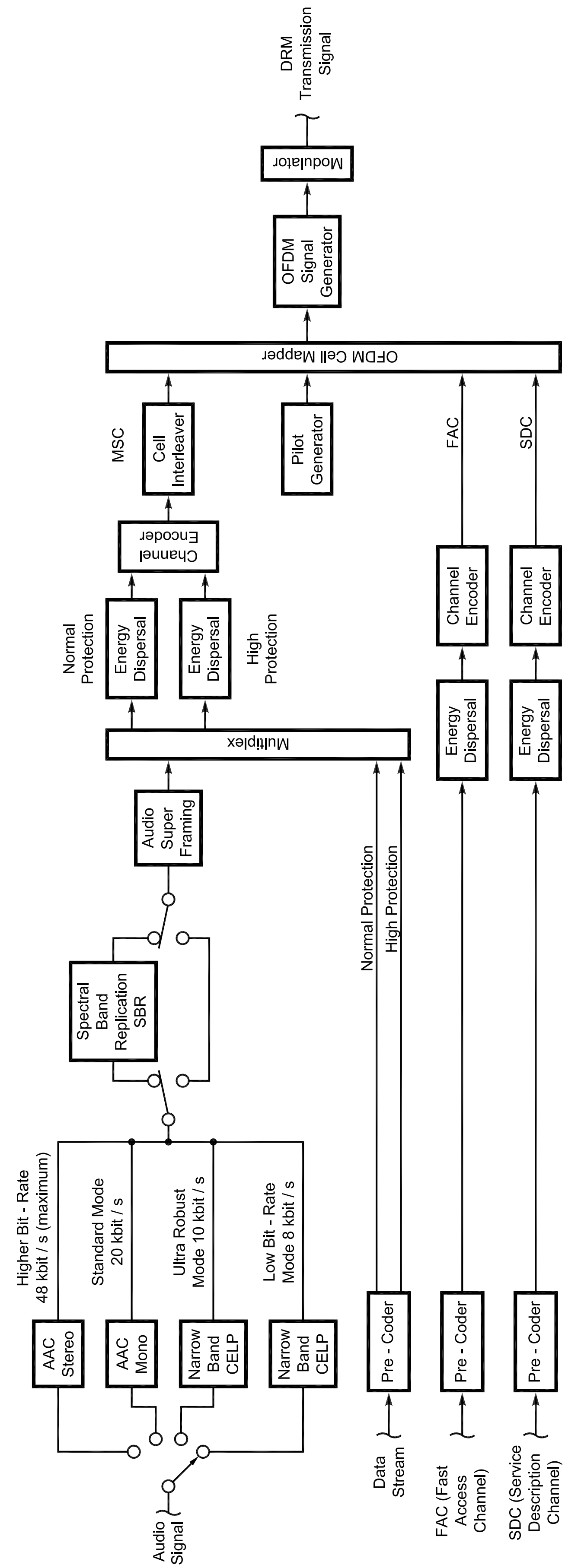


Fig 4—Complete DRM architecture for transmit side.

appears lower in the spectrum (overtones of instruments or voiced sounds).

At the sender, the highest-frequency band of the audio signal is examined to determine the spectral distribution and whether it falls into category 1 or 2 above. A small amount of side information is then prepared for transmission to help the decoder. The highest-frequency band is then removed before the remaining main band of the audio signal is passed to the AAC coder, which codes it in the conventional way.

At the receiver, the AAC decoder first decodes the main band of the audio signal. The SBR decoder then adds the synthetic upper band, helped by the instructions sent in the side information. Overtones are derived from the output of the AAC decoder, while noise-like sounds are synthesized using a noise generator with suitable spectral shaping.

An alternative for speech-only programming is to use a coder designed expressly for speech, in which case the bit-rate can be reduced much more than with a waveform audio coder, while retaining the same speech quality. Although this offers broadcasters further flexibility, there is, however, some doubt whether this approach would be used much in practice (apart from multilingual news broadcasting). Even speech-only broadcast material contains jingles, background sounds in interviews and so on, all of which can cause serious problems to speech coders. Figure 3 summarizes the various basic channel-content possibilities. Thus the complete system is given in Fig 4 (transmit side).

Adaptation to Amateur Radio Use

From the beginning, Thales has proposed to limit the kernel group of carriers to 3 kHz, thus enabling easily a 3-kHz mode for use in radio amateur HF or military applications.[6] So far, this mode has not been retained by the DRM consortium to keep the basic set of modes as limited as possible, thus lowering the development cost of the first version of the receiver chipsets.

Sensing the importance of such a mode, a demonstrator was developed anyway. Further modifications of the DRM system were applied while keeping the same overall structure:

- To limit the kernel group of carriers to less than 3 kHz.
- To adapt to push-to-talk mode, and other features of amateur transceivers.
- To include lower bit-rate vocoders (waveform coders are not usable at such low bit rates) imposed by the further reduction in available bandwidth (1200, 2400, 3200 bits/s voice coders are implemented in the demonstrator).
- To simplify the multiplex scheme and its associated description to reduce the bit rate for signals and have a minimum transmission delay (see above).
- To simplify intellectual-property issues.

The main similarities between the proposed system and the original DRM standard are the following.

Unmodified Features

- Symbol duration and guard times
- The positions of the three unequally spaced unmodulated carriers (pure tones), which help find rapidly the actual frequency shift of the received signal.
- The repartition and positions of the gain-reference symbols
- The modulation schemes (constellations)
- The interleaving principles (short and long interleaving)
- The convolutional codes—in practice, we use only the simplest one, for both vocoded audio and free user data. Its performance is, of course, the same as with other more complex encoding schemes.

Modified Features

- The SDC stream has been completely suppressed since it is no longer necessary. All the configuration information is located in the FAC only, so that the super-frame of DRM (three frames) now only contains one frame (latency reduction, software simplification).
- The FAC contains only 40 symbols (instead of 65), 26 useful bits (instead of 64) and is more protected (code rate 0.5 instead of 0.6). It contains the full set of parameters of the current transmission, as well as other information like the frame contents: A flag indicating the number of vocoder frames in the transmission frame, or the end-of-message (EOM) flag.
- The positions of the FAC cells and some frame-synchronization cells have been modified to fit in the reduced bandwidth.
- The whole original signal is shifted towards dc to fit in a bandwidth of 300-3000 Hz. The 3-kHz bound is always respected. The 300-Hz bound is only approximately respected because of the spacing between tones (333 Hz, 281 Hz and 273 Hz for modes A, B and C).
- Finally, each vocoder frame and each free data block is completed by an eight-bit CRC (parity check) to detect errors and minimize their impact on the audio quality.

In practice, the signal-processing software for transmission and reception are essentially the same as for the DRM system. The only changes are software simplifications and different primary data (constants or data arrays). The demonstrator was developed to show the highest quality achievable with current voice-compression techniques and the possibility to add new digital data services.

The voice codecs used are part of the Thales portfolio of coders (the HSX family for Harmonic Stochastic eXcitation). Thales Communications has been active in low-bit-rate voice codecs since 1968. In particular, the 800 bits/s Stanag (Standard NATO Agreement) and the ETSI Tetra coders (for Professional Mobile Radio, the European equivalent of mode-25 radios) came out of these laboratories.

The advantage of having coders from the same family is that they share a lot of common parts, allowing for simple switching of coders, reduced code size and thus simple maintenance or even hierarchical error-correcting schemes. Such schemes correspond to transmis-

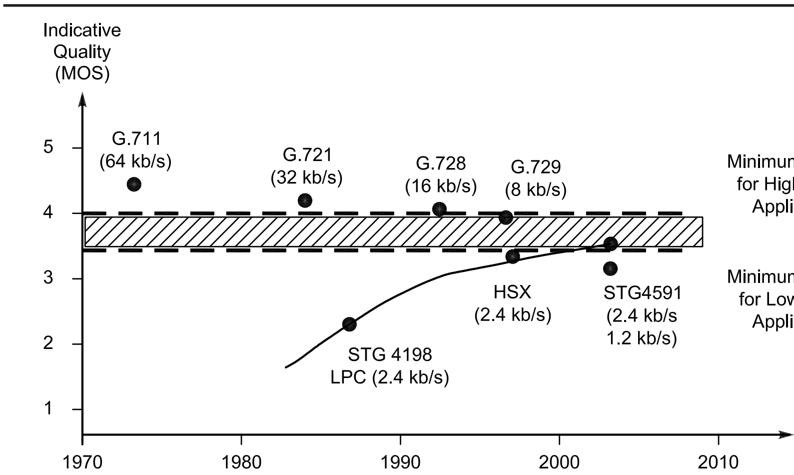


Fig 5—Evolution of voice coders quality through time.

sions where the additional information necessary to go from a given quality to a better one is less protected. So quality at reception varies depending on the ability of the modem to decode part or all the information, thus insuring at least the minimum quality and at best, the best quality when the channel permits it. Such a scheme is not implemented yet, but it could be added in a future version of the system.

Fig 5 shows the evolution through time of the quality of voice coders. The middle-range quality lying between MOS (mean opinion scores) of 3.5 and 4 are considered as the range acceptable for low- to medium-quality commercial applications. The latest 2400-bits/s coders have equivalent or better quality than older 4800-bits/s ACELP coders (adaptive code-excited linear prediction), while-1200 bit/s coders become of acceptable quality. New research in further rate reduction allows hope for 600 bits/s in a few years. Competition for a high-quality 4-kbits/s codec has been ongoing for several years without anything yet satisfying all the very stringent constraints (such as a very low delay). All the G.xxx codecs are standardized by the ITU, while the STG coders are military NATO standards—also named Stanags.

Other potential coders could have been:

• DVSI: proprietary format (IMBE and AMBE: 2 to 9.6 kbits/s) mainly chosen for satellite communications systems (Inmarsat), Iridium and APCO 25.

• MELP and EMELP: basis for the STG 4591 Stanag

The main reason for not choosing them is that the first is a proprietary, not easily available format, and that both come with IPR licensing fees and conditions, without any important gain in quality.

Our demonstrator software is 100% PC based for simple integration with standard off-the-shelf transceivers. A simple and intuitive man/machine interface (MMI) based upon the use of *LabWindows* software has been developed. The demonstrator has been fully integrated with Ten-Tec off-the-shelf transceivers. Transatlantic tests with ARRL are underway. A screen dump or the MMI in receive mode is given in Fig 6.

The six sub-windows show (left to right, top row first): the time values of the received signal, the spectrum of the received signal (filtered within 3 kHz), the reception level (estimation of the signal-to-noise ratio, SNR), the audio output signal, the received constellation (after time and frequency synchronization) and the estimated channel impulse response (here two paths are present with around 90° shift between them).

Other information are given such as the text message received, the mode detected, the estimation of the Doppler shift, the estimated instantaneous SNR.

The demonstrator is able to provide various transmission applications:

• Real-time voice transmissions, using a microphone/speaker and a sound card in the PC

• Transmit digital data (hand-written text or even complete files). The maximum available digital data rate is directly related to the vocoder bit rate and the current error correcting code and modulation.

• Test the modem performance by sending a test sequence for bit-error-rate estimation at the receiver

To test the demonstrator in a controlled environment, a real-time SW propagation simulator was included to show the insensitivity of the digital part of the system to multipath, whereas at the same time, an analogue AM equivalent is seriously perturbed. The simulator may generate a maximum of four paths. All the characteristics (amplitude, average power, frequency offset, Doppler spread, time spread) are adjustable in real time using a natural man/machine interface. Narrow-band interference may be added for system tests. All these impairments are added at the transmit side.

The DRM consortium normalized laboratory test conditions to get a fair comparison between various proponents. The channel model selected is described hereafter.

The approach is to use stochastic time-varying models with stationary statistics and define models for good, moderate and bad conditions by taking appropriate parameter values of the general model with stationary statistics. One of those models with adaptable parameters is the *wide-sense stationary uncorrelated scattering model* (WSSUS model). The justification for the stationary approach with different parameter sets is that results on real channels lead to BER curves between best and worst cases found in the simulation.

A tapped-delay-line model is then used for multipath generation.[7, 8] The time-variant tap weights are zero-mean, complex-valued stationary Gaussian random processes. The magnitudes are Rayleigh-distributed and the phases are uniformly distributed. For each weight, there is one stochastic process, characterized by its variance and power density spectrum (PDS). The variance is a measure for the average signal power, which can be received via this path and the PDS determines the average speed of variation in time. This type of channel model is known as the "Watterson" model.

A number quantifies the width of the PDS and this quantity itself (the width) is commonly referred to as the

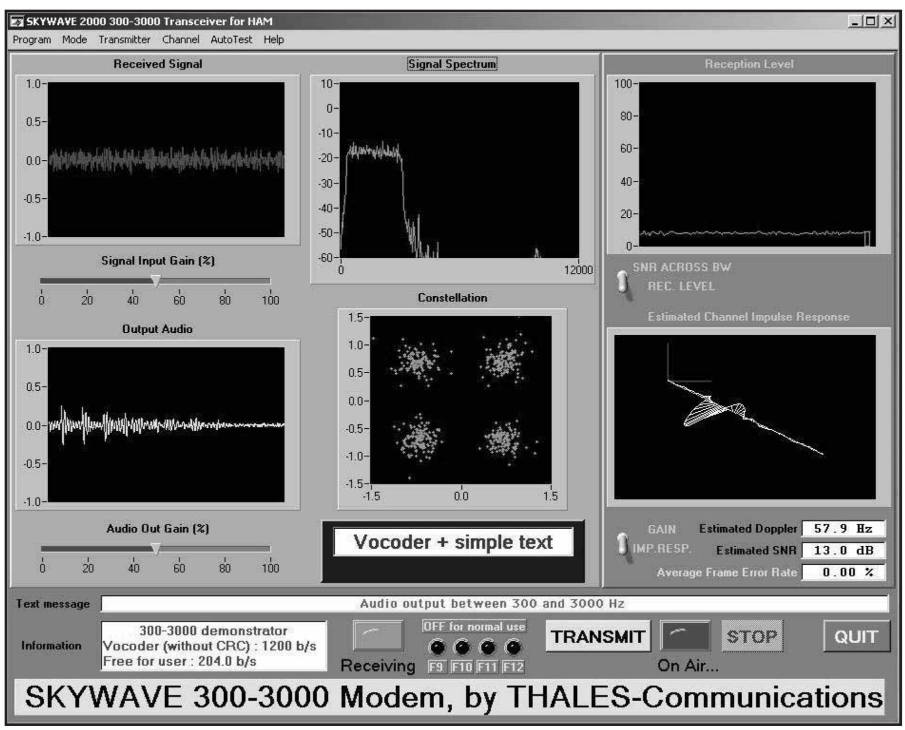


Fig 6—Demonstrator MMI while in receive mode.

Doppler spread of that path. There might be also a non-zero center frequency of the PDS, which can be interpreted as an average frequency shift (or Doppler shift).

It is common to assume a Gaussian amplitude statistic for the process, which is a reasonable assumption for an ionospheric channel. The stochastic processes belonging to every individual path then become Rayleigh processes. WSSUS does not define the shape of the PDS. For ionospheric channels, Watterson has shown a Gaussian shape to be a good assumption.[9, 10] The one-sided Doppler spread is then defined as the standard deviation (s) of the shape of the PDS.

Five channels are currently used by DRM:

- AWGN (additive white Gaussian noise): one path of constant amplitude (for ground-wave propagation).
- Ricean with delay: one constant path of unit amplitude, a second with half amplitude delayed 1 ms, with 0.1-Hz Doppler spread (for flat fading at MW and SW).
- US Consortium: four paths of amplitudes (1, 0.7, 0.5, 0.25), delay (0, 0.7 ms, 1.5 ms, 2.2 ms) Doppler spread (0.1 Hz, 0.5 Hz, 1 Hz, 2 Hz) and frequency shifts (0.1 Hz, 0.2 Hz, 0.5 Hz, 1 Hz).
- CCIR poor: two paths of equal amplitude delayed 2 ms and equal 1-Hz Doppler spread (SW propagation).
- Similar to channel 4, but with 4-ms delay and 2-Hz Doppler spread (bad SW propagation).

Conclusion

This paper presents a complete system proposal for Amateur Radio digital voice transmission derived from the DRM standard. It shows the current state of the art in COFDM modem technology. COFDM was chosen because of its very good robustness when a high spectral efficiency is desired, while at the same time severe propagation conditions must be endured. Such a system should start wide interest in digital voice, as the necessary computing power required is well within current PC technology reach (or DSP for future integration inside transceivers).

This system proposal is currently being tested (real-time demonstrator measurements) using a transatlantic link.

GLOSSARY

AAC	Advanced Audio Coding (MPEG-2/4)
ACELP	Adaptive Code Excited Linear Prediction
AFS	Alternative Frequency Switching
AM	Amplitude Modulation
ARIB	Association of Radio Industries and Businesses
AWGN	Additive White Gaussian Noise
BER	Bit-Error Rate
COFDM	Coded Orthogonal Frequency Division Multiplex
DAB	Digital Audio Broadcasting
DC	Direct Current
DRM	Digital Radio Mondiale
DVB	Digital Video Broadcasting
DVB-T	DVB - Terrestrial
EBU	European Broadcasting Union
ETSI	European Telecommunications Standard Institute
FAC	Fast Access Channel
FFT	Fast Fourier Transform
FM	Frequency Modulation
HF	High-Frequency
IBC	International Broadcasting Conference
IEC	International Electrotechnical Commission
IFFT	Inverse Fast Fourier Transforms
ISO	International Standard Office
ITU	International Telecommunication Union
ITU-R	ITU - Radiocommunication Sector
LF	Low-Frequency
LW	Long-Wave
MF	Medium Frequency
MLC	Multi-Level Coding
MMI	Man Machine Interface
MOS	Mean Opinion Scores
MPEG	Moving Picture Experts Group
MSC	Main Service Channel
MW	Medium Wave
NADIB	Narrow Band DIgital Broadcasting
OFDM	Orthogonal Frequency Division Multiplex
PAD	Program Associated Data
PDS	Power Density Spectrum
QAM	Quadrature Amplitude Modulation
RDS	Radio Data System
RF	Radio Fequency
SBR	Spectral Band Replication
SDC	Service Description Channel
SFN	Single Frequency Network
SNR	Signal-to-Noise Ratio
SSB	Single Side Band
SW	Short Wave
UEP	Unequal Error Protection

Notes

[1]C. J. Demeure, P.A. Laurent, "A new Modem for High Quality Sound Broadcasting at Short Waves," IEEE 4441, 7th Conference on Radio Systems and Techniques, Nottingham, UK; pp 50-54, July 1997.

[2]J. Stott, "DRM—Key Technical Features," *EBU Technical Review* #286, March 2001.

[3]M. Cronk, "DRM—Implementation Issues," *EBU Technical Review* #286, March 2001.

[4]H. IMAI and S. HIRAKAWA, "A New Multi-level Coding Method Using Error Correcting Codes," *IEEE Transactions on Information Theory*, Vol. IT-23, pp 371-377, May 1977.

[5]U. Wachsmann, R. Fisher, and J. Huber, "Multilevel Codes: Theoretical Concepts and Practical Design Rules," *IEEE Transactions on Information Theory*, Vol. IT-45, pp 1361-1391, July 1999.

[6]*DAB system: ETSI Norm*, ETS 300 401 ed. 2, May 1997

[7]*DVB-T system: ETSI Norm*, ETS 300 744, March 1997

[8]P. A. Laurent, C. Demeure, D. Castelain, B. Le Floch, "Thomson—CCETT Common Proposal for a Digital Audio Broadcasting System at Frequencies below 30 MHz," *NADIB Report*, June 1998.

[9]J. Lindner, D. Castelain, F. Nicolas, "Specification of an Ionospheric Channel Model for AM Radio Broadcasting Bands," *NADIB Report* , March 1998.

[10]Annex B to STANAG #4285, "Evaluation of Modems Employing the Stanag 4285 Waveform."

References

C. C. Watterson, J. R. Juposher and W. B. Bensema, "Experimental Verification of an Ionospheric Mode," *ESSA Technical Report*, ERL 112-ITS, 1969.

CCIR Recommendation 520, *Use of High Frequency Ionospheric Channel Simulators*.

RF

By Zack Lau, W1VT

Easy-to-Build 50:300 Ω and 50:450 Ω Transformers

Simple wire antennas often have high resistive impedances, typically from 200 to 500 Ω. It is often desirable to match these impedances to 50 Ω, to take advantage of the convenience of 50-Ω coaxial transmission line. If an impedance change is desired over a wide frequency range, a popular solution is to use a transformer with a ferrite core. The high permeability of ferrite allows less wire to be used for a given inductance, increasing the transformer bandwidth. Increasing the permeability too much, however, results in excessive loss—a balance needs to be struck for best performance. Transformers for 50:200 Ω are relatively common; I'll describe transformers for getting to 300 and 450 Ω.

When designing with these transformers, power-handling capacity can be a problem. Hams want transformers to work with a variety of antennas, such as center-fed dipoles, long wires and off-center-fed dipoles (OCFDs). OCFDs are particularly problematic, as it can be difficult to predict and remove unwanted shield currents.[1] A large shield current forced through a transformer can significantly increase power dissipation. Increasing the impedance presented to the shield current will help, but it is not always obvious that there is a problem.

How does one detect a hot ferrite core 50 feet in the air? One solution is a remote thermometer, but the consumer product I have seen is a potential source of 70-cm interference. It operates under Part 15 of the FCC rules. If the core gets hot enough, its impedance will change, but you may cook the core in the process. The impedance change is easily detected as a slow rise in SWR. Some cores will not survive this abuse—their permeability may be permanently changed—the SWR will not return to normal.

[1]Notes appear on page 58.

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

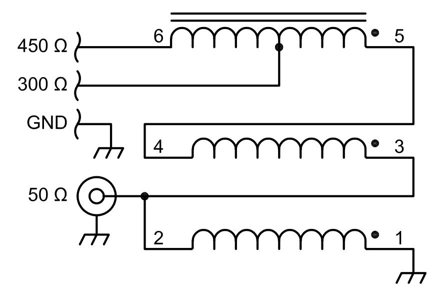


Fig 1—A schematic of the matching transformer. Wind 9 trifilar turns of #22 PVC-insulated hook-up wire on an FT-140-61 toroid ferrite core.

One solution is to use a large core as a separate shield choke. If the common-mode impedance of the choke is much higher than that of the transformer, most of the power will be lost in the choke, rather than the transformer. An advantage of this approach is that the transformer core can be optimized for low loss, while the shield core can be optimized for high impedance. The shield-choke core will typically have a higher permeability than the transformer core.

Fig 1 shows a broadband HF design. It has 9 trifilar turns of #22 AWG PVC-covered wire on an FT-140-61 core. It is tapped five turns from the high-impedance output to provide transformations to both 300 and 450 Ω. The return loss is better than 16 dB from 3.5 to 30 MHz (1.4:1 SWR). If only a transformation to 300 Ω is required, I'd suggest using #20 PVC covered wire, as this improves the return loss at lower frequencies. The insertion loss for a pair of 50:450-Ω transformers at 30 MHz is just 0.33 dB. It is even less over the rest of the HF amateur frequencies. My March 1995 RF column in *QEX* describes how to calculate a power rating based on toroid surface area and temperature rise.[2] For a 25°C temperature rise, the properly matched transformer should be able to handle 45 W. The 50:300-Ω transformers will handle even more—51 W. An FT-140 core will dissipate 1.66 W with a 25° temperature rise. The practical power rating is likely to be between these two extremes.

A smaller design uses 10 trifilar turns of #22 gauge PVC covered wire on an FT-114-61 core. The tap for the 300-Ω connection is 6 turns from the output. Tapping 4 turns from the output results in optimum output impedance around 360 Ω, rather than 300 Ω. It has more than 19-dB return loss from 7 to 30 MHz (1.25:1 SWR) and 12.5 dB return loss at 3.5 MHz (1.6:1 SWR). The worst-case insertion loss was at 3.5 MHz: 0.60 dB for a pair of 50:450-Ω transformers back to back. The FT-114 will dissipate 0.90 W with a 25°C temperature rise. The transformer should handle 13 W. The 50:300-Ω transformers had 0.45 dB loss at 3.5 MHz. They should handle 18 W.

Both designs are intended for use with a 1:1 choke balun. A design that works well from 1.8 to 54 MHz is 16 turns of RG-58C on an FT-240-43 core. I measured over 30 dB of isolation across the entire frequency range with a 50-Ω insertion-loss sweep setup. The coax is wound in a single layer, filling the core. A less expensive design that works well for the higher HF bands is 11 turns of RG-58C on an FT-140-43 core. It measures over 30 dB of isolation between 10 and 30 MHz. Type-43 ferrite can be used for broadband transformers if you don't mind the loss. A 6 turn trifilar design had nearly twice the loss. The loss is more consistent with frequency—it might be useful for amplifiers where loss is required for stability. An FT-240 core has 77 cm^2 of surface area—it will dissipate 3.7 W with a 25°C temperature rise.

Construction

The cores are wound with ordinary PVC hookup wire. The outer diameter of the #22 AWG wire is 0.058 inches, the #20 wire measured 0.062 inches. The thickness of the insulation is important—it helps determine the impedance of the windings. Thicker insulation spaces the windings more, raising the impedance. This makes the windings more useful for higher impedances. Conversely, you can use thinner insulation for lower-impedance windings. I used #20 enameled wire on a

50:200-Ω design (see Note 1).

Enameled #20 wire has a diameter of just 0.033 inches, much less than that of PVC coated hookup wire. This is counterintuitive to the idea that you always want the biggest wire that will fit on the core.

I strongly recommend winding and labeling each wire, one at a time. I cut scrap mailing labels into thin strips to make numbered markers for each wire. Assuming the labels don't fall off, it is a relatively easy task to connect the windings in the proper phase. However, before connecting the wires, I always separate the turns into groups of three. This helps to insure that the number of turns is correct. I also make sure that the wires do not cross over each other, except where the windings are connected.

Testing

I strongly recommend testing the transformers with an antenna analyzer to measure the SWR. Carbon-film or carbon-composition resistors (1/4-W, 300 Ω and 470 Ω) can be used as high-impedance loads. The SWR should be low from 3.5 to 30 MHz. A 4:1 or 5:1 SWR typically means that the wrong tap point is selected for the 50-Ω input. This technique is useful for designing your own transformers. If the optimum frequency range is too high, you need to add turns. *The upper frequency tends to be determined by the overall length of the wire.* If it is too low, remove turns. If the impedance of the winding is correct, removing half the turns should double the upper frequency. Since the impedance varies as the square of the turns ratio, halving the lower frequency only requires a 40% increase in turns. Thus, a 10-turn winding would be scaled to 5 or 14 turns.

The antenna analyzer is also useful for selecting tap points to accommodate output impedances other than 300 or 450 Ω. Obtain a carbon-film or carbon-composition resistor close to the value you need and use it to terminate the transformer at different tap points. Two parallel resistors generally have less series inductance than two in series. Remember this if you need to combine resistors to hit an exact resistance—such as two 910-Ω resistors in parallel to form a 455-Ω load.

You can measure insertion loss by hooking two transformers back to back, terminating them in a 50-Ω dummy load and measuring the power lost in the pair. Alternately, Jerry Sevick has suggested a "soak test," running 200 W or 1 kW for several minutes.[3] Then, with the power turned off, the cores are checked for a temperature rise.

The transformer designs presented should work quite well in QRP applications (5 W PEP). They may also be useful in higher power applications. A challenge for expert hams—how does one accurately determine power ratings for transformers and baluns that are fed through tuning networks?

Notes

[1] Z. Lau, W1VT, "Making Off-Center Fed Dipoles Work," (RF) Mar/Apr *QEX*, pp 55 to 58.

[2] This article also appears on pages 2-20 to 2-26 of *QRP Power* (Newington, Connecticut: ARRL, 1996) Order #5617.

[3] J. Sevick, W2FMI, *Transmission Line Transformers*, 2nd edition, (Newington, Connecticut: ARRL, 1990) pp 12-15.

Down East Microwave Inc.

We are your #1 source for 50MHz to 10GHz components, kits and assemblies for all your amateur radio and Satellite projects.

Transverters & Down Converters, Linear power amplifiers, Low Noise preamps, Loop Yagi and other antennas, Power dividers, coaxial components, hybrid power modules, relays, GaAsFET, PHEMT's, & FET's, MMIC's, mixers, chip components, and other hard to find items for small signal and low noise applications.

We can interface our transverters with most radios.

Please call, write or see our web site www.downeastmicrowave.com for our Catalog, detailed Product descriptions and interfacing details.

Down East Microwave Inc.
954 Rt. 519
Frenchtown, NJ 08825 USA
Tel. (908) 996-3584
Fax. (908) 996-3702

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

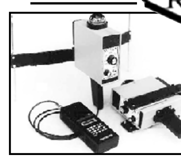
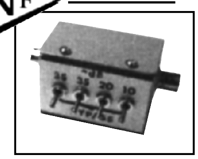
The *EZNEC 3.0* demo is the complete program, with on-line manual and all features, just limited in antenna complexity. It's free, and there's no time limit. Download it from the web site below.

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

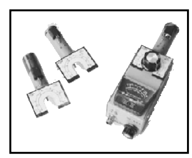
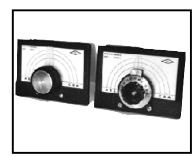
http://eznec.com

NATIONAL RF, INC.

VECTOR-FINDER
Handheld VHF direction finder. Uses any FM xcvr. Audible & LED display.
VF-142Q, 130-300 MHz $239.95
VF-142QM, 130-500 MHz $289.95

ATTENUATOR
Switchable, T-Pad Attenuator, 100 dB max - 10 dB min BNC connectors
AT-100, $89.95

DIP METER
Find the resonant frequency of tuned circuits or resonant networks—ie antennas.
NRM-2, with 1 coil set, $219.95
NRM-2D, with 3 coil sets (1.5-40 MHz), and Pelican case, $299.95
Additional coils (ranges between 400 kHz and 70 MHz avail.), $39.95 each

DIAL SCALES
The perfect finishing touch for your homebrew projects. 1/4-inch shaft couplings.
NPD-1, 3 3/4 × 2 3/4 inches 7:1 drive, $34.95
NPD-2, 5 1/8 × 3 5/8 inches 8:1 drive, $44.95
NPD-3, 5 1/8 × 3 5/8 inches 6:1 drive, $49.95

S/H Extra, CA add tax

NATIONAL RF, INC
7969 ENGINEER ROAD, #102
SAN DIEGO, CA 92111

858.565.1319 FAX 858.571.5909
www.NationalRF.com

Letters to the Editor

Theory of Intermodulation and Reciprocal Mixing: Practice, Definitions and Measurements in Devices and Systems, Part 1 (Nov/Dec 2002)

Doug:

Ulrich Rohde's article seems to have something missing. I realize this is a two-part article, but Part 1 seems to stop in mid-paragraph on page 15. I would expect steps 5, 6 and so forth concerning cascaded noise figure.

If things are missing, I hope it is not too late for the Jan/Feb 2003 issue. Otherwise, could we have errata posted to *ARRLWeb*? This very interesting article will require several readings.—*Jack Cavanagh, KB4XF, 223 N Randolph Rd, Fredericksburg, VA 22405-2927;* **cjcav@earthlink.net**

Doug,

This had to do with intercept point rather than noise figure, and since the length of a published paper is always an issue, this was omitted. The Friis noise-figure calculation should be well-known and is also covered in my receiver book.

By the way, there are errors in Eq 4. It should read as shown on this page.—*Ulrich Rohde, KA2WEU, Synergy Microwave Corporation, 201 McLean Blvd, Paterson, NJ 07504;* **ULRohde@aol.com**

A Software-Defined Radio for the Masses, Part 3 (Nov/Dec 2002)

Hi Doug,

I have been made aware that the download link for my Part 3 demonstration code said that it is "non-functional." That is incorrect because it is fully functional demonstration code. I downloaded it last night and ran it to verify that. Readers have also run the code successfully. There is some confusion about the SPL versus IPP libraries. The SPL license allows me to freely distribute the DLLs with my code. I would appreciate your help in getting the Web site corrected.—*Gerald Youngblood, AC5OG, 8900 Marybank Dr, Austin, TX 78750;* **gerald@sixthmarket.com**

Hi Gerald,

It is done! Doug Smith, KF6DX, ***kf6dx@arrl.org***

The ATR-2000: A Homemade, High-Performance HF Transceiver, Part 2 (May/Jun 2000)

Hello John,

I am building your design for the TRF noise receiver and noise gate for my homebrew transceiver. From looking at the circuit for the noise gate on p 40, it appears to me that the diagram is incorrect.

It looks to me like when there is no noise, the 0.022-µF capacitor is fully charged and the MPS2222 is turned on. By the way, I am using a 12-V dc supply, not 15 V. The way it is wired, there is approximately 3.42 V on the anodes of the diodes and 12 V on the cathodes. Therefore, the gate is reverse-biased, turning the gate off when it should be on.

When a noise pulse arrives on the noise gate input and discharges the 0.022-µF capacitor, the MPS2222 turns off and the voltage at the junction of the 390-Ω and 1 kΩ resistor goes to 12 V. Therefore, there is now 12 V on both the anodes and the cathodes. Thus, the diodes are turned on (just barely?) when they should be turned off.

It seems to me that we should either reverse the direction of the diodes—as we are dealing with an RF signal path—or move the top of the 33-µH inductor to the left side of the gate and the junction of the 1-kΩ/390-Ω resistor to the right side of the gate.

I'm not an electrical engineer so maybe I am missing something. I would appreciate your comments. In addition, if there are any other circuit errors, recent modifications or recommendations regarding the AT2-2000 other than those pointed out in regard to the receiver/transmit IF circuit, I would appreciate hearing about them.

I really appreciate the work you have done. The article on proper mixer termination was a great help. I have added diplexers to all the receiver input filters per your article and saw the IP_3 rise from +7.7 dBm to +22.7 dBm. I'm using using a +17-dBm LO to feed a doubly balanced mixer. Pretty amazing I would say. Thanks!

If you are interested in seeing a picture of my receiver, it is posted on my Web site at **www.qsl.net/ve7ca.**—*Markus Hansen, VE7CA, 674 St Ives Crescent, North Vancouver, BC V7N 2X3; Canada;* **ve7ca@rac.ca**

Hi Markus,

I'm sorry for the delay in answering your e-mail, but I've been away. In the published schematic, the diodes are drawn backwards. Current flows from the +15-V supply through the 33-µH inductor, the center tap of T3, the diodes, the center tap of T2, the 390-Ω resistor and the collector of the MPS2222. Several corrections have been published in the "Letters to the Editor" column of *QEX,* but I don't have copies here so I can't give you dates.—*John Stephensen, KD6OZH, 3064 E Brown Ave, Fresno, CA 93703-1229;* **kd6ozh@gte.net**

(Feedback about the ATR-2000 appeared in the May/Jun 2001 *QEX* on p 62. I've added a PDF of the feedback to ATR-2000.ZIP at **www.arrl.org/qexfiles/**.—*Bob Schetgen, KU7G,* QEX *Managing Editor)*

The Dirodyne: A New Radio Architecture? (Jul/Aug 2002)

Dear Rod,

[That is a] very interesting *QEX* article on the Dirodyne. I was somewhat confused by a statement you made relative to the Tayloe Detector. In the third paragraph of your article under the heading "The Tayloe Filter,"

$$y = k_1\left(A_1\cos\omega_1 t + A_2\cos\omega_2 t\right) + k_2\left[A_1^2\frac{1}{}\right.$$

$$+k_3\begin{cases}\left[A_1^3\left(\dfrac{\cos\omega_1 t}{2}+\dfrac{\cos\omega_1 t}{4}+\dfrac{\cos 3\omega_1 t}{4}\right)\right. \\ +A_1^2A_2\left[\dfrac{3\cos\omega_2 t}{2}+\dfrac{3\cos(2\omega_1+\omega_2}{4}\right. \\ +A_2^2A_1\left[\dfrac{3\cos\omega_1 t}{2}+\dfrac{3\cos(2\omega_2+\omega_1}{4}\right.\end{cases}$$

Equation 4 as per KA2WEU (see above)

you made the statement "… so there is no need for an audio phase-shift network to regenerate SSB at the same or at any other frequency…." What did you mean by that? Are you stating that this is another method of generating SSB without the use of an audio phase-shift network? If this is the case, could you please explain—perhaps mathematically—how a 90° phase shift at RF, IF or LO can provide a 90° phase shift at audio. The two frequencies (baseband and RF) are completely independent even in a discrete system and produce different phase shifts at baseband than would be experienced with an audio phase-shift network.

I think the point of the Tayloe Detector is that a gain term of two can be achieved by "differencing" the outputs of the out-of-phase components of the Tayloe Detector. This recovers energy that would be lost through detection by conventional means.

If I have missed the boat here please let me know. Thanks for the article and any further thoughts you have.—*Pete McNulty, WA1SOV, 8 Settlers Ln, Sandy Hook, CT 06482-1312;* **wa1sov@earthlink.net**

Hi Pete,

Normally, to generate SSB, there must exist either two or four audio paths in quadrature, as I'm sure you are aware. I simply mean that there are four quadrature baseband or audio paths from the output of the first Tayloe detector. Thus, there is no need to generate the four paths using a phase-shift network. In fact, it was this that made me suspect that a Tayloe filter was indeed a possibility; however, I had to build it to see if the system would behave as predicted, and it did.

I have the feeling that you are mathematically inclined. I am definitely not. This can be a burden sometimes, but a trace on an oscilloscope means more to me than a mathematical expression. The other engineer at my work is the opposite way, and he is—and I don't use the word lightly—a genius in almost anything he does, but especially complex math. I just get by and I struggle! My chief attribute is imagination, which I have in abundance; that makes the two of us a great team.

I am afraid I cannot comment about the differentiating the outputs of the out-of-phase components or the gain of two. I think the gain of two may come from the fact that the mixer does not offer a 50-Ω load to the source. Thus it can *appear* to have again of two, because of the generator's internal impedance not being loaded.

Actually, Gareth, the above-mentioned engineer, has a very good understanding of the Dirodyne, possibly better than mine! I can forward any math questions to Gareth if you have any.—*Rod Green, VK6KRG, 106 Roseberry St, Bedford, Perth WA-6052, AUSTRALIA;* **rodagreen@bigpond .com**

Dear Rod,

Thanks for your response to my inquiry. I was somewhat confused by the article because of the statement relating to generating SSB. This lead me to believe that your premise was that because the Tayloe generated four quadrature components, they could simply be summed and differenced to obtain a SSB signal without any additional phase-shift network at audio baseband. If this is what you meant, then I would disagree. While the Tayloe will obviously produce four phases in quadrature, they were produced in quadrature at a higher frequency. The baseband signal that is produced if the Tayloe is run at four times the IF input frequency, must be passed through additional 90° phase-shift networks at the baseband in order to produce an SSB signal by summing and differencing.

Perhaps I misread the article and your intent was not to say this at all. Anyway great work, keep it up. By the way, if you are interested, I have designed a DSP-driven exciter that can produce high-fidelity stereo independent or single sideband with or without carrier suppression. This was for a commercial application, but there may be some ham radio applications as well. It is in field testing at WBCQ on 9335 kHz and powers one of their 50-kW amplifiers. I recently gave a presentation at the New England Division ARRL convention in Boxboro, Massachusetts. Anyway, you can see why I have a keen interest in this stuff.—*Pete, WA1SOV*

Dear Doug,

I want to thank the *QEX* staff again for the excellent way you presented the Dirodyne article. I wonder if you could forward this with my thanks to Alvin Schmitt. He was kind enough to write with a cure for my birdie-noise problem *[see Letters, Nov / Dec 2002—Ed]*.

The cure was a great success! I can now receive signals down to –114 dBm for a 20-dB SINAD.—*Rod Green, VK6KRG*

Tower and Antenna Wind Loading as a Function of Height (Jul/Aug 2001)

Hi Frank,

I suppose that you have received a lot of mail about [your] subject. I looked over my tower statistics after reading [your article] and was made aware of an important factor: icing. If someone were to go by your calculations, they would be close, but only if they lived in Miami. Here in Germany, statistics must include an icing factor that varies with tower height. It comes out to an average of about 2.5 times the wind loading without ice.

For completeness, I think you should make readers aware of this. Thanks for a fine article!—*Steve Takacs, DJØIA, Uckerather Strasse 65, Hennef-Sieg Germany 53773;* **steve.takacs@tiscali.de**

Hello Steve,

Thank you very much for your comments about icing. Yes, you are absolutely correct. The icing factor should have been addressed in the "Assumptions" section of my article.—*Frank Travanty, 808 Pendleton CT, Waukesha WI 53188;* **w9jcc@juno.com**

Using the HP Z3801A GPS Frequency Standard (Nov/Dec 2002)

Doug,

I would like to start by saying that I enjoy *QEX* as possibly the only remaining Amateur Radio publication that still remembers that part of the purpose of the Amateur Radio Service is advancing the state of the art. While most of the ham world has been sitting around whining about the demise of CW, I have observed that we have a lot more appliance operators than we used to. I look forward to receiving [every] copy of *QEX*.

Having said that, I must express a matter of concern. I have noticed in the most current issue a note under a schematic stating *[p 53]* that Maxim would provide free samples via their Web page and/or 800 number. Over the years, it has become harder for those of us working in the electronics industry to retain our ability to access free samples. Many manufacturers are reluctant to supply them to all but the larger companies who will be purchasing tens of thousands of parts. For a small company, it can be problematic to get samples quickly and often we must resort to ordering from Digi-Key with the associated minimums. I realize that the author meant no harm, but I do believe that it is questionable editorial policy to include such a note. I am fairly sure that Maxim and other companies would not be happy if they were aware that the amateur community was abusing their sampling policy.

I believe that it is an abuse of the

policy of making free samples available to engineers involved in legitimate commercial development to request free samples for hobby purposes *unless* the company providing them is made aware that they are not going to be used for commercial purposes. Thanks and 73—*Ron Kumetz, N1WT, 1348 Peet Rd, Cornwall, VT 05753-9244;* **ron@senix.com**

Hi Ron,

Thanks for your note. We appreciate your kind words about QEX.

I am not aware that Maxim would be unhappy to provide samples to experimenters in the Amateur Radio Service, but I will check it out. We are a service, after all, and not just a hobby. Were we to present ourselves in that way, we might find quite a different reaction from manufacturers. I would think the kind of exposure provided in publications like QEX might actually encourage manufacturers to sample. Many of our readers are well placed in industry and academia in addition to being hams.

I personally have found it relatively easy to get samples using requests via various manufacturers' Web sites. Samples usually show up quite quickly and I don't get questions or follow-up sales calls. Just put ARRL after your name—if you are a member—and that may help things along. In the bad old days, a distributor or representative had to handle the requests and months sometimes went by before I saw samples, if I ever saw them at all.—Doug Smith, KF6DX

A Guy-Wire-Interaction Case Study (Nov/Dec 2002)

Doug,

Dick Weber's article brings to mind a way to prevent a problem in antenna installations. The basic steps are these:

1. Model the antenna and determine the high-current points on the tower and guys; currents less than about 5% of radiator current probably do no damage. If the antenna exists, look at the problem of getting to the point; and, if necessary, move to one that can be reached.

2. Insert a small reactance in the high-current or selected segment, and adjust its value until the pattern stabilizes; repeat for each operating frequency, leaving the other reactances in place.

3. Determine the number of ferrite beads needed to get the reactance. If the "spec" sheet doesn't give the equation for this, measure the inductance with a Q-meter, or jumper a single turn with a known capacitor and measure the resonant frequency with a "grid dipper." Up to 7- and possibly 10-MHz ferrites from old TV's are useful.

4. Install the ferrite—clamp-on for an old installation, or toroid type as a new guy is being assembled. Use silicon rubber to hold it in place and for weather protection.

An alternate for an existing antenna is to mount an RF current probe on a trolley to be pulled along a guy wire. See the *ARRL Antenna Handbook* for a design, although it may be easier to use a LED indicator, flashing faster as the current increases.—*R. P. (Bob) Haviland, W4MB, 1035 Green Acres Circle, N. Daytona Beach, FL, 32119;* **bobh@695online.com**

Brainteaser

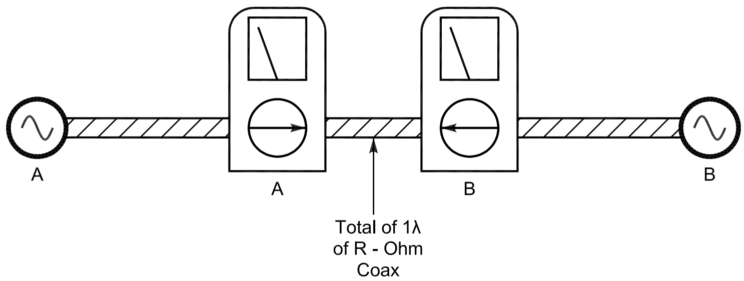


Fig 1—A test setup for the brainteaser.

Consider two RF generators connected to each other with a 1-λ line (Fig 1). The line's characteristic impedance is $R + j0$. The source impedance of generator A is $R + j0$ and that of generator B is $3R + j0$.

At $t = 0$, each generator delivers a signal to the line that produces a wave of power P traveling toward the other generator. The frequencies and phases of the generator signals are identical. In the steady state, what are the readings from two directional wattmeters inserted in the line to register the power in each direction?

Change generator B to have source impedance $R/3 + j0$ and keep everything else the same. What do the wattmeters read?

Next Issue in QEX/Communications Quarterly

In a follow-on to his recent article in *antenneX*, Dan Handelsman, N2DT, brings us a tutorial on intellectual property law in the US. He covers the basics of patents, copyrights and trademarks: constitutionally provided protection for the products of inventive minds. *antenneX* publisher Jack Stone was kind enough to grant us permission to reprint some of Dan's original material. Dan and *QEX* have added to it considerably.

Also in the Mar/Apr 2003 issue, several of our series on software radio continue. A lot is happening in that field. Both hardware and software are improving at a seemingly exponential rate.

QEX Subscription Order Card

ARRL
225 Main Street
Newington, CT 06111-1494 USA

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

In the US
☐ ARRL Member $24.00
☐ Non-Member $36.00
In the US by First Class mail
☐ ARRL Member $37.00
☐ Non-Member $49.00
Elsewhere by Surface Mail
(4-8 week delivery)
☐ ARRL Member $31.00
☐ Non-Member $43.00
Canada by Airmail
☐ ARRL Member $40.00
☐ Non-Member $52.00
Elsewhere by Airmail
☐ ARRL Member $59.00
☐ Non-Member $71.00

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 to ARRL

Charge:

☐ ☐ ☐ ☐

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.

06/01

HP® GPS RECEIVER DISCIPLINE CLOCK

Supply is limited!

$249
(Org. list $4,800)

Model: Z3801A® (Refurbished)

- Disseminating precise time and frequency (time acc. <1 µS)
- NIST traceable frequency reference, 10 MHz
- Manual and software included
- 48V dc/600 mA power supply and GPS antenna available

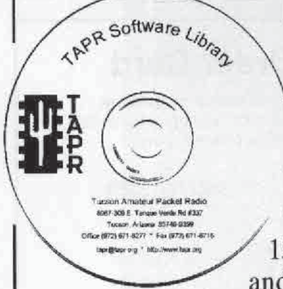
www.buylegacy.com **info@buylegacy.com**

San Marcos, CA
760-891-0810 • 800-276-1010 • Fax 760-891-0815

HP® and Z3801A® are registered trademarks of Hewlett Packard.

TAPR

***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 $20 worldwide.

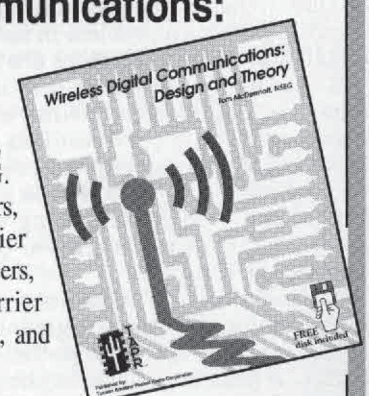


TAPR CD-ROM

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!

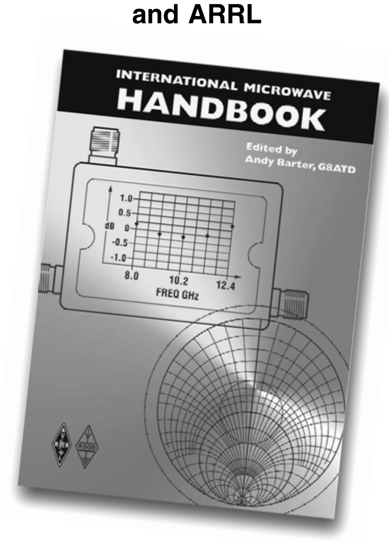


Tucson Amateur Packet Radio Corp.

Tucson Amateur Packet Radio
8987-309 E Tanque Verde Rd #337 • Tucson, Arizona • 85749-9399
Office: (972) 671-8277 • Fax (972) 671-8716 • Internet: tapr@tapr.org www.tapr.org
Non-Profit Research and Development Corporation

International Microwave Handbook

— Published by RSGB and ARRL



Edited by Andy Barter, G8ATD

Reference information and designs for the microwave experimenter: operating techniques; system analysis and propagation; microwave antennas; transmission lines and components; microwave semiconductors and valves; construction techniques; common equipment; test equipment; bands 1.3 GHz, 2.3 GHz, 3.4 GHz, 5.6 GHz, 10 GHz, 24 GHz, and above.

The precursor to this significant work was the three volume **Microwave Handbook** published by the RSGB in the late eighties and early nineties. This new book includes contributions from radio amateurs, organizations, publications and companies from around the world.

ARRL *The national association for* ***AMATEUR RADIO***

Order toll-free **1-888-277-5289** (US)

www.arrl.org/shop

ARRL Order No. 8739 — $39.95*

*shipping $7 US (UPS)/$9.00 International

tel: **860-594-0355** fax: **860-594-0303**

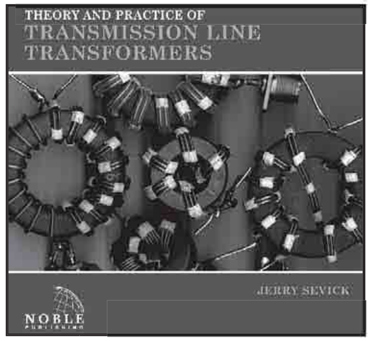
e-mail: **pubsales@arrl.org**

INTRODUCING

Jerry Sevick's NEW CD-ROM Course

Theory and Practice of Transmission Line Transformers

INSTRUCTIONAL CD-ROM SERIES



This tutorial introduces the theory and practice of transmission line transformers (TLT). In an innovative approach to the subject, Sevick divides TLTs into four classes: TLTs with ratios of 1:1, 1:4, less than 1:4 and greater than 1:4. The first two sections in this course cover 1:1 baluns and 1:4 baluns and ununs, as discussed by Guanella and Ruthroff. Additional sections review TLTs with ratios less than 1:4 and greater than 1:4, such as 1:6, 1:9, 1:12. The course concludes with a discussion of information on diode mixers and power combiners\splitters.

The material covered in this course provides incentive to investigate

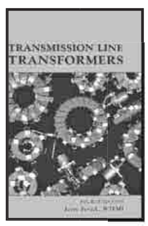
- **Tthe application of the latest TLT designs in broadband amplifiers and logic circuits**
- **The advancement of the theory and broadband response of the TLT**
- **The research in ferrite materials for high permeabilities with high bulk restivity**

2002, CD-ROM. ISBN 1-884932-33-9

NP-52 . . . . List Price $99.00 Introductory Offer $79.00*

***Please mention code 0352 when placing your order to take advantage of our introductory offer. Offer expires 03/15/2002.**

ALSO AVAILABLE



Transmission Line Transformers

Jerry Sevick

This fourth edition contains all the valuable information from earlier editions: the classic techniques of Guanella and Ruthroff and real transformers constructed and measured to established practical levels of bandwidth and loss performance that can be obtained with transmission transformer techniques. The new chapters cover •Transmission line transformer efficiency•Power combiners and mixer transforms •Equal-delay transforms.

2001, 4th edition, 312 pages. ISBN 1-884932-18-5

NP-9 . . . . $39.00

Order on-line! Visit us at www.noblepub.com

NOBLE PUBLISHING

Noble Publishing Corporation
630 Pinnacle Court
Norcross, GA 30071
USA

4 WAYS TO ORDER

CALL **770-449-6774** · FAX **770-448-2839**

E-MAIL **orders@noblepub.com**

INTERNET **www.noblepub.com**

Software Radio Now!

The RF Time Machine is a high-performance I-Q modulator and demodulator

- ***Receive*** a block of RF—up to 80 kHz wide—& ***record it*** to the audio tracks of a Hi-Fi VCR, to a computer through a sound card or to other recording devices.
- Hook to the antenna port of an HF RX & ***tune through*** the recorded portion of spectrum ***just like in real time!***
- Terrific for contest & DX analysis, radio demos, OO, EME & research.
- Assembled, $170; kit, $135 (+S/H). 1 Band Filter board & xtal included. 80, 40, 30, 20, 15 & 10 meters available.
- Daughter board now available for direct connection to a signal generator.

Rockwell Collin Mechanical Filters for the FT-817

- 300 and 500 Hz CW Filters
- 2.5 kHz SSB Filter

24 GHz Dielectric Rod Antenna

- Mounts Directly to a 24-GHz Gunnplexer
- Much Lighter than a Horn

Cylindrical Crystals

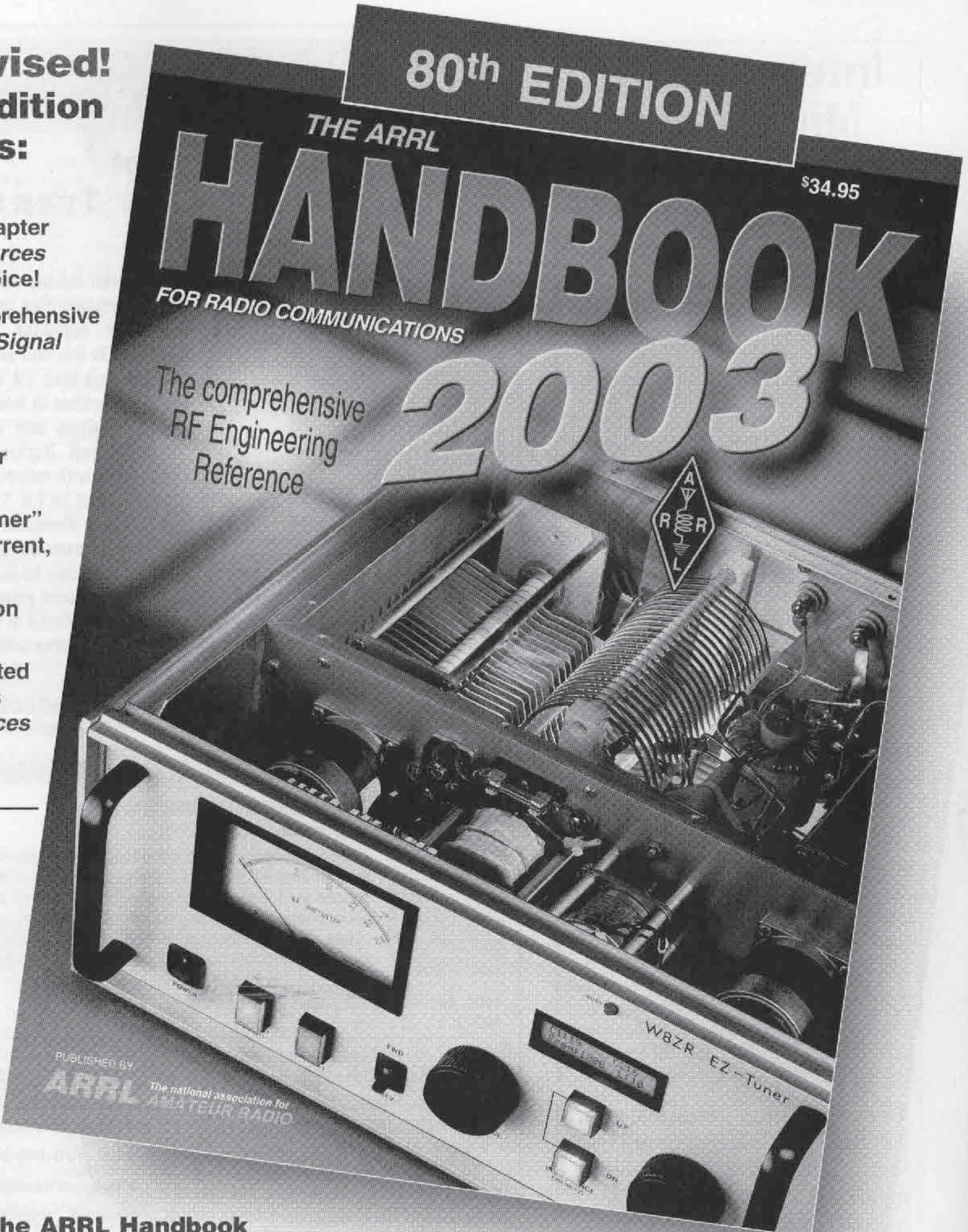
- 3560, 7040, 7190, 10106, 10125, 14025, 14060, 14200, 21026, 21060, 28060 kHz
- +/–100 PPM, 18 pF, 3 x 8 mm (3560 - 3 x 10 mm)

Expanded Spectrum Systems • 6807 Oakdale Dr • Tampa, FL 33610

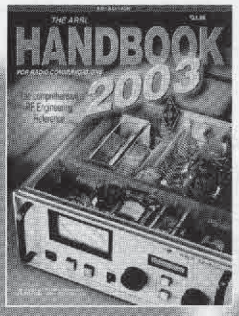
813-620-0062 • Fax 813-623-6142 • www@expandedspectrumsystems.com

Always revised! The 2003 edition includes:

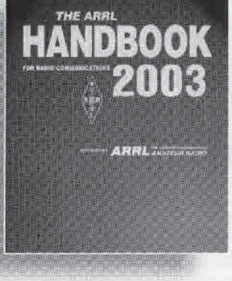
- An updated and comprehensive chapter on *Modulation Sources* including digital voice!
- A revised and comprehensive chapter on *Digital Signal Processing* (DSP) technology.
- A new high-power, automatic EZ-Tuner project by W8ZR.
- An "Ugly Transformer" project for high current, 120-V ac stations.
- A revised chapter on *Safety Practices.*
- A completely updated *Handbook* Address List in the *References* chapter.



THE STANDARD!



The ARRL Handbook for Radio Communications



Softcover—
ARRL Order No. 1921 .... **$34.95***
*plus shipping $7 US (UPS) /$9 International

Hardcover—
ARRL Order No. 1948 .... **$49.95***
*plus shipping $8 US (UPS) /$10 International

Available from ARRL publication dealers, everywhere!

Sales tax is required for orders shipped to CA, CT, VA, and Canada.
Prices subject to change without notice.

The ARRL Handbook CD for Radio Communications

Version 7.0—for *Windows* and Macintosh**
View, Search and Print from the entire 2003 edition book! CD-ROM.
ARRL Order No. 1956 .... **$39.95***

*plus shipping $5 US (UPS)/$7 International
**Some supplementary software utilities included—for *Windows* and DOS only.

Fully Searchable!

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 1/2003

New Volume!

The ARRL Antenna Compendium Volume 7

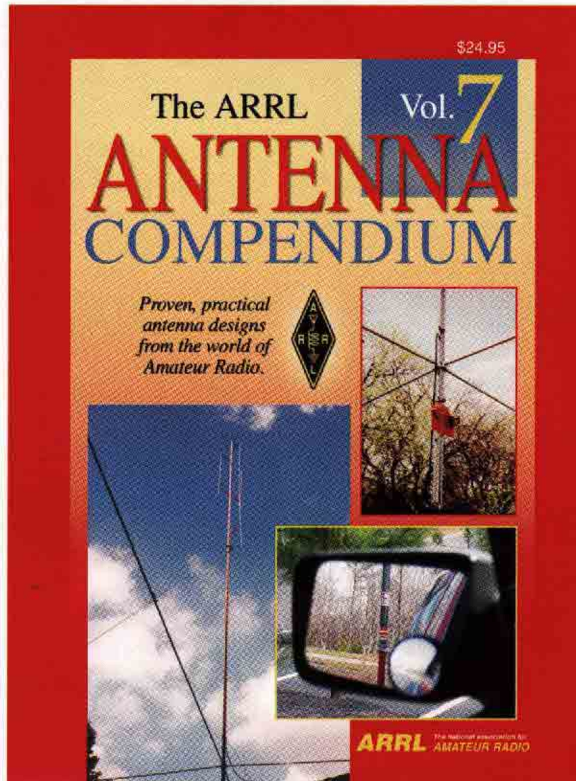
Proven, practical antenna designs from the world of Amateur Radio.

Hooray for antennas! This is the seventh in the very popular **ARRL Antenna Compendium** series. Inside, you'll find articles covering a very wide range of antenna-related topics:

- 30, 40, 80 and 160-Meter Antennas
- Measurements and Computations
- Mobile Antennas
- Multiband Antennas
- Practical Tips
- Propagation and Ground Effects
- Quad Antennas
- Special Antennas
- Stealth Antennas
- Tuners and Transmission Lines
- Vertical Antennas
- VHF/UHF Antennas
- Wire Antennas
- Yagi Antennas

And, this volume includes even more articles on low-band antennas and operating, and great designs for operating on the road —from cars, vans or motor homes.

Available from ARRL Dealers everywhere!

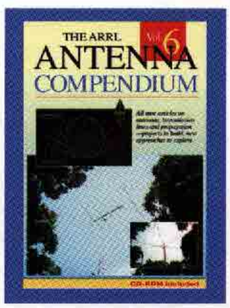


***The ARRL Antenna Compendium* —Volume 7**
ARRL Order No. 8608 **Only $24.95** plus shipping*

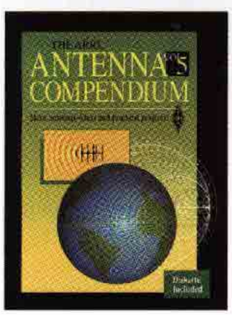
ARRL Antenna Compendiums

More antennas — ideas and practical projects

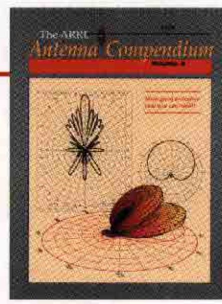
Volume 6
All new articles covering low-band antennas and operating, 10-meter designs, multiband antennas, propagation and terrain assessment. CD-ROM included with propagation prediction software!
#7431 .... **$22.95***—Includes software



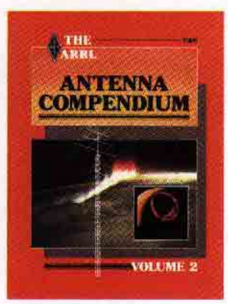
Volume 5
Enjoy excellent coverage of baluns, an HF beam from PVC, low-band Yagis, quads and verticals, curtain arrays, and more!
#5625 .... **$20***—Includes software



Volume 4 is out-of-print

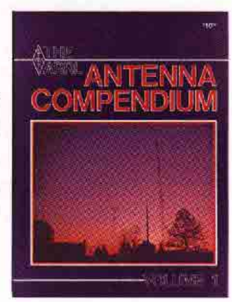


Volume 3
Quench your thirst for new antenna designs, from Allen's Log Periodic Loop Array to Zavrel's Triband Triangle. Discover a 12-meter quad, a discone, modeling with MININEC and VHF/UHF ray tracing.
#4017 .... **$14***



Volume 2
Covers a wide range of antenna types and related topics, including innovative verticals, an attic tri-bander, antenna modeling and propagation.
#2545 .... **$14***

Volume 1
The premiere volume includes articles on a multiband portable, quads and loops, baluns, the Smith Chart, and more.
#0194 .... **$10***



***Shipping:** US orders add $5 for one item, plus $1 for each additional item ($10 max.). International orders add $2.00 to US rate ($12.00 max.). US orders shipped via UPS.

Sales Tax is required for shipments to CT, VA, CA and Canada.

Order Toll Free
1-888-277-5289
www.arrl.org/shop
8 AM-8 PM Eastern time Mon.-Fri.

ARRL *The national association for AMATEUR RADIO*

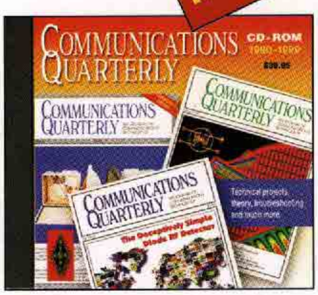
225 Main Street, Newington, CT 06111-1494
tel: 860-594-0355 fax: 860-594-0303

e-mail: pubsales@arrl.org

World Wide Web: http://www.arrl.org/

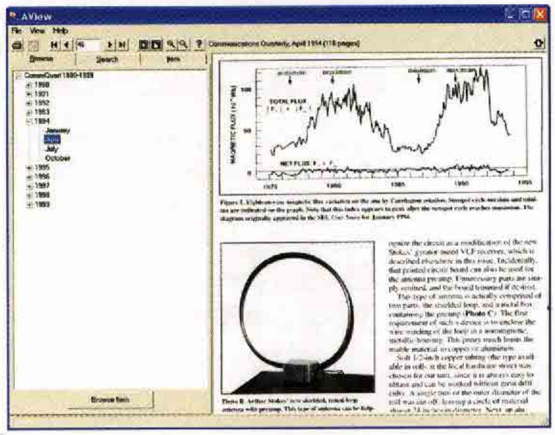
ARRL CD-ROM Collections

NEW!



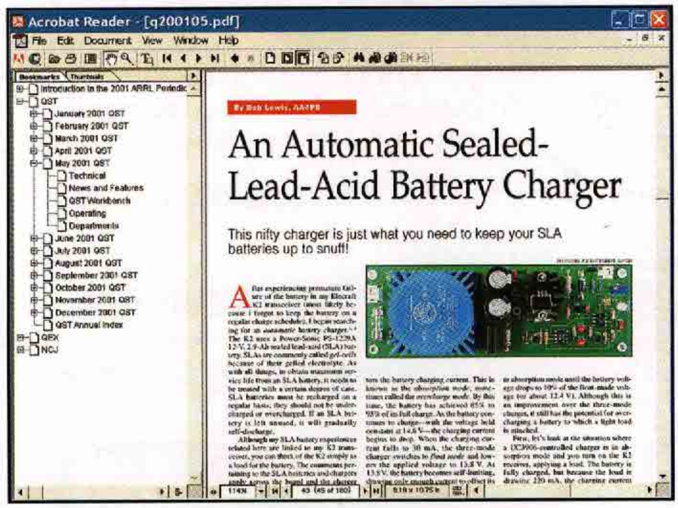
Communications Quarterly CD-ROM

This CD-ROM collection covers volumes of *Communications Quarterly* published from 1990-1999. Gain access to advanced technical topics in articles which cover transmitter, receiver and transceiver projects, theory, antennas, troubleshooting and much more. High quality black-and-white page scans can be read on your computer screen or printed. Quickly **search** for articles by title and author, **select** specific year and issue, and **browse** individual articles and columns. Requires Microsoft *Windows*™.



ARRL Periodicals on CD-ROM

are fully-searchable collections of popular ARRL journals. Every word and photo published throughout the year is included!



QST The official membership journal of ARRL

NCJ National Contest Journal

QEX Forum for Communications Experimenters

SEARCH the full text of every article by entering titles, call signs, names—almost any word. **SEE** every word, photo (including color images), drawing and table in technical and general-interest features, columns and product reviews, plus all advertisements. **PRINT** what you see, or copy it into other applications. **WEB LINKS** appearing in any article can be used to launch your existing Web browser to view additional information (feature available for versions beginning with 1996).

System Requirements: Microsoft *Windows*™. 1999, 2000, and 2001 editions support *Windows* and Macintosh systems, using the industry standard Adobe *Acrobat Reader*® (included).

ARRL CD-ROM Collections

ARRL Periodicals CD-ROM
(includes *QST, QEX, NCJ*)
#8632 Year 2001 **$19.95**
#8209 Year 2000 **$19.95**
#7881 Year 1999 **$19.95**
#7377 Year 1998 **$19.95**
#6729 Year 1997 **$19.95**
#6109 Year 1996 **$19.95**
#5579 Year 1995 **$19.95**

QST View CD-ROM
#7008 Years 1915-29 **$39.95**
#6710 Years 1930-39 **$39.95**

QST View CD-ROM (cont.)
#6648 Years 1940-49 **$39.95**
#6435 Years 1950-59 **$39.95**
#6443 Years 1960-64 **$39.95**
#6451 Years 1965-69 **$39.95**
#5781 Years 1970-74 **$39.95**
#5773 Years 1975-79 **$39.95**
#5765 Years 1980-84 **$39.95**
#5757 Years 1985-89 **$39.95**
#5749 Years 1990-94 **$39.95**
#8497 Years 1995-99 **$39.95**
#QSTV (all 12 sets) **$399**

NEW!
#8780 **Communications Quarterly CD-ROM** (1990-1999) **$39.95**

Ham Radio Magazine CD-ROM*
#8381 Years 1968-76 **$59.95**
#8403 Years 1977-83 **$59.95**
#8411 Years 1984-90 **$59.95**
#HRCD (all three sets) **$149.85**

#7660 **QEX Collection CD-ROM** (1981-1998) **$39.95**

#7733 **NCJ Collection CD-ROM** (1973-1998) **$39.95**

www.arrl.org/shop
1-888-277-5289 (US)

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

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.) Or, contact ARRL to locate a dealer. **Sales Tax** is required for orders shipped to CA, CT, VA, and Canada.