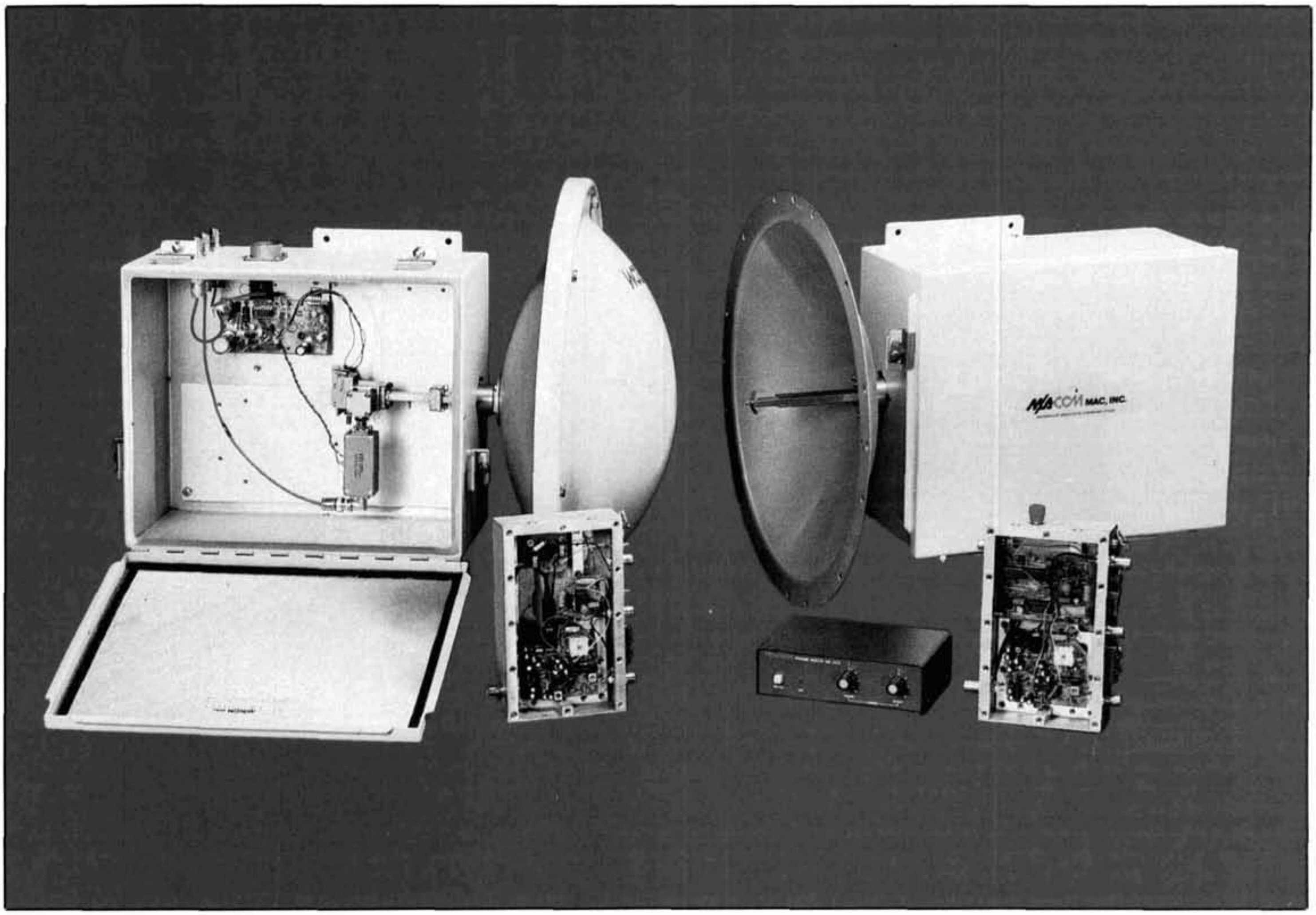
COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

Premier Issue $9.95 Fall 1990



- A Communications System Using Gunnplexer Transceivers
- Digital Signal Processing: Working in the Frequency Domain
- The Enhancement of HF Signals by Polarization Control
- High Dynamic Range Receiver
- Basic Concepts of Scattering Parameters
- Transmitter IMD Measurements on Utility-Grade Spectrum Analyzers
- Interface Your Computer to the "Poor Man's Spectrum Analyzer"
- Super Narrowband Techniques Equalize Power Inequity on 1750 Meters

ISSN 1053-9433

WORLD CLASS PERFORMANCE

WRTC Champions:
John Door KIAR
Doug Grant KIDG

Team Finland in operation:
Jukk Keshiaho OH8PF
Hannu Nieminen OHIXX

IC-765
HF TRANSCEIVER

Selected as the official radio of the World Radiosport Team Championships, a cultural exchange event of the 1990 Goodwill Games™, the IC-765 exemplifies ICOM's commitment to excellence in performance. The IC-765 incorporates the finest technology with the best designs to produce unbeatable HF operation for competitors worldwide.

The IC-765 sports: **Band Stacking Registers**. Each band's VFO's retain the last selected frequency, mode and filter choice when changing bands. Produces the equivalent of 20 VFO's; two per band. Great for multiband DX'ing! **99 Fully Tunable Memories.** Store frequency, mode and filter selections. **Direct Digital Synthesizer (DDS).** Assures ultra-fast PLL switching and lock-in for excellent PACKET, AMTOR and CW QSK operations.

Unlimited Operating Capability! The three step attenuator cuts multi-station overloads. Additional features include a **Built-in AC Supply, 100 percent duty cycle** for consistent high quality operation, **Fully Automatic Antenna Tuner, Iambic Keyer, Narrow 500Hz CW Filters** and **CW Pitch Control.**

The IC-765 general coverage receiver covers all bands and all modes.

Backed by ICOM's full **one-year warranty**, the IC-765's world class performance and superb reliability make it . . . the Choice of Champions!

CORPORATE HEADQUARTERS: ICOM America, Inc.
2380-116th Ave. N.E., Bellevue, WA 98004
CUSTOMER SERVICE HOTLINE (206) 454-7619
CUSTOMER SERVICE CENTERS:
3150 Premier Drive, Suite 126, Irving, TX 75063
1777 Phoenix Parkway, Suite 201, Atlanta, GA 30349
3071 - #5 Road, Unit 9, Richmond, B.C. V6X 2T4 Canada
2380-116th Ave. N.E., Bellevue, WA 98004

All stated specifications are subject to change without notice or obligation. All ICOM radios significantly exceed FCC regulations limiting spurious emissions. 765990

For a brochure on this or any other ICOM product, call our Toll-Free Literature Request Hotline 1-800-999-9877.

ICOM
First in Communications

KENWOOD

Triple Play!

TM-941A

TRI-BAND FM Transceiver

Kenwood brings you yet another breakthrough–the TM-941A *TRI-BAND* FM TRANSCEIVER. Now you can operate on *three* bands–144, 450, or 1200 MHz–with one radio! This rig even gives you full duplex, cross-band, triple-band repeat!

- **High power output.** 50 W on 144 MHz, 35 W on 450 MHz, and 10 W on 1200 MHz. (Selectable low power: 5 and 10 W, 1 W on 1200 MHz.)
- **Wide band receiver coverage.** 118-174, 438-450 (400-475 after modification), 1240-1300 (1210-1330 after modification) MHz. TX on Amateur bands only. Modifiable for MARS/CAP. Permits required.
- **CTCSS encode built-in.** 38 sub-tones selectable from the front panel.
- **Cross-band repeat function.** Selectable single or dual input! Off-set function on output, allows simplex to repeater repeat!
- **Simultaneous tri-band receive.** Individual volume and squelch controls help you "sort out" the signals.
- **Detachable front panel.** Use the optional PG-4K or PG-4L to mount the front panel remotely.
- **Selective calling option (DTU-2).** Selectively call a single station, or call a group with DTMF tones.
- **303 memory channels.** Store everything you need for efficient operation. All channels allow you to store "odd split" repeaters.
- **Versatile scanning functions.** Band scan, memory scan and programmed scan with carrier or time operated stop.
- **NEW! Auto memory scan.** Automatically memorizes a busy frequency while scanning the band!
- **Automatic repeater offset on 2 m.** Plus or minus 600 kHz for 144 MHz, ±5 MHz on 450 MHz, and ±12 or 20 MHz for 1200 MHz. (Manual offset for 450 and 1200 MHz.)
- **Fixed detect output.** For packet operators!
- **Multi-function DTMF mic supplied.**
- **Auto power off and time-out timer.**
- **4-step dimmer control.** Selectable 4-step dimmer control.
- **Three separate antenna and speaker connectors.** For maximum performance.

Optional Accessories:
• **DTU-2** Digital paging (DTMF) unit • **PG-4K, PG-4L** Front panel cable kits • **MC-45** Multi function mic. • **MB-11** Extra mounting bracket • **SP-41, SP-50B** External mobile speakers • **PG-3B** DC line noise filter • **PS-430** Power supply • **PG-2N** DC power cable • **TSU-7** CTCSS decode unit.

Complete service manuals are available for all Kenwood transceivers and most accessories.
Specifications, features, and prices are subject to change without notice or obligation.

KENWOOD U.S.A. CORPORATION
COMMUNICATIONS & TEST EQUIPMENT GROUP
P.O. BOX 22745, 2201 E. Dominguez Street
Long Beach, CA 90801-5745
KENWOOD ELECTRONICS CANADA INC.
P.O. BOX 1075, 959 Gana Court
Mississauga, Ontario, Canada L4T 4C2

KENWOOD

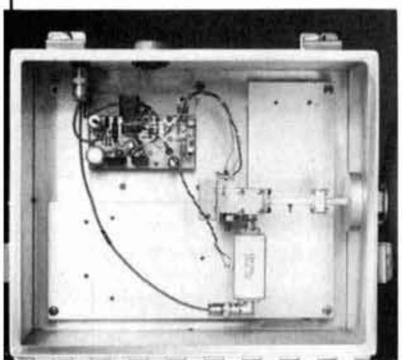
...pacesetter in Amateur Radio

COMMUNICATIONS QUARTERLY

THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

CONTENTS

Premier Issue November 1990



EDITORIAL

Thank You for Being a Charter Subscriber!

The ability to express oneself in an open forum like a magazine is an opportunity that many never have. I'm lucky, however, as *COMMUNICATIONS QUARTERLY* gives me an outlet to share my opinions with you. This month, in lieu of a specific topic, I want to talk about what has been going on here in Greenville in recent months.

You're holding the result of almost six months of mind numbing, tedious, and difficult work from a number of very dedicated people.

But before I get ahead of myself, I'd like to tell you who WE are, what WE want to be, and where WE want to go. I emphasized the WE because *COMMUNICATIONS QUARTERLY* is written for you, the technically minded Amateur. Amateur Radio needs a technical publication – one that dedicates itself to the pursuit of excellence in electronics.

COMMUNICATIONS QUARTERLY will cover a broad spectrum of articles in the very diverse field of electronics, from sophisticated antennas to the latest in zener diode applications. You, our reader, want to be challenged by thought provoking ideas. You want to learn how to apply sophisticated techniques in both your Amateur and professional activities.

This is what you're going to get.

A number of authors have already contacted us with suggestions, ideas, and articles for inclusion in upcoming issues. We've been carefully screening manuscripts and accepting only those that meet our rigid standards. It's not an easy task, though. In fact, each article may have been reviewed by three different people before an acceptance letter is sent to the author. This what we're doing to establish and ensure a reputation for running only the very best technical papers.

Looking over projected articles for upcoming issues, I can see that, while some may bemoan the lack of technical credibility in some areas of the Amateur service, it's alive and well in the pages of *COMMUNICATIONS QUARTERLY*. You'll be seeing articles on subjects like direct digital synthesis, high speed data communications, computer control and design, to name just a few. You'll be able to explore state-of-the-art electronics with a unique view to Amateur and professional applications.

I'd like to thank Dick Ross of CQ Communications for giving Terry and me the opportunity to bring you this first of many issues of *COMMUNICATIONS QUARTERLY*. Neither of us have ever done a start up before and it has been a real education. Late, sleepless nights have left us both tired. But we're pleased with what you now hold in your hands. We're convinced that the idea is sound and that there are enough readers to make *COMMUNICATIONS QUARTERLY* a viable publication. Response to our subscription mailing has more than borne that out. In fact, the response has been nothing short of outstanding. We're flattered!

Bob Wilson, WA1TKH, Peter Bertini, K1ZJH, Alf Wilson, W6NIF, and all the members of our editorial review board have made major contributions so all this could happen. Their knowledge of technical matters and editorial skill contribute greatly to the readability of *COMMUNICATIONS QUARTERLY*. I also want to thank our "unsung" heros. Their contributions are immeasurable – and both Terry and I appreciate all they have done.

Finally, I want to thank Terry for her effort. Her skill as an editor is seen on each page of every article. Attention to detail and a willingness to invest time in ensuring each article says exactly what the author intended it to, are an integral part in making the concept of *COMMUNICATIONS QUARTERLY* work. She has seen to it that this has all been done.

Please send us your thoughts, comments, ideas, and manuscripts. Prevail on those you know who might be able to write for us. Make us aware of new ideas, trends, and techniques. Give us the opportunity to share them with the other readers of *COMMUNICATIONS QUARTERLY*. In turn, we'll give you the best Amateur Radio technical journal being published today.

Thanks for the support. This is going to be fun!

Craig Clark, NX1G
Associate Publisher

A publication of
CQ Communications, Inc.
76 North Broadway
Hicksville, NY 11801-USA

Editorial Offices: Main Street, Greenville, NH 03048. Telephone: (603) 878-1441. FAX: (603) 878-1951.

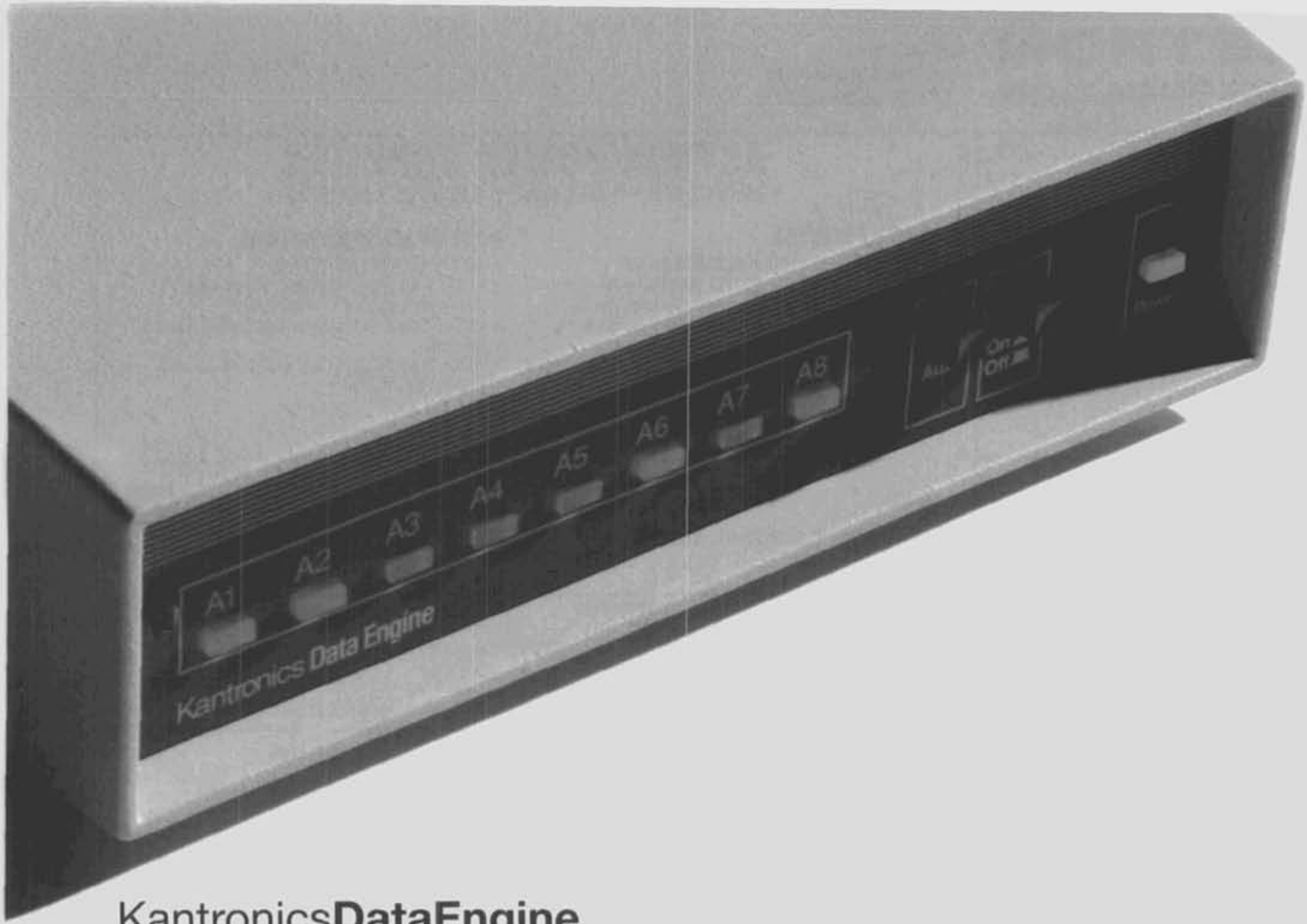
Business Offices: 76 North Broadway, Hicksville, NY 11801. Telephone: (516) 681-2922. FAX: (516) 681-2926.

Communications Quarterly is published four times a year by *CQ Communications, Inc.* Subscription prices: Domestic – one year $29.95; Foreign – $39.95.

Communications Quarterly does not assume responsibility for unsolicited manuscripts. Allow six weeks for change of address.

Postmaster: Please send change of address to Communications Quarterly, CQ Communications, Inc., 76 North Broadway, Hicksville, NY 11801.

ISSN 1053-9433



Kantronics**DataEngine**

If you'd like to run fast packet today, perhaps experiment with faster modes too, yet be compatible with today's 1200 baud system, then install a DataEngine at your station or node. The Kantronics DataEngine runs 1200, 2400, and 9600 baud packet with matched plug-in modems and appropriate transceivers now. For example, you can install a DE9600 modem in the DataEngine, attach a Kantronics DVR2-2 9600 baud data ready transceiver and run 9600 on 2-meters! Most off-the-shelf transceivers are not 9600 ready.

Better yet, you can add a free G8BPQ networking EPROM (download from our factory TELCO BBS) and run the 9600 combination as a node! Further, with the BPQ code, you can run the DataEngine in multi-drop KISS mode, attaching a KPC-2 or KPC-4 to the RS-232 port, creating a three or four port node!

Or, plug in a DE9600 or DE19200 modem and couple the DataEngine with the Kantronic's D4-10 data ready transceiver for 440 MHz high speed packet operation. (The D4-10 shall be released in early 1991).

Or, if you're interested in the DX spotting network or in transferring computer files to a buddy at 2400 baud, plug in the DE2400 modem and couple the DataEngine to any off-the-shelf transceiver. The 2400 baud QPSK modem is compatible with existing narrow band FM rigs and matches the industry standard Kantronics KPC-2400 modem.

Or, roll your own, attaching an experimental modem(s) via the disconnect headers! You might want to work satellites, play with a new form of modulation, or design your own HF modem. The disconnect headers on the DE pc board leave room for two internal or external modems. Experimenters already report interfacing the DataEngine with a PSK modem for working satellites. To aid in this fun process, you can order the DE Developer's Manual.

After all, the DataEngine is designed with an open architecture. It's a dual port, full duplex TNC with 16-bit V40 microprocessor running at 10 MHz, 85C30 communications controller, capable of speeds to 56KB per port. It comes with 64K of EPROM and 64K of RAM, and has socket space for up to .5 megabytes of EPROM and .5 megabytes of RAM!

Even better, the DataEngine comes factory stock with a DE1200 modem already plugged in and an end-user EPROM which supports terminal, KISS, HOST and BBS modes. It's ready to go on existing 1200 baud channels! The host mode enables use of sophisticated terminal programs, such as the DataEngineer, developed for the DataEngine, including windows/split screen etc. Or again, dial your own terminal program! In effect the DataEngine is "developer and user ready."

Specs: size 1-3/4"x6"x9", weight 2-1/2 lbs, power requirements nominally 12VDC at 150 ma. Input sensitivity 20 mvpp, Audio output drive jumper selectable from 10mvpp to 2vpp.

Options include: DE19200, DE9600, DE2400 and DE1200 modems, plus a developer's manual for modem and protocol experimenters. Plus more modems from Kantronics on the way!

For detailed specifications contact Kantronics. The Kantronics DataEngine, the TNC that can run fast today, and the platform to develop the next generation of advanced TNC applications.

Kantronics 1202 E. 23rd St., Lawrence, KS 66046, 913.842.7745 TELCO BBS 913.842.4678

ASTRON CORPORATION

9 Autry
Irvine, CA 92718
(714) 458-7277

MODEL VS-50M

ASTRON POWER SUPPLIES

• HEAVY DUTY • HIGH QUALITY • RUGGED • RELIABLE •

SPECIAL FEATURES

- SOLID STATE ELECTRONICALLY REGULATED
- FOLD-BACK CURRENT LIMITING Protects Power Supply from excessive current & continuous shorted output
- CROWBAR OVER VOLTAGE PROTECTION on all Models except RS-3A, RS-4A, RS-5A, RS-4L, RS-5L
- MAINTAIN REGULATION & LOW RIPPLE at low line input Voltage
- HEAVY DUTY HEAT SINK • CHASSIS MOUNT FUSE
- THREE CONDUCTOR POWER CORD except for RS-3A
- ONE YEAR WARRANTY • MADE IN U.S.A.

PERFORMANCE SPECIFICATIONS

- INPUT VOLTAGE: 105-125 VAC
- OUTPUT VOLTAGE: 13.8 VDC ± 0.05 volts (Internally Adjustable: 11-15 VDC)
- RIPPLE Less than 5mv peak to peak (full load & low line)
- All units available in 220 VAC input voltage (except for SL-11A)

SL SERIES

• LOW PROFILE POWER SUPPLY

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
SL-11A	•	•	7	11	2¾ x 7⅝ x 9¾	11

RS-L SERIES

• POWER SUPPLIES WITH BUILT IN CIGARETTE LIGHTER RECEPTACLE

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-4L	3	4	3½ x 6⅛ x 7¼	6
RS-5L	4	5	3½ x 6⅛ x 7¼	7

RM SERIES

MODEL RM-35M

• 19" RACK MOUNT POWER SUPPLIES

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
RM-60A	50	55	7 × 19 × 12½	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50
RM-60M	50	55	7 × 19 × 12½	60

RS-A SERIES

MODEL RS-7A

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A		•	2.5	3	3 × 4¾ × 5¾	4
RS-4A	•	•	3	4	3¾ × 6½ × 9	5
RS-5A		•	4	5	3½ × 6⅛ × 7¼	7
RS-7A	•	•	5	7	3¾ × 6½ × 9	9
RS-7B	•	•	5	7	4 × 7½ × 10¾	10
RS-10A	•	•	7.5	10	4 × 7½ × 10¾	11
RS-12A	•	•	9	12	4½ × 8 × 9	13
RS-12B		•	9	12	4 × 7½ × 10¾	13
RS-20A	•	•	16	20	5 × 9 × 10½	18
RS-35A	•	•	25	35	5 × 11 × 11	27
RS-50A	•		37	50	6 × 13¾ × 11	46

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46

VS-M AND VRM-M SERIES

MODEL VS-35M

• Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES

MODEL RS-12S

• Built in speaker

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	•	•	5	7	4 × 7½ × 10¾	10
RS-10S	•	•	7.5	10	4 × 7½ × 10¾	12
RS-12S	•	•	9	12	4½ × 8 × 9	13
RS-20S	•	•	16	20	5 × 9 × 10½	18

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

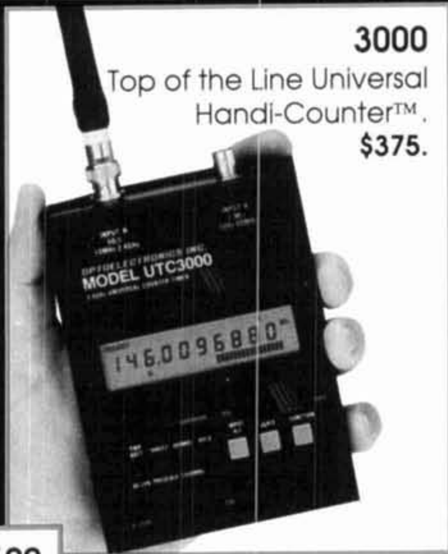
OFF THE BENCH AND RUNNING

When *OPTOELECTRONICS* took frequency counters off the bench we created a whole new dimension - Frequency Finding.

Our Handi-Counters™ make Frequency Finding a reality by allowing you to pick-up radio transmitters at the maximum possible distance.

Monitoring Enthusiast, Security Specialists, Broadcast and Design Engineers, Two Way and Amateur Radio Operators and Service Technicians have all made the move. ***Shouldn't you?***

3000
Top of the Line Universal Handi-Counter™.
$375.



2600
Super Sensitive RF Frequency Finder.
$325.

2210A
Full Range Pocket Size LED.
$239.

1300H/A
Low Cost Ultra Sensitive (HF, UHF, VHF).
$179.

8030
Bench Portable (Fits in an attache case) with *ALL* the Handi-Counter™ Features plus *More*. **$579.**

Model	Range *Low* *High*	Display *8 Digit LED*	Display *10 Digit LCD*	Signal Strength Bargraph	Universal Counter	TCXO *Option*
8030	10Hz 2.4GHz		•	•	•	± .1ppm add $125.
3000	10Hz 2.4GHz		•	•	•	± .2ppm add $ 80.
2600	1MHz 2.4GHz		•	•	•	± .2ppm add $ 80.
2210A	10Hz 2.4GHz	•				± .5ppm add $ 75.
1300H/A	1MHz 1.3GHz	•				± .5ppm add $ 75.

Sensitivity: <1 to <10mV typical. NiCads & AC Charger/Adapter included. Carry Case, Antennas and Probes extra. One year parts & labor warranty on all products.

OPTOELECTRONICS

Toll Free Order Line:
1-800-327-5912
In Florida call (305)771-2050
FAX (305)771-2052

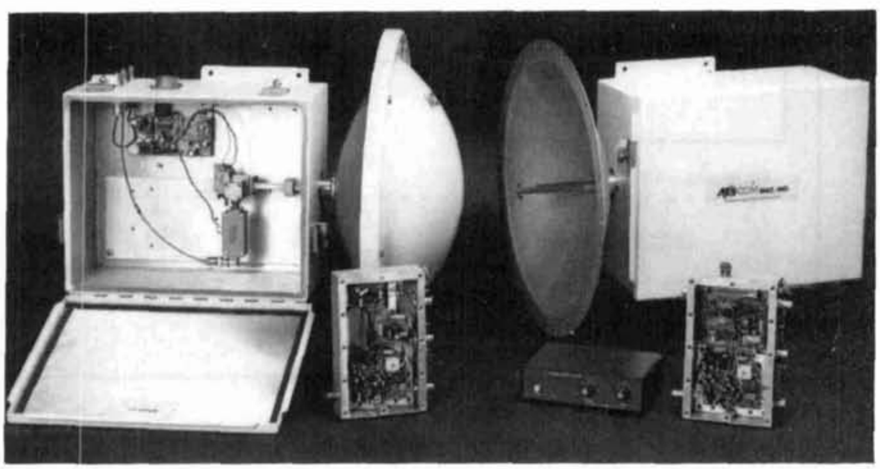
5821 NE 14th Avenue • Ft. Lauderdale, FL 33334
Visa, MC, COD, Cash, M.O. accepted. Personal Check allow 3 weeks. 5% Shipping, Handling, (Maximum $10) U.S. & Canada. 15% outside continental U.S.A.

By John V. Bellantoni, *WB1ALZ*
777#88 San Antonio Road
Palo Alto, California 94303, and
Steven P. Powell, *N2BU*
Cornell Amateur Radio Club
W2CXM, 401 Barton Hall
Cornell University
Ithaca, New York 14853.

A COMMUNICATIONS SYSTEM USING GUNNPLEXER TRANSCEIVERS

Want a fast and easy way to get on the microwave bands? Here's a scheme that uses Gunnplexers™ and inexpensive portable (Walkman™) stereo radios.

A former Cornell Amateur Radio Club president, who now works in the microwave industry, donated a pair of 24-GHz M/A-COM Gunnplexer™ transceivers for our use. These Gunnplexers are similar to the MA-87127 series, which are popular with hams, but operate at 24 instead of 10 GHz. We wanted to put them on the air with a minimum of hassle and circuit building. We tried using FM radios for the IF/AF circuit, as suggested in the MA-87127 data sheet and *The ARRL Handbook*,[1] but immediately discovered two problems. First, the Gunn oscillators drifted with time and temperature, so the FM radios required constant adjustment. Second, the Gunnplexers'100-MHz IF signal overloaded the front end of the FM radios, resulting in poor audio quality. To rectify these problems, we modified a pair of inexpensive portable FM stereo radios (Gemini AS10K*), used a analog AFC circuit to keep the two Gunnplexers "locked" to each other, and installed a digital circuit which searches out and finds the lock when power is first applied, or if the lock is lost. The result was a high-fidelity communications link for which we have found a variety of uses at W2CXM. We have used the system for such applications as DXing, demonstrations, and as a link between our remotely located 2-meter repeater and the club room.



*Available at many K-Mart stores.

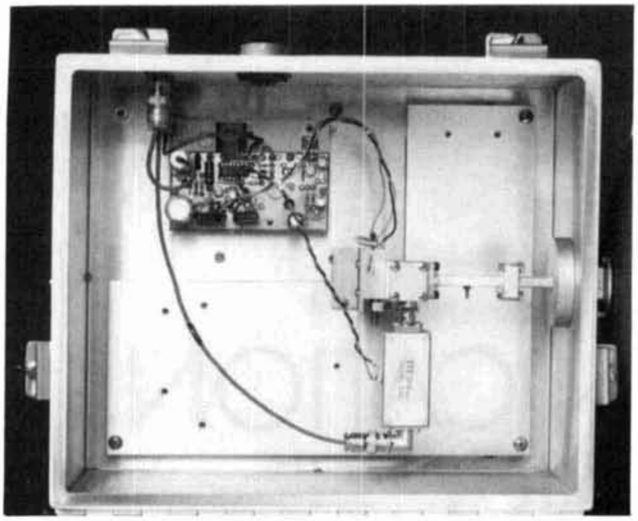


Photo A. A 24-GHz Gunnplexer. From right to left: waveguide tuning screw, circulator, and Gunn oscillator. Just below the circulator is the mixer and IF preamplifier. Above and to the right of the Gunn oscillator is the Gunn diode power supply board.

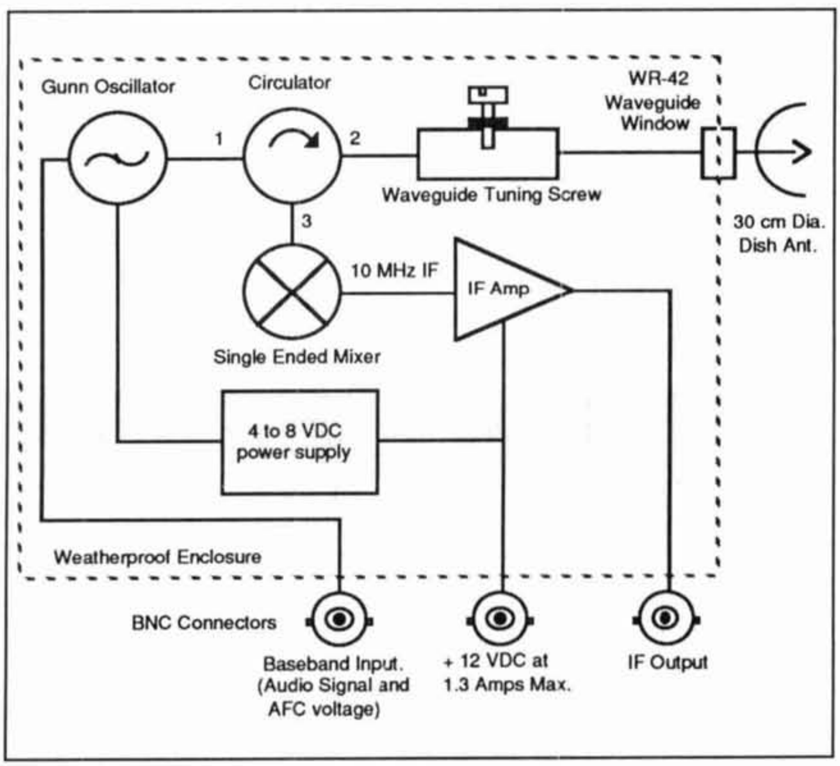


Figure 1. Block diagram of the 24-GHz Gunnplexer.

Gunnplexer Transceiver

Photo A and **Figure 1** show the 24-GHz Gunnplexer and a block diagram. The heart of the unit is the Gunn oscillator. It consists of a Gunn diode mounted in a waveguide cavity. The cavity acts like a very high-Q resonant circuit with a center frequency, in this case, of 24 GHz. Applying a DC bias (4 to 8 volts DC at 1 amp is typical) to the Gunn diode causes the circuit to oscillate at the frequency of the cavity. A mechanical tuning screw on the side of the Gunn cavity lets you shift the center frequency by several hundred MHz.

Gunnplexers are primarily used as the front end for two-way communications. In this mode, two units are used with their carrier frequencies offset by the IF frequency – typically in the 10 to 200-MHz range. The Gunn oscillator acts as a transmitter and a local oscillator for the receiver down-converter, simultaneously. Full duplex is an inherent benefit of this configuration. It allows concurrent reception and transmission – much like a telephone conversation. Most Amateur applications will only require simplex operation. This means that, at any given time, one unit will be used as the transmitter, while the other will act as a receiver down-converter. The transmitted microwave signal generated by the Gunn oscillator travels from port 1 to port 2 of the circulator and past the tuning screw on its way out to the antenna. However, a small amount of the transmitted signal is reflected back by the tuning screw through port 2 to port 3 of the circulator, and into the mixer. This is the LO signal. The incoming signal received by the antenna travels from port 2 to port 3 of the circulator into the mixer, where it's combined with the LO signal to generate the IF signal.

You can apply FM modulation to a tuning varactor mounted inside the Gunn oscillator cavity. The varactor is set close to the Gunn diode in the waveguide cavity, and its input is usually called the baseband input. Any change in the varactor's capacitance due to a modulation voltage, will change the oscillating frequency. The frequency change for the 24-GHz Gunnplexers is on the order of 60 to 90 MHz for a 0 to 12-volt signal at the varactor input (**Figure 2**) – $\simeq$ 7 MHz/volt deviation. To keep the IF signal from the Gunnplexers within the 70-kHz bandwidth of the FM stereo radio IF section, apply no more than $\simeq$ 10 mV modulated signal to the baseband input. Note that increasing the voltage into the varactor increases the operating frequency. This relationship is essential to the operation of AFC circuit detailed in the next section.

W2CXM Gunnplexer communications system

A diagram of the W2CXM Gunnplexer system is shown in **Figure 3**. The Gunnplexers are offset by the 10-MHz IF frequency. The Gunnplexer set to the lower frequency is controlled by an analog AFC; the upper Gunnplexer is controlled by a digital lock circuit.

Figure 4 helps explain the operation of the analog lock circuit. The output from the FM detector in the KA2248A chip has two components. One is the modulated signal which is fed into the audio circuits (pin 1 of the KA2264 chip) via a blocking capacitor. The other is a DC voltage proportional to the IF frequency, as shown in **Figure 4**. If, for example, the two Gunnplexers start to drift apart, the IF frequency and the DC voltage start to increase. This voltage is fed back into the varactor of the lower Gunnplexer by way of the analog lock circuit (**Figure 5**). Increasing the voltage increases the frequency, so the lower Gunnplexer "chases" after the upper one. You'll get the opposite effect if the two Gunnplexers drift toward each other. The IF frequency begins to decrease, as does the DC voltage feed into the lower Gunnplexer. In other words, it starts to "run away" from the upper Gunnplexer. This is an example of negative feedback. What results is that the IF frequency remains essentially constant, despite changes in the operating frequency. The analog circuit is an excellent way to lock the two units, but its range is limited. Every time power is applied or the lock is lost, the free running or "upper" Gunnplexer must be adjusted until the lock is found.

To eliminate the need to tune one of the Gunnplexers each time power is applied or the lock is lost due to misaligned antennas, we designed a digital frequency lock circuit (see **Figure 6**). The circuit scans the voltage applied to the upper Gunnplexer until a signal appears. This signal is indicated by a digital low voltage at the signal LED driver (pin 7 of the KA2248A chip). The circuit then stops scanning so the analog lock can take over. U3 is configured as a simple digital-to-analog (D/A) converter clocked by an NE555. Its output is buffered by a LM358 and applied to the Gunnplexer's baseband input. The voltage range that U3 can scan is set anywhere over the 12-volt range by the two 1-k resistors at the base of Q4 and Q5. Q4 sets the upper voltage, while Q5 sets the lower one. The 555 is turned on and off by the signal LED driver output of the KA2248A IF chip via the U1 circuit. U1 serves to AND together two different time

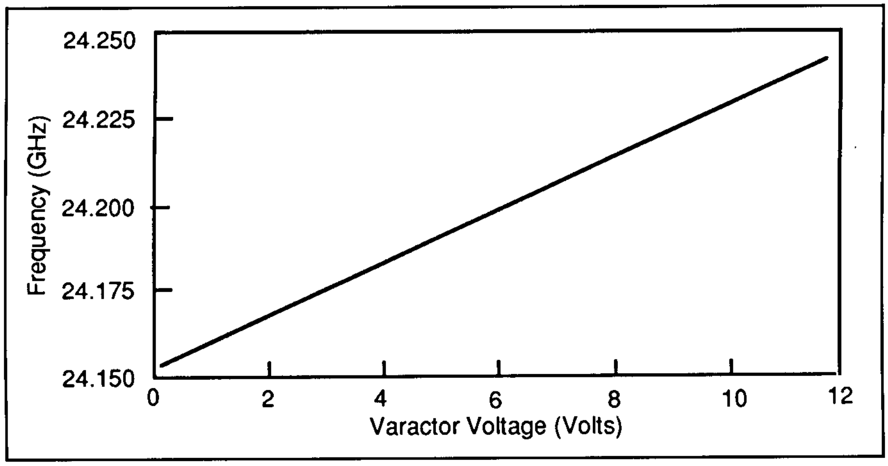


Figure 2. Typical Gunnplexer tuning curve.

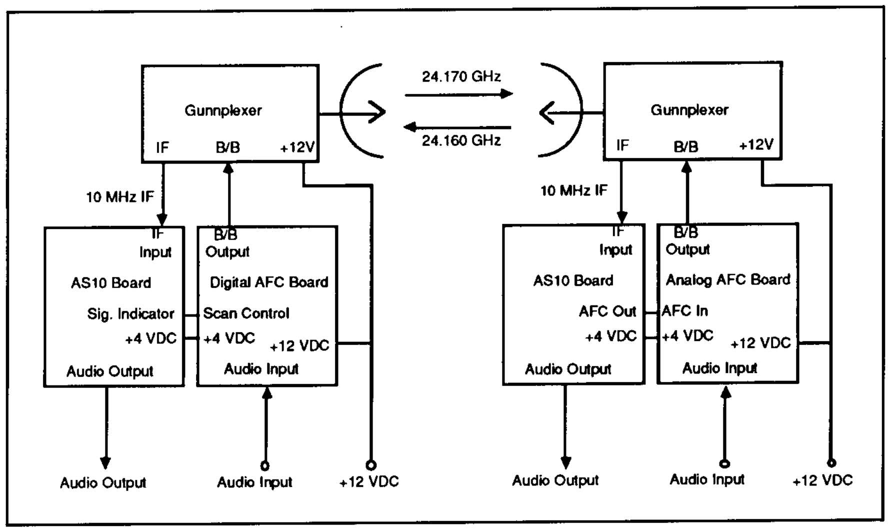


Figure 3. Block diagram of the W2CXM Gunnplexer communications system.

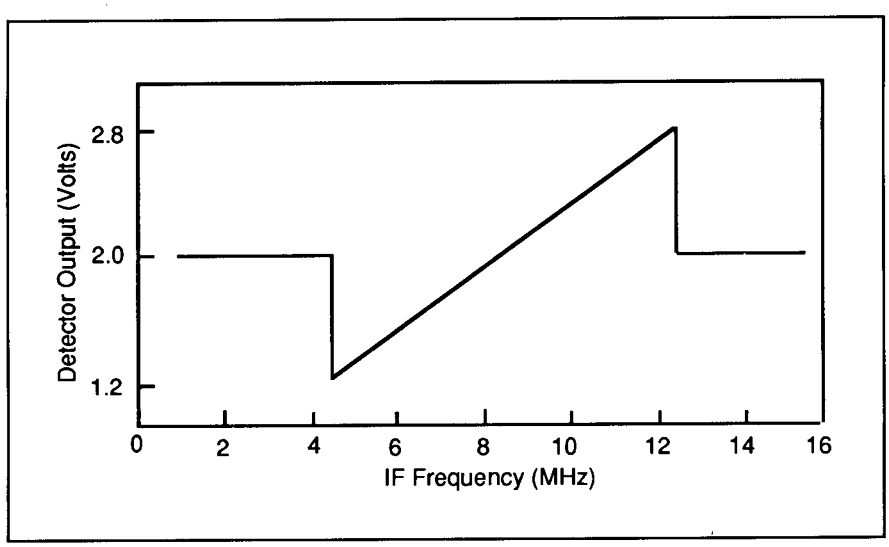


Figure 4. DC output voltage out of the FM detector in the KA2248A chip. Within the range of 4 to 12 MHz the DC voltage increases with the IF frequency. AFC can be performed by feeding the DC voltage back into one of the Gunnplexers.

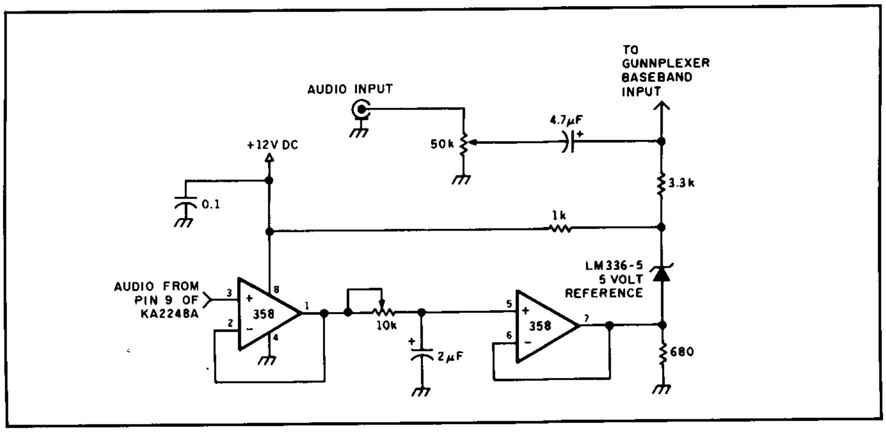


Figure 5. Analog lock circuit schematic. The DC voltage from the FM detector is buffered, filtered, and level shifted up 5 volts before being applied to the baseband input of the Gunnplexer. The 10-k trimpot sets how fast the lock responds to changes in frequency.

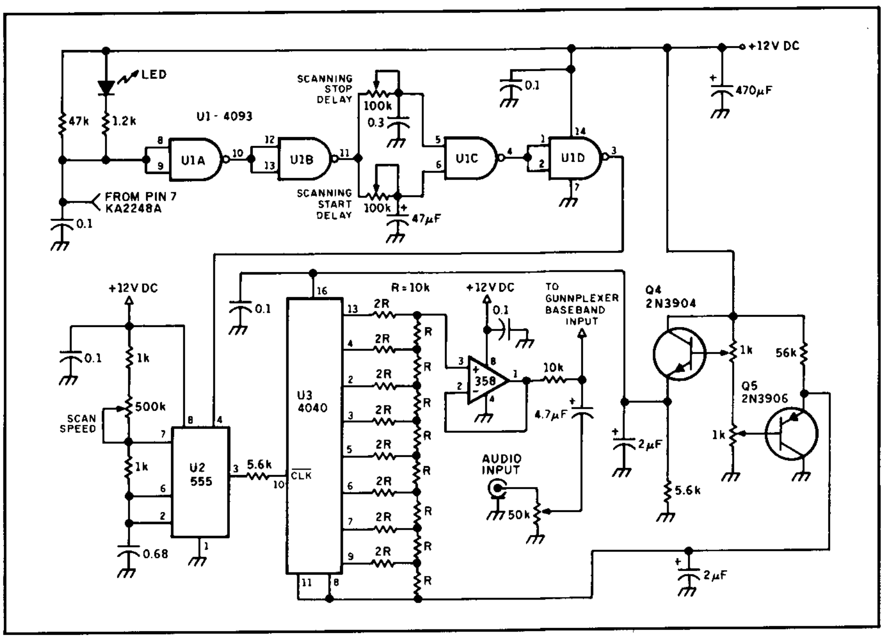


Figure 6. Digital AFC circuit.

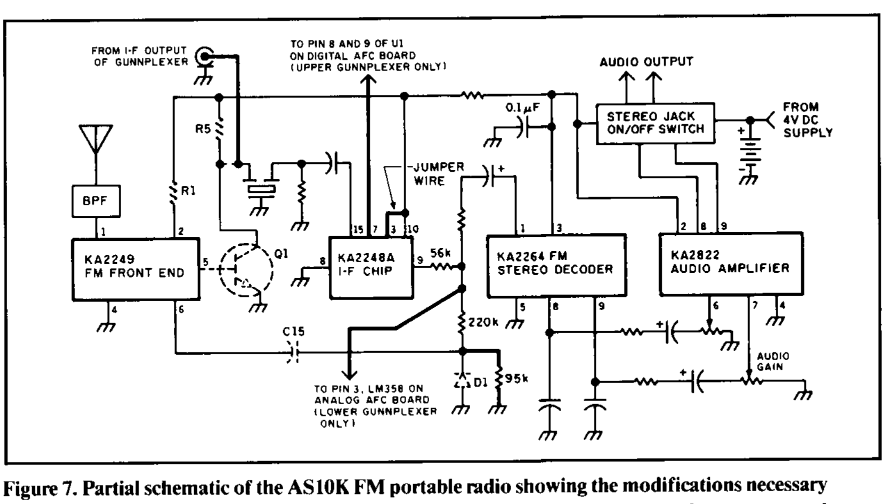


Figure 7. Partial schematic of the AS10K FM portable radio showing the modifications necessary to interface with a Gunnplexer. Added modifications are in bold, components that are removed are shown by dashed lines.

constants set by the 100-k resistors. The slow time constant (47-µF capacitor) prevents the circuit from starting to scan if there's a temporary loss in lock when, say, a bird flies through the beam or wind buffets the dish antenna. The fast time constant (0.3-µF capacitor) sets the amount of time after a signal is found before the scan stops. This lets the circuit scan into the middle of the analog lock range before stopping, as opposed to stopping right at the edge where the signal first appears.

AS10K modifications

Figure 7 gives the modifications you must make to the AS10K. The circuit board is very crowded, so you'll find some type of magnifying lens (like a 10X microscope) and a low-wattage soldering iron (12 to 15 watts) helpful for some of the work.

- Short out the on/off switch located in the headphone jack. Use a multimeter to locate the switch terminals on the PC, then solder a wire across them.
- Disconnect power from the KA2249 FM front end chip by removing R1, a 4.7-ohm resistor. The 4.7 ohm also supplied +V to pin 3 of the KA2248A chip so, to put +V back on pin 3, run a jumper wire from pin 10 to pin 3 of the KA2248A chip.
- Connect coax to the crystal filter. Make room for the coax by removing the components in line between the KA2249 chip and the crystal filter. These are R3, R5, Q1, C5, and C9. Run a length of coax (RG-174/U) from the filter input to a BNC jack mounted on the side of the enclosure.
- Tap off an AFC control voltage. Remove tuning diode D1. Replace with a resistor in the 90 to 100-k range. Run a wire from the top of the 220-k resistor to the analog AFC circuit (pin 3 of the 358 in **Figure 5**). Do this only for the Gunnplexer tuned to the "lower" frequency.
- Tap off a "lock found" signal. In most FM radios, this signal is the one that drives the tuning LED. It goes to a digital low (0 volts) when a signal is present. Connect a wire from pin 7 of the KA2248A IF chip to the digital AFC circuit (pins 8 and 9 of U1 in **Figure 6**). Do this only for the Gunnplexer tuned to the "upper" frequency.
- Set the mono/stereo switch to mono.
- Connect the audio and power. Cut the headphone cable to a convenient length, and run it to an audio jack of your choice mounted on the side of the enclosure. Connect the wires for the battery to the +4 volts DC power supply.

As AS10K radios may not be available in some locations, it's useful to look at the functions inside the KA2248A FM IF/AM chip with an eye to modifying other FM/AM portable radios. **Figure 8** is a block diagram of the chip taken from the Samsung data book. The IF signal input to the chip is FM IN (pin 15). After amplification, the signal is fed into a quadrature detector. Coupling capacitors aren't present at the output of the detector, and this allows both DC and audio components to be sent to AF OUT (pin 9) via a buffer amplifier. The detector output is also used for AGC and the LED DRIVER function. To modify other types of FM radios, locate the detector output before the blocking capacitor and connect it to the AFC circuit as described in the fourth step of the AS10K radio modifications. Be sure to have some type of resistor divide circuit [56 k/(220 k + 95 k) in **Figure 8**] to knock the change in the DC voltage down to a level suitable for the Gunnplexer. Also check the logic supplied to the signal LED with a voltmeter. The LED IND (pin 7) of the KA2248A goes to logic 0 when a signal is found. Some radios may have a logic 1, which means you should remove one of the inverting gates in the digital AFC board (connect pin 4 of U1 to the 555 instead of pin 3).

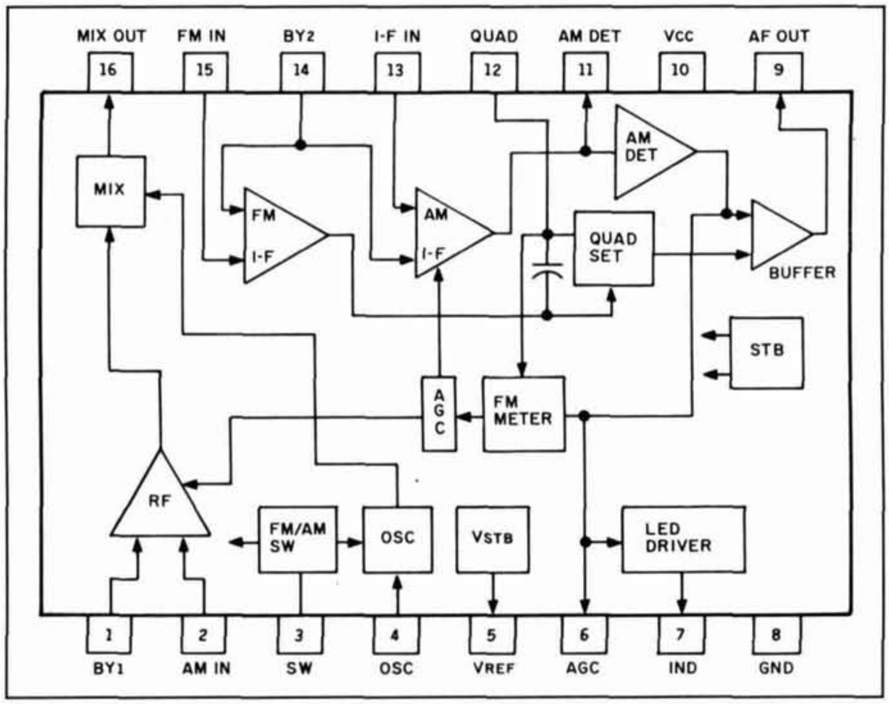


Figure 8. Block diagram of the KA2248A FM/AM IF chip.

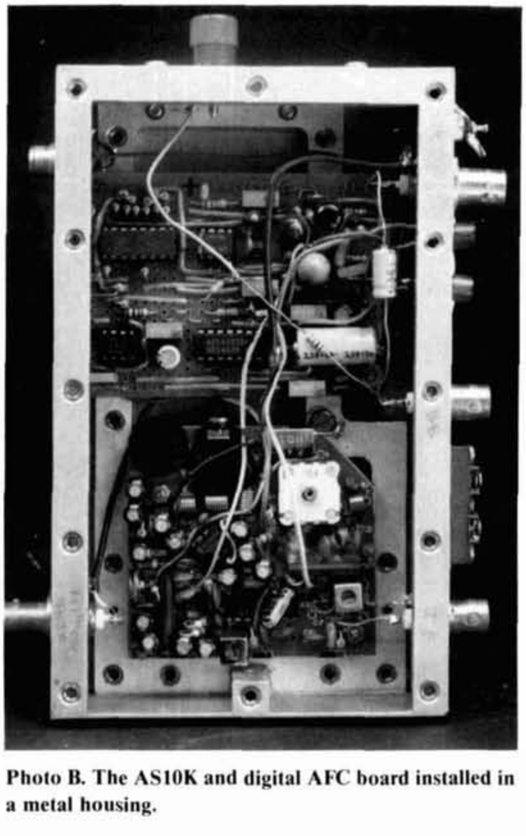


Photo B. The AS10K and digital AFC board installed in a metal housing.

Construction and alignment

There's very little that's critical in the construction of the system. We removed the AS10K boards from their plastic cases, added the modifications, and mounted them – along with the AFC circuit boards – with 0.5-inch standoffs in metal boxes (see **Photo B**). We used BNC connectors and coaxial cable to connect the IF, baseband, and 12 volts to the Gunnplexers. The AFC circuits were built on a couple of pieces of scrap experimenter's pc board (Douglas Electronics Vector Board). Stranded wire from a section of ribbon cable interconnects the AS10K and the AFC boards. We also built a +4 volt power supply for the AS10K on each AFC board. The power supply design (shown in **Figure 9**) was based on what we had in the junkbox. Almost any circuit will work, as long as it's well filtered and can supply at least 50 mA. Dig up a pair of microphones, speakers, and 12-volt (1.5-amp) power supplies from the junkbox and the system is ready to go on the air.

Alignment is a two-step process. You begin by aligning the Gunnplexers. **Appendix A** lists the steps we used. Hopefully, your Gunnplexers will already have been aligned at the factory, because some fancy microwave equipment is required for alignment. The only difficulty you may have with a pair of pre-tuned Gunnplexers will be that the manufacturer usually sets the IF frequency to 30 MHz. You can, however, use the mechanical tuning screw on the side of

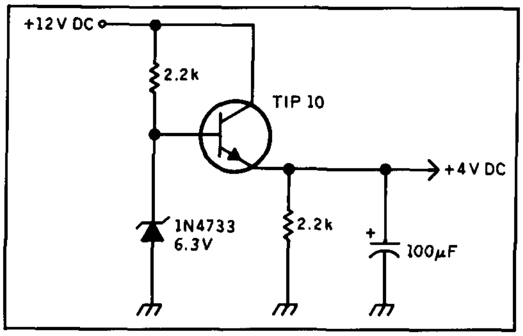


Figure 9. Schematic for the +4 volt supply. Each AFC board has one of these supplies on it to power the AS10K board.

the Gunn oscillator to change the IF to 10 MHz (the tenth step in **Appendix A**). First, power up the Gunnplexers with the 12-volt supplies and place identical DC voltages of, say, 6 volts-DC on each baseband input. Connect a frequency counter at the IF port, aim the Gunnplexers at each other, and observe the IF frequency. Now adjust the mechanical tuning screw very slowly, until the frequency counter reads 10 MHz. If you happen to pick the "upper" Gunnplexer, you'll notice that the screw will have to be turned in slightly (clockwise), while the "lower" unit will require that you back the screw out a bit (counter-clockwise). Mark the Gunnplexers, so you'll know which Gunnplexer to connect to the digital and which to connect to the analog AFC board.

Continue the alignment procedure by adjusting the trimpots on the AFC boards. On our unit, the trimpot on the analog AFC board is set to filter out signals above 10 Hz; however, the system didn't seem to be very sensitive to its setting. The trimpots on the digital AFC board require more care in adjustment. The scanning voltage range is set by the 1-k resistors at the base of Q4 and Q5. The lower voltage – set by Q5 – should be high enough that the system doesn't try and lock up at its image frequency; that is, at the frequency where the "upper" Gunnplexer is actually 10 MHz below the "lower" Gunnplexer, instead of 10 MHz above it. Four volts at the emitter of Q5 should be sufficient. The upper scan voltage is set to ensure that it isn't possible to damage the varactor in the Gunn oscillator by an overvoltage condition. Adjust the 1-k resistor at the base of Q4 so there's no more than 10 volts available there. We set the scan start trimpot to a 1.5-second delay. We found that adjusting the speed of scan, as set by the 500-k resistor on the 555, so it took 1 to 3 seconds to scan through the voltage range worked best. Some tinkering will be necessary between the 500-k scan speed and the 100-k scan stop, as the two settings interact with each other. The best way to set the scan stop is to put a voltmeter at pin 7 of the KA2248A chip, drop the lock (aim one of the dishes away for a moment), and bring it back up to check to see if the voltage is in the middle of its range as shown in **Figure 2**. The 50-k trimpots at the audio inputs on both boards should be set so the audio signal going into the baseband input is on the order of 10 to 15 mV.

Once aligned, operation couldn't be simpler. Aim the two dishes at each other and turn on the 12-volt power supplies. The digital AFC board will find the lock automatically, and the analog AFC will keep it there. Adjust the audio volume to a comfortable range and QSO away.

Conclusion

So there you have it: a way to get a pair of Gunnplexers on the air with high quality audio, but without building a lot of IF/AF stuff. Because we were given the Gunnplexers, the project was inexpensive, too. It cost the club $26 for the two FM portables; the rest of the parts were scrounged from various junkboxes. The link has found a variety of uses here at the club. Probably the most fun we've had, besides putting it together, has been to use the system for demonstrations and lectures on ham radio. We also used the system to connect the phone line at the club room up to our remotely located repeater. The KA2264 stereo decoder chip on the AS10K board proved ideal for this application. With the mono/stereo switch in stereo position, the stereo LED indicator (pin 6) lights up every time a 19-kHz stereo pilot tone is present. We installed a 19-kHz oscillator at the repeater site which is controlled by the repeater's computer. The 19-kHz oscillator turned on every time someone wanted to use the phone patch. At the club room, we tapped off the signal that lights up the stereo LED to a relay driver. The relay connected a Heathkit phone patch to the phone line. The net result was a phone patch via 24 GHz! Whatever application you have in mind for a pair of Gunnplexers, we're sure that the W2CXM system will help you get the job done.

Acknowledgments

Many thanks to Dennis Cleary, WA2PKP, for donating the Gunnplexers and Professor Rick Compton of the Electrical Engineering School at Cornell University for letting us use equipment in the microwave lab. ■

Reference

1. *The 1990 ARRL Handbook*, ARRL, Newington, CT, pages 32-15 to 32-21.

APPENDIX A: 24-GHz Gunnplexer alignment procedure

- Connect a WR-42 matched load to the antenna port.
- Adjust the Gunn oscillator power supply for the proper Gunn diode bias voltage, as specified on the oscillator (5.5volts DC typical). Set the varactor voltage to 6-volts DC (middle of the 2 to 10-volt tuning range).
- Disconnect the mixer/IF preamplifier assembly from the circulator.
- Connect a WR-42 10-dB coupler to the circulator. Connect a frequency counter to the coupled arm and a power meter to the through arm.
- Adjust the waveguide tuning screw for +5 dBm LO power at the mixer port. Set the Gunn oscillator mechanical frequency screw to the desired operating frequency (clockwise lowers frequency, counter-clockwise raises frequency). Readjust the waveguide tuning screw for +5 dBm at the mixer port if necessary. There may be some frequency pushing/pulling due to the tuning screw.
- Power down the Gunnplexer. Disconnect the coupler from the circulator and reconnect the mixer/IF preamp assembly.
- Disconnect the matched load from the antenna port and connect a power meter. Power on the Gunnplexer and check the transmitted output power (+18 dBm typical). If output power is low, check to be sure the Gunn diode bias voltage is correct, and verify that the power at the mixer port is +5 dbm. Power down the Gunnplexer.
- Align the other Gunnplexer as described in the preceding seven steps. Be sure to offset the operating frequency by the IF frequency.
- Connect the two Gunnplexers together through a 40-dB WR-42 attenuator.
- Connect a spectrum analyzer or frequency counter to the IF output BNC connector on one of the Gunnplexers. Make sure 6-volts DC is set on both varactors. Verify that the IF frequency is correct based on the difference of the Gunnplexers frequency set in the fifth step. If the IF frequency is incorrect, carefully readjust the tuning screw on one of the Gunnplexers.
- Using a spectrum analyzer or power meter, verify that the receiver conversion gain (RF to IF) is approximately 12 dB (11-dB conversion loss + 23-dB gain in the IF preamp). Note that the receiver input power is 40 dB down from the output power of the transmitting Gunnplexer. If the gain is low, check that the IF preamp bias voltage is 12-volts DC. Verify that the IF preamp gain is ≈23 dB. If all the above are correct, the mixer diode is probably damaged and needs replacement. Check that the LO power at the mixer port of the circulator is +5 dBm.
- Repeat the preceding two steps for the other Gunnplexer.
- Power down the Gunnplexer and reinstall the antennas.

By R.P. Haviland, *W4MB*
1035 Green Acres Circle North
Daytona Beach, Florida 32119

USING MININEC FOR ANTENNA ANALYSIS

Elements of MININEC theory

The computer program "MININEC" solves a set of complex electromagnetics relations. You don't need a detailed understanding to make practical use of the program, but you'll find a basic knowledge of the principles helpful when setting up the data input and run conditions. (See references for the full theory.) You'll need to do some planning to reduce the time needed for a run, obtain the best accuracy, and give proper interpretation of the results.

The basic concepts

MININEC deals with straight wires, or straight elements thin enough to be treated as wires. Each wire is assumed to be divided into segments (for instance, – – – – – –), with the segments long enough to connect to each other. MININEC treats wire ends as half-segments, so the wire shown in the parentheses would be a four-segment dipole.

Somewhere in the assembly of wire segments which make up the antenna, there must be at least a single segment which is fed by an external source. For example, assume that you have a single driven segment which is the center segment of the lower of a pair of wires, as = = = = =.

The radio frequency current flowing in this driven segment induces an electric field in the surrounding space. Its intensity is a function of the distance and angle from the driven segment. Due to the difference in intensity, the top wire segments see a varying intensity which is at maximum at the center segment and decreases toward the ends. The field difference causes a variation in the electric charge along the wire; that is, a potential difference from one place to another.

As a consequence of this potential difference, current will flow in the segments of the top wire. There will also be a potential difference for the non-driven segments of the bottom wire and the current flowing in each segment. Because current is now flowing in all segments, each will produce a field component and, in turn, a current component in all other segments. You must evaluate each of these currents to obtain an antenna performance solution.

Because the process relates to voltages and currents, it's convenient to express their interrelationship in accordance with the alternating current version of Ohm's law, as:

Current = Voltage/Impedance

Where each of the three quantities has many components, one for each wire segment for current and voltage, and the square of this number for impedance.

MININEC solves these relations from the antenna geometry plus the source voltage. In the process, it provides data for other important items. The major ones are:

- driving impedance
- input power
- mutual impedance

Because you now know the current in each segment, along with the geometry, you can have the program calculate the radiation field using array theory. This will give the radiation pattern.

Refinements

A number of refinements are available. Two or more wires can form a beam. Two or more wires can also be connected at a point to form antennas like the "T." Impedances can be introduced at any segment, as in loaded antennas, or for impedance matching.

One important factor is the presence or

absence of the Earth. MININEC assumes either that the Earth is remote, or that the part directly under the antenna is perfect. This corresponds with adding an image antenna identical to the source, or underground by the antenna height. However, the program calculates the radiation pattern for a remote Earth or a near Earth, using the Earth constants provided. An alternative to these constants is a perfect Earth.

Practical computer limits

MININEC is a large program. Storing the arrays used by MININEC takes memory. As a result, each computer has a limit to the antenna complexity it can handle. Also, because of the enormous number of computations needed, MININEC will be slow when dealing with antennas having a large number of wire segments.

Don't let the length of the run time discourage you from using MININEC on large problems. Just accept the fact that the time the program takes is necessary, and do something else while the computer is running. (Just be certain that there's a surge protector on the power line!)

Data needs of MININEC

MININEC has a good set of prompts which make data input relatively easy. I'd like to continue by reviewing the input requirements, not in the order they occur, but in their order of importance to the final results. At the end of this article you'll find an actual printout of a final run. You may find it helpful to refer to this for an overall view of the prompts. Each specific data request is covered in detail in the following paragraphs.

Frequency

Enter the desired frequency in MHz. If you enter a zero the frequency will be set to 299.8 MHz.

Selecting the number of wires

MININEC deals only with straight wires. You must treat any design which has a bend, like a V, as an assembly of two or more wires. For example, a dipole is a single wire, and a square loop quad element is made of four wires.

In a few cases, you may find it necessary to have two or more divisions of a single straight wire. For example, if a dipole is to be fed as a Windom, there must be wire ends at the Windom feed point to allow connection of the single wire feeder.

The first version of MININEC specifically required that you input the fact that wires were connected. Connection input errors were common, so later versions were changed to assume wire connection if the end coordinates are identical. This must be in three dimensions. For example, if the start end of wire one has the X,Y,Z, coordinates 1,1,1, and the finish end of wire two is also at 1,1,1, MININEC assumes these wires to be connected. It makes no difference whether the start or finish end is involved. However, in data output, the direction of current in the wire is affected. Current always flows to the lowest potential point, so the sign of current is reversed.

There's no limit on the separation distance of two wires, or on the angle between them. But versions of MININEC earlier than 3.12 lose accuracy if two connected wires make a small angle. The error is negligible for angles greater than about 10 degrees, so this limitation isn't severe.

Segmenting the wires

MININEC is based on the concept of constant current in a short segment of a wire. This can be a poor approximation when the current varies along the wire and there are only a few segments. You'll need a large number of segments to get a good approximation of current.

On the other hand, the number of calculations which must be made increases at a level which is approximately the square of the number of segments, so the time needed to obtain a solution increases rapidly. Also, information must be stored for each segment. This too varies as the square. Consequently, small computer users must resign themselves to:

- staying with small problems if high accuracy is needed, or
- accepting long run times, or
- choosing the number of segments for reasonable accuracy.

The accuracy versus number of segments does vary with the antenna design. This makes it impossible to set down hard and fast rules for segmentation. I've found the following a useful guide:

- Use four to five segments per half-wave of wire for an indication of performance. Expect an appreciable error in reactance values.
- Use eight segments per half-wave for reasonable practical accuracy.
- Use sixteen or more segments if you need good accuracy.

When in doubt, try runs with differing segmentation and compare the results.

Remember that super accuracy is unnecessary for practical antennas. Construction tolerances, element sag, presence of tower and boom, and nearby objects all affect performance. Reasonable accuracy, say ± 20 percent, is nearly always enough.

The specific input requests of MININEC, and the needed keyboard responses are:

NO. OF WIRES
-Enter number; 0=* (load from disk)
NO. OF SEGMENTS
-Enter number
END ONE COORDINATES
-Enter X, Y, Z in meters
END TWO COORDINATES
-Enter X, Y, Z in meters
RADIUS
-Enter wire radius, meters

(Note: The star indicates that a jump to another routine occurs if the input is as indicated.)

At this point, the program makes wire connections and prints a table of connections.

CHANGE WIRE NO.
-Input Y for a change or N=next wire

After you've input the last wire, the program outputs a second geometry table. Check to make sure the location of the segment(s) to be fed or loaded and the connections are correct.

CHANGE GEOMETRY
-Enter N= no change, or Y= repeat at NO. OF WIRES

The feed point

MININEC requires that at least one segment be designated as a feed point, though you may choose to have many feed points. The feed voltage and phase angle may be set to any value you desire. Remember that control of the feed conditions is a major factor in antenna pattern control, so be certain that the feed voltage and phase are the values you need.

In simple antennas with one feed, it's convenient to set the feed voltage to unity and the angle to zero. This allows currents in segments to be read directly as mutual admittances from the drive point ($Y = I/E$, $Z = 1/Y$). Use 1000 volts to obtain values in millimhos. Multiple runs with different feed points will give all inter-element impedances. Note that this is usually defined only for the impedance to a current maximum, which is often the center of an element.

There's an interrelationship between the number of segments in a wire and the allowable feed-point location. MININEC prints the coordinates of each wire end point, and follows this information with the coordinates of each segment center. Because feeds are assumed to be at the center of a segment, this gives you the coordinates of allowable feed points. Corner-fed antennas are a problem, and you may find it necessary to introduce a very short wire of one segment.

The MININEC input requests and keyboard responses are:

NO. OF SOURCES
-Enter number of drive feed points; 1 is assumed, if a lesser number is entered
PULSE, VOLTAGE, PHASE
-Enter number of the segment to be driven, drive in volts, phase in degrees for each driven segment

Environment: introducing the Earth

MININEC lets you introduce Earth conditions so the program can calculate vertical patterns of elevated antennas. Several options are allowed, with one restriction.

The simplest option is to assume an ideal Earth. If an option other than ideal Earth is chosen, you may select up to five Earth conditions. With the exception of a single condition, this applies everywhere (except as it applies to drive impedance). You can set both the Earth dielectric constant and resistance. If you want more information, check the book *Reference Data for Radio Engineers*, or ask the engineer of your local broadcast station for the values used on their FCC application. You may set the conditions to apply to concentric circles about the antenna, or to parallel strips. Also, or alternatively, you can set the elevation of each such area. This lets you simulate an antenna on a circular hill, a ridge, a hillside, or in a valley. Because you're limited to five sets of parameters, there are some accuracy limitations.

Antennas fed against ground must have at least one end with a height of zero. It's also possible to introduce a ground screen directly under the antenna. Radial wires are assumed.

The MININEC input requests and keyboard responses are:

ENVIRONMENT
-Enter 1 for free space conditions *, −1 for ground plane
NUMBER OF MEDIA
-Enter 0 for ideal ground *, 1 to 5 for number of media
TYPE OF BOUNDARY
-Enter 1=Linear, 2=Circular
MEDIA
-Enter, in order, the dielectric constant,

conductivity in mhos/meter for
each media

NUMBER OF RADIAL WIRES
IN GROUND SCREEN
-Enter 0=no screen *, or number

RADIUS OF RADIAL WIRES
-Enter radius, meters

X OR R COORDINATE OF
NEXT MEDIA
-Enter distance, meters

HEIGHT OF MEDIA
-Enter height relative to 0,0,0 point, meters

Loaded antennas

MININEC lets you introduce resistance or reactance at the center of each segment. If introduced at a feed point, the addition represents a shunt, which is often used for impedance matching. If introduced away from the feed, the addition represents a loading element – usually a reactance.

It's also possible to enter the added resistance-reactance as an S-parameter. Make your entries in this order: the parameter, followed by the numerator and denominator of each term, starting with the highest order. See texts on Fourier transforms for details on such usage. For single frequency analysis, it's easier to calculate the reactance separately, and use this value as input.

The MININEC input requests and keyboard responses are:

NUMBER OF LOADS
-Enter number of segments to be loaded, 0=none *

S-PARAMETER LOAD
-Enter Y if needed, or N *

PULSE NUMBER, ORDER OF S
-Enter number of the segment to be
-loaded, highest S exponent

NUMERATOR, DENOMINATOR
COEFFICIENTS OF S
-Enter coefficients, ohms, henries, and farads, as appropriate, for each loaded segment

PULSE, RESISTANCE, REACTANCE
-Enter number of segment to be loaded, load resistance, load reactance, ohms, each loaded segment

Obtaining desired output

MININEC has a number of output possibilities. The major items to consider are:

- **Output location**: near to or far from the antenna. The far field is more commonly used.
- **Orientation**: making certain that output data is at the correct angles with respect to the antenna.
- **Incrementation**: making certain that the output increments give the necessary details.

I suggest you use a standard orientation for all problems. A useful set of rules is:

- Align the dominant wires to the X axis. For example, Yagi elements are on or parallel to the –X to +X line.
- For antennas occupying a third dimension, such as quads or stacked Yagis, try to have this dimension along the Z axis. Thus, a quad loop lies in the XZ plane or is parallel to it.
- For symmetrical antennas, place the point of symmetry at 0,0,0 for free space antennas, or at 0,0 height for antennas above the earth.
- For unsymmetrical single feed antennas, place the feed point at 0,0,0 for free space, or at 0,0 height for above Earth conditions. If you're using multiple feeds, try to keep them symmetrical about 0,0,0.
- Place a little sketch on the printout sheet if you don't follow these rules.
- Note unusual conditions on the output printout.

If you follow these suggestions, two planes will give a good picture of the pattern. The first is the XY plane – the horizontal pattern. The second is the YZ plane – the vertical pattern. Sometimes you'll need the XZ pattern.

The patterns can show the vertical, horizontal, or total polarization component. Where Earth reflection is involved, there's no horizontally polarized far-field radiation along the horizon, so the true XY plane pattern may be useless. The procedure in such cases is to set an elevation angle (90-zenith angle) and make an azimuth sweep. Usually, the elevation you select will be that of the nose of the main lobe. However, in studying the effectiveness of a design, angles can be related to propagation modes. Elevations of 5, 10 and 30 degrees are often used to indicate maximum F-layer, normal F-layer, and E-layer effectiveness.

The angle increment should be related to lobe widths. For two dipoles reasonably closely spaced, 20 to 30 degrees will be satisfactory. But a multi-element antenna, wide-spaced elements, or an elevated 2-meter antenna may have lobes of 10 degrees or so width, so you may need increments of 1 to 5 degrees. You can often reduce the amount of pattern data needed by considering antenna symmetry.

With a new antenna type it's probably worthwhile to generate the complete pattern and file it on disk. In other words, azimuth should cover 0 to 360 degrees and elevation

0 to 180, or 0 to 90 if ground is present. Note that there will be data output for angles below the horizon if the Earth is present, but this data is useless.

There are three output possibilities, two for permanent data retention:

OUTPUT TO CONSOLE = C *,
PRINTER = P *, Disk = D
-Enter selected letter
FILENAME
-Enter the filename and filepath to disk
P = COMPUTE FAR-FIELD PATTERNS
VARIABLES IN DBI OR
VOLTS/METER
-Enter D * for dBi or V for field intensity, in μvolts/meter
ZENITH ANGLE
-Enter the initial angle in degrees from zenith, increment per step, number of calculation steps
AZITMUTH ANGLE
-Enter initial angle in degrees from zero, increment per step, number of calculation steps

(Note: Zero degrees azimuth is along the +X axis, zero degrees zenith is along the +Z axis. Minus angles are allowed. For complete coverage of the free space pattern, 0 to +180 zenith and 0 to 360 azimuth is required. Choose an increment to show smallest lobe width.)

FILE FAR FIELD DATA
-Enter y for disk copy or N *
FILENAME
-Enter filepath + filename

A second output alternative is:

N = COMPUTE NEAR FIELDS
ELECTRIC OR MAGNETIC
-Enter E = electric, H = magnetic
FIELD LOCATION(S)
-Enter initial location from 0,0,0, increment distance, number of steps
CHANGE POWER LEVEL
-Enter Y=change or N *
NEW POWER LEVEL
-Enter new level, watts, O = old
SAVE TO A FILE Y/N
-Enter Y for a disk file or N = No *
FILENAME
-Enter the filepath + filename

Currents

The program can display each wire segment separately or as a part of pattern computation. The setup input for this is:

C = COMPUTE/DISPLAY CURRENTS

(Note: Current computation is required for any form of solution. If C is not entered, the currents will be computed but not displayed. A table of feed point(s) conditions and power will be output in all cases. If C has been entered, this is followed by a tabulation of current in each segment.

The fastest way to learn MININEC is to practice using antennas of known performance. ■

Bibliography

1. Alfredo J. Julian, James C. Logan, and John W. Rockway, "MININEC: A Mini-Numerical Electromagnetics Code," *Naval Ocean System Center Technical Document 516*, September 6, 1982.
2. James C. Logan and John W. Rockway, "The New MININEC (Version 3): A Mini-Numerical Electromagnetic Code," *Naval Ocean System Center Technical Document 938*, September 1986.

Both sources are available from The National Technical Information Service, 5285 Port Royal Road, Springfield, Virginia 22161.

Obtaining MININEC

Two versions of MININEC are available from CQ's Ham Radio Bookstore.

The program disk, "Practical Antenna Design and Analysis," by R.P. Haviland, W4MB, includes a copy of the original version of MININEC. This is the only one suitable for small computers, including the C-64 and the APPLE. Amiga and IBM-PC disks are also available for $39.95.

W4MB's most recent set of programs, "MININEC For Amateurs," includes four versions of MININEC 3.09. These versions provide for standard operations and swept frequency, variable load, and optimization. Files of designs and programs to design dipoles, verticals, Yagis and quads are included. Disks are available for the Amiga and the PC for $39.95.

Other versions are described in ads in various publications. A Fortran version is available from RF Design Software Service, Denver, Colorado. A "commercial grade" version is available from Artech House, Norwood, Massachusetts.

```
          ****************************************
            MINI-NUMERICAL ELECTROMAGNETICS CODE
                          MININEC
             04-16-1990                 10:41:53
          ****************************************
FREQUENCY (MHZ): 14
    WAVE LENGTH =  21.41429  METERS

ENVIRONMENT (+1 FOR FREE SPACE, -1 FOR GROUND PLANE):  1

NO. OF WIRES: 1

WIRE NO. 1
            COORDINATES                            END       NO. OF
   X                  Y             Z     RADIUS CONNECTION  SEGMENTS
-5                  0            0                 0
 5                  0            0         .001    0           8

                  **** ANTENNA GEOMETRY ****

WIRE NO.  1  COORDINATES                         CONNECTION PULSE
X                  Y            Z         RADIUS    END1 END2  NO.
-3.75              0            0          .001      0    1     1
-2.5               0            0          .001      1    1     2
-1.25              0            0          .001      1    1     3
 0                 0            0          .001      1    1     4
 1.25              0            0          .001      1    1     5
 2.5               0            0          .001      1    1     6
 3.75              0            0          .001      1    0     7

NO. OF SOURCES :  1
PULSE NO., VOLTAGE MAGNITUDE, PHASE (DEGREES):  4 , 1 , 0
NUMBER OF LOADS 0
ZENITH ANGLE : INITIAL,INCREMENT,NUMBER:  90 , 0 , 1
AZIMUTH ANGLE: INITIAL,INCREMENT,NUMBER:  0 , 10 , 10
```

Printout 1. The MININEC header section, which includes the wire segmentation coordinates. The example is for a dipole.

```
FILL MATRIX  : 0:20
FACTOR MATRIX: 0:04
********************      SOURCE DATA      ********************
        FREQUENCY, MHZ.=  14
          RESISTANCE LOAD, OHMS =  0
          REACTANCE LOAD, OHMS =  0
PULSE  4        VOLTAGE = ( 1 , 0 J)
                CURRENT = ( 8.794121E-03 , 8.016245E-03 J)
                IMPEDANCE = ( 62.10681 ,-56.6132 J)
                POWER =  4.39706E-03  WATTS

********************      CURRENT DATA      ********************
WIRE NO.  1 :
PULSE     REAL            IMAGINARY       MAGNITUDE       PHASE
 NO.      (AMPS)          (AMPS)          (AMPS)          (DEGREES)
E          0               0               0               0
 1         3.563988E-03    2.907439E-03    4.59948E-03     39.20696
 2         .0063537        5.326857E-03    8.291255E-03    39.97602
 3         8.165337E-03    7.083316E-03    1.080954E-02    40.94118
 4         8.794121E-03    8.016245E-03    1.189944E-02    42.35061
 5         8.165342E-03    7.083319E-03    1.080954E-02    40.94118
 6         6.353706E-03    5.326861E-03    8.291262E-03    39.97601
 7         3.563992E-03    2.907441E-03    4.599484E-03    39.20695
E          0               0               0               0
```

Printout 2. Source and current output of MININEC. Source is always printed, but current only on demand. Note that this dipole is below resonance at the selected frequency.

```
********************      FAR FIELD      ********************
********************      PATTERN DATA      ********************
ZENITH    AZIMUTH    VERTICAL        HORIZONTAL      TOTAL
 ANGLE     ANGLE     PATTERN (DB)    PATTERN (DB)    PATTERN (DB)
 90        0         -138.1358       -999            -138.1358
 90        10        -138.2148       -14.81311       -14.81311
 90        20        -138.4676       -8.77082        -8.77082
 90        30        -138.9415       -5.237245       -5.237245
 90        40        -139.7213       -2.76949        -2.76949
 90        50        -140.9441       -.9447931       -.9447931
 90        60        -142.8462        .4006619        .4006619
 90        70        -145.9184        1.335945        1.335945
 90        80        -151.6594        1.889931        1.889931
 90        90        -274.8774        2.073665        2.073665
```

Printout 3. Pattern data output. The right columns are for the vertically polarized, horizontally polarized and total intensity.

By B. Sykes, *G2HCG*
Reprinted from *Practical Wireless*
November 1989

THE ENHANCEMENT OF HF SIGNALS BY POLARIZATION CONTROL

As any Amateur knows, to get the best out of your VHF/UHF equipment, you need to be able to control the polarization. But what about HF?

At VHF and UHF the ability to control the polarization response of an antenna from the shack is very useful, indeed. In fact it is almost essential. The polarization of signals received from satellites is rarely known, and depends not only on the attitude and spin of the satellite, but on the fact that signals have to traverse one or more of the ionized layers above the Earth's surface.

Polarization at HF

Little seems to be known about the changes in polarization which occur at HF when signals are reflected from the ionosphere as they are in normal over-the-horizon long distance propagation. Polarization control at HF is unheard of; the antenna and control requirements have always seemed to be quite out of the question. The advent of the Polarphaser* and crossed Yagi system, which gives a simple means of polarization control, has so far only been used at VHF and UHF. Consideration of practical antenna size seems to rule the system out at HF, but a year or so ago, the first experiments were conducted on 21 MHz with a Polarphaser made up for that band – together with crossed dipoles mounted vertically in an X formation. Results were encouraging, but inconclusive, due to the lack of directivity of the antenna system. Control of polarization is only possible when the signal is in the beam of the antenna, and it proved a very difficult practice to ensure that the dipoles were broadside to the received signal, thus enabling control to be achieved.

Openings on 28 MHz

When it became apparent that 28 MHz was really opening regularly, a 28-MHz Polarphaser was made up – together with a suitable antenna system. The antenna consisted of two 3-element Yagis mounted in cross formation on a common boom, with the elements at 45 degrees to the horizontal. The Yagis were designed for optimum back-to-front ratio, using 0.2-wavelength spaced reflectors and 0.1-wavelength spaced directors. Particular emphasis was placed on ensuring a good match to the two 50-ohm feeders. The system was mounted on a rotator at a height of 7.5 meters.

Take off was completely clear in all directions but the South, where the presence of a house impeded the path. Local tests were conducted with G3BFC some 500 meters away to prove the effectiveness of the system. Thirty dB of polarization discrimina-

*The name Polarphaser refers to commercially available antenna phasing control units offered by South Midland Communications Ltd. of Southampton, England for use at VHF. The author has taken the same basic design and adapted it to the HF bands.
The unit accepts two input signals and provides a combined output signal for a specified phase shift between input signals. Using a front panel control, the phase shift can be varied from zero to 360 degrees. Ed.

tion was found to be achievable. Due to the lack of a clear path to the South, most tests were conducted to the East using the Cyprus and Perth beacons – together with numerous contacts with DX stations. The practicality of the project is always in mind with new antenna systems, in particular the vital "Gain/Aluminium" ratio. Six elements are being used and a comparison must be made with the same number of elements used as a straight Yagi, which would add 3 dB to the gain of a single 3-element. A crossed Yagi mounted in X formation with feeders to each half, together with a polarphaser, will give the following polarizations: **horizontal**, through elliptical to **clockwise circular**, through elliptical **vertical**, through elliptical to **anticlockwise circular**, and through elliptical again back to **horizontal**. Slant polarization at 45 degrees is not available, unless provision is made to switch to individual halves of the crossed Yagi.

Serious testing

Before the commencement of serious testing, it was necessary to prove that the sensitivity of the receiver did not vary with rotation of the polarization control. The effect of altering the phase of the feeders is precisely the same as altering their length, which in many installations will be found to alter the sensitivity of the receiver. There are two reasons for this. First – although the feed system is supposed to be 50 ohms – cable, SWR meters, plugs, and sockets are only "nominal" 50 ohms, and may vary widely. The other, more important, reason is that the input impedance of the transceiver on receive is not 50 ohms. The input impedance of the transceiver in use (a Yaesu FT-757 GX2) was measured and, without the attenuator switched in, was found to be 60 ohms. This was good enough to ensure negligible change of S-meter reading when the acid test was applied – namely adding a quarter wave to the feeder length.

Interesting results

Results proved very interesting and very worthwhile, particularly on reception – fully satisfying the previously mentioned Gain/Aluminium ratio. Most QSB proved to be caused by polarization changes, and signals could be peaked up by choosing the optimum polarization – sometimes by as much as 20 dB (some 3 or 4 S-points). Only occasionally did rotation of the polarization control make little difference, and this probably reflected the lack of a 45 degree position in the system.

The most important factor is the rate of change of polarization, and whether it would be possible to compensate for it manually. The number of hops would appear to govern the change, with the expectation that multi-hop propagation would result in random high-speed change. This did not prove to be the case, although tests were conducted at varying ranges to prove or disprove the point.

Single hop

The Cyprus beacon on 28.200 MHz was particularly useful as an example of single-hop propagation. This beacon was the first indication that the band was opening. During the first half hour or so, polarization was changing quite slowly – even remaining constant for up to a minute, with distinctly predominant polarizations differing from day to day. Although it was quite possible to constantly rotate the control to maintain optimum signal strength, tests were conducted to try and ascertain which was the best average polarization to use. One of the two circulars, with a slight advantage to clockwise, was usually predominant – but often vertical was the best, with horizontal the least effective. Although the Cyprus beacon is itself vertically polarized, there did not seem to be any particular difference between the beacon and other Middle East stations using horizontal polarization.

Two hop

Two-hop propagation to India did not differ greatly from single hop, with very similar characteristics and rate of change. The Perth beacon, and many contacts with Australian stations, provided examples of multi-hop propagation and, although there were a number of occasions when the signal could not be peaked manually, quite often the signal behaved as it did with single hop propagation. It became apparent that the ionosphere was having a "combing" effect and imposing the prevailing polarization on the last reflection, regardless of that of the arriving signal. Again, circular gave the best average signals. Leaving the control set at horizontal, as with a normal installation, was a distinct advantage. It always proved possible to bring a weak signal out of the noise by changing to another polarization.

The apparent combing effect of the ionosphere is particularly interesting, and was the subject of further tests. Two receivers were used connected to the same antenna system. One was tuned to the Cyprus beacon known to be vertically polarized; the other was tuned to a Middle East station using horizontal polarization. The correlation of polarization change between the two stations was 100 percent. A further test between the Cyprus beacon (one hop) and the Perth beacon (multi hop) showed a correlation of polarization at least 80 percent of the time.

No doubt

There is no doubt, therefore, that the ionosphere controls the polarization of reflected signals, and that the final received polarization is that of the prevailing ionospheric reflection. The polarization of the signal arriving at the ionosphere is totally irrelevant. The ionosphere can perhaps be visualized as a hanging curtain of wires, swinging in direction and reradiating the signals at the polarization of the direction of the wires at the time of reflection.

Interference reduction was found to be very useful. The 28-MHz band seems to be very prone to man-made hash from motors and computers. Always, this interference is strongly polarized and, being local and unvarying, can be very effectively nulled out. Even the racket from a computer in the shack can be reduced; the amount of reduction depends on the screening integrity of the feeder system in the shack.

Baluns

The use of baluns on the antennas is absolutely vital if locally induced noise levels are to be reduced. Although at first it was expected that QRM would be of differing polarization and could therefore be reduced, the combing effect precluded this – although local stations which had not been reflected from the ionosphere would be of constant polarization and could therefore be nulled out. It is interesting to note that with a good receiver, when the band is open, the general noise level from the ionosphere can be heard varying with polarization change.

The system is capable of accepting transmitter power and circular is normally used. It is very tempting in a "pile up" situation to choose the polarization giving the highest received signal strength, just as the station goes over, thus perhaps the first vital words come in at enhanced signal strength. The problem is whether the received polarization shift is the same as the transmitted shift. The "curtain of wires" analogy seems to indicate that it will be the same, but investigating this aspect requires polarization control at the other station, and this does not exist at the present time.

The use of a computer to automatically choose the optimum polarization gives considerable food for thought. The possibility of a constant 2 or 3 S-point improvement in received signals is a very strong incentive. The old adage – "if you can't hear them, you can't work them, regardless of transmitting power" is very true. The necessary software should not prove too difficult, but the hardware problems would be considerable. Interference is the first – though computers are bad enough on 28 MHz without deliberately connecting one to the antenna system. A very fast a.g.c. line to control the computer may be necessary, bearing in mind that with SSB the computer may have to sweep through all polarizations and choose the optimum – all within the envelope of a syllable of speech. Using average signal strength would be much simpler, but not so effective. The ability to choose optimum signal-to-noise ratio instead of maximum interference levels is high.

Computer control

Computer-controlled active antenna systems have been used at VHF for optimum reception of TV signals for rebroadcast purposes. Commercial techniques at commercial prices undoubtedly exist. The concept of enhancing reception of HF signals by up to three S-points at reasonable prices is now evident and must be pursued.

Displaying the polarization visually

During extended periods of listening to changes in signal strength resulting from polarization change of the reflected signal from the ionosphere, continual turning of the Polarphaser control resulted in somewhat tired fingers. Much thought was given to achieving a method of displaying the polarization of the incoming signal. This would not only allow considerably more research into the varying properties of the ionosphere, but would enable use to be made of those properties in the form of much enhanced signal strengths.

In order to compare signals from different sources in the same part of the world, two receivers were already in use with inputs in parallel from the Polarphaser connected to the two Y inputs of a double-beam oscilloscope. This provided a means of seeing the variations in amplitude of signal resulting from changes in polarization, but the actual polarization could only be read from the dial of the Polarphaser.

Variations in polarization input to a crossed Yagi result in variations of phase and amplitude between the two feeders. It was felt that it should be possible to display these variations on an oscilloscope. One receiver was connected to the 45-degree half of the crossed Yagi, the other to the 135-degree Yagi. The outputs of the receivers were connected to the X and Y inputs of the oscilloscope. Ideally the RF signal before demodulation would be used, but the normal SSB receiver beats the RF signals with a BFO, and if this oscillator is stable, the resulting audio signal will contain the same phase shifts as the RF signal and be much easier to display on an oscilloscope. First tests were unsuccessful, as it was totally impossible to keep the two BFOs in the two receivers in phase. With one oscillator only, replacing the two BFOs in the two receivers was obviously necessary. Not wishing, at this stage, to delve too far into the insides of the modern black box, the answer was to tune the receivers to AM and insert one 28-MHz signal from a stable oscillator into the input of each receiver – thus producing an in-phase BFO on each.

Initially a signal generator was used, and the system worked first time. The first oscillators of each receiver were still in use, but the PLL circuits held frequency and phase exactly constant at all times.

The optimum way to mount a crossed Yagi is with the elements at 45 and 135 degrees to the horizontal. This balances the antenna in relation to the support mast and feeders, and avoids mismatch problems caused by proximity to the mast of the vertical elements. Although none of the elements are vertical or horizontal, polarization in these directions can be generated electrically by altering the phase length of the feeders.

At first the Yagis were mounted in this way, but oscilloscope X and Y plates are mounted at 0 and 90 degrees. This has the effect of rotating the display by 45 degrees. Unfortunately, this cannot be corrected by altering the feeder lengths and, therefore, the phase to each receiver. There are only two ways of correcting the display. You must rotate the tube of the oscilloscope (which is preferable but not always possble), or rotate the antenna by 45 degrees making the elements vertical and horizontal. This was done, and the resulting imbalance between the two halves of the crossed Yagi was minimized by placing the support mast midway between the wide spaced reflector and dipole. Feeder lengths from each Yagi must be equal.

The system is illustrated in **Figure 1**. The outputs of the receivers are connected to the X and Y plates of the oscilloscope. Phase and amplitude are displayed directly on the screen, giving an exact picture of the phase and amplitude of signal on each feeder, and showing the polarization of the incoming signal. Signals being at audio frequencies, an expensive wideband oscilloscope is not required. Unfortunately, even oscilloscopes boasting 20-MHz bandwidth on the Y amplifiers often have only 1 MHz on the X amplifiers. This method of displaying polarization requires two identical receivers, with the audio taken from the output provided prior to the audio amplifiers and loudspeaker. Different receivers have different audio characteristics, resulting in audio phase shifts varying with frequency. This will not affect results on a constant carrier such as a beacon, but will prevent the polarization of SSB audio signals being shown.

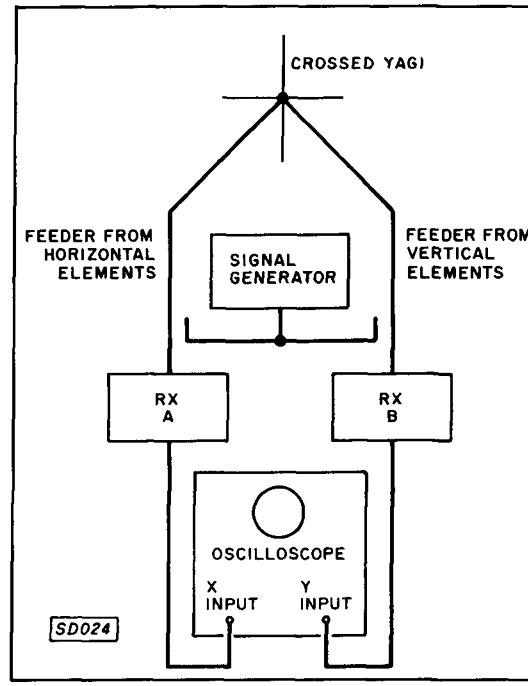


Figure 1. System diagram.

Waveform generation on the oscilloscope

The way in which the waveforms are generated on the oscilloscope is as follows:

Horizontal Polarization is received only by the horizontal elements of the antenna. They are connected, via receiver A, to the X plates of the oscilloscope. The display is therefore a horizontal line.

Vertical Polarization is received only by the vertical elements. Receiver B and the Y plates give the vertical display.

Forty-five Degree Polarization is received equally by each antenna, and is in phase. The display is therefore a 45-degree line.

One hundred and thirty five-degree Polarization is also received by each antenna, but is 180 degrees out of phase, giving a 135-degree line.

Circular Polarization is received equally by each antenna, but is 90 degrees out of phase — resulting in a circular trace.

Discrimination between clockwise and anticlockwise is not possible. The spot forming the circular trace, although rotating in opposite directions in each case, is moving far too fast for the difference to be visible.

Using the equipment

Using the equipment was quite fascinating and highly instructive. Stations in different parts of the world could be tuned to, and not only was the polarization of the incoming signal immediately visible, but the rate of change of polarization could be immediately seen — therefore the usefulness of polarization control could be assessed. The various beacons in different parts of the world were particularly useful, together with the constant carriers available from the American repeaters.

It was immediately apparent that the number of hops to the ionosphere and back was totally irrelevant. The polarization of signals must therefore be controlled by the last hop. This was a conclusion drawn before when turning the Polarphaser knob, but very graphically illustrated now. The rate of change varied with the time of day and the degree of opening of the band. Generally, the better the opening, the slower the rate of change. But the initial opening of the band was always characterized by a very slow rate of change. The polarization on the Cyprus beacon in the morning stayed constant for periods of up to a minute. The Brazil beacon on 28.270 MHz (which must be at least four hops) was a particularly constant signal, with a definite preponderance of vertical polarization which stayed constant for minutes at a time. The difference between using horizontal and vertical polarization on this particular beacon was some 4 S-points in favor of the vertical — far outweighing the usual increase of noise level prevalent with vertical polarization. Short term QSB, although showing on the S-meters of the individual receivers which were on fixed polarization, was nonexistent on the combined signals as displayed on the oscilloscope. This perfectly illustrates that the majority of QSB is caused by polarization change. Long-term QSB brought the overall signal down on both receivers, and was accompanied by a flurry of polarization changes. The pictures of the oscilloscope screen, taken while receiving the Cyprus beacon at 1600 on 19 April 1988, can be seen in **Photo A**.

The polarization of a Yagi antenna is only correct in the main beam. This is shown on the screen when receiving a local signal of fixed polarization. Rotation of the beam results in rotation of the polarization, to the extent that a horizontal Yagi may show vertical polarization off the back and circular at the sides.

How do we use this display?

The next thought is how, in practice, could use be made of this display. The requirement is not only to be able to see the optimum polarization for the particular signal, but to be able to use it for transmission and reception. Since the use of two receivers in addition to the one in use would be rather expensive, another means of obtaining the same result is necessary. The main cost of receivers these days lies in the sophistication of control provided by the microprocessor, and the mechanical design of the case and control knobs — none of which are needed for the polarization display receiver. Consumer ICs, at very reaonable prices, provide a complete receiver on one chip, and use could be made of these. A possible system, which might well be the

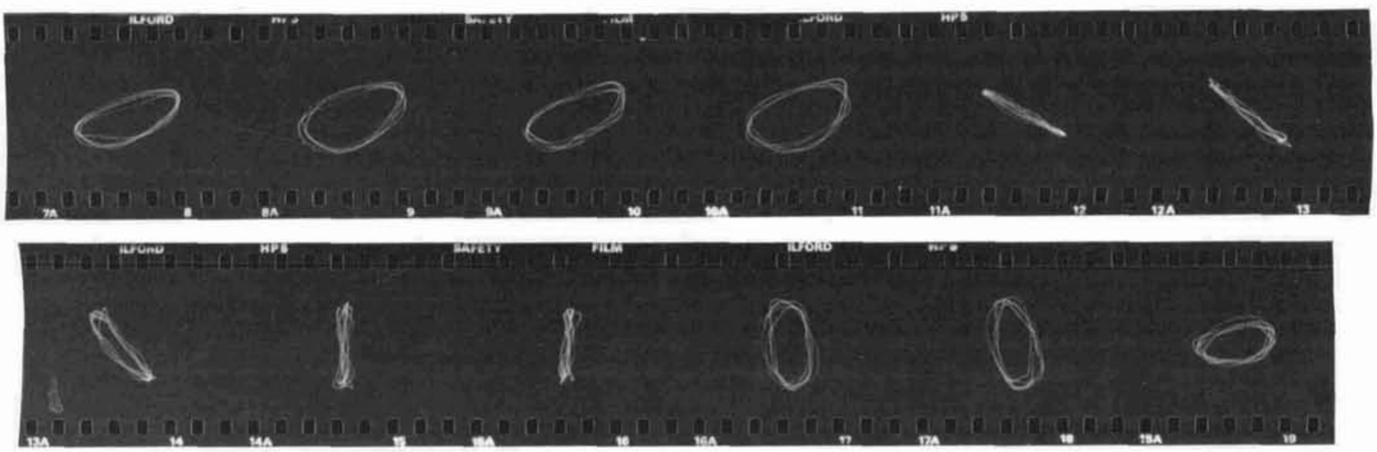


Photo A. Pictures of the oscilloscope screen taken while receiving the Cyprus beacon on April 19, 1988.

subject of a constructional article at a later date, is shown in **Figure 2**. Perusal of the circuit diagrams of a number of modern transceivers showed that the provision of a take-off point for the common local oscillator feed is quite feasible. The system of separate boards for oscillator signal generation, with coaxial feeders supplying the signal to the various parts of the set, provides a simple connection point without causing damage or needing modification in any way.

Many hours spent watching the quite fascinating changes in polarization of signals reflected from the ionosphere show that, although the polarization is often changing too rapidly for manual compensation to be feasible, there are many occasions when the ability to pick the optimum is very advantageous. Knowing the polarization of an incoming signal is an entirely new facet of Amateur Radio, and might even form part of the signal report. One thing is manifest: namely, that a horizontal beam is not the optimum. Circular polarization from a crossed Yagi would be far better.

My sincere thanks are due to Bill Wheeler, G3BFC, my near neighbor, for his extreme patience in providing low power local signals for calibration, and for his assistance in the erection of the test antennas. ■

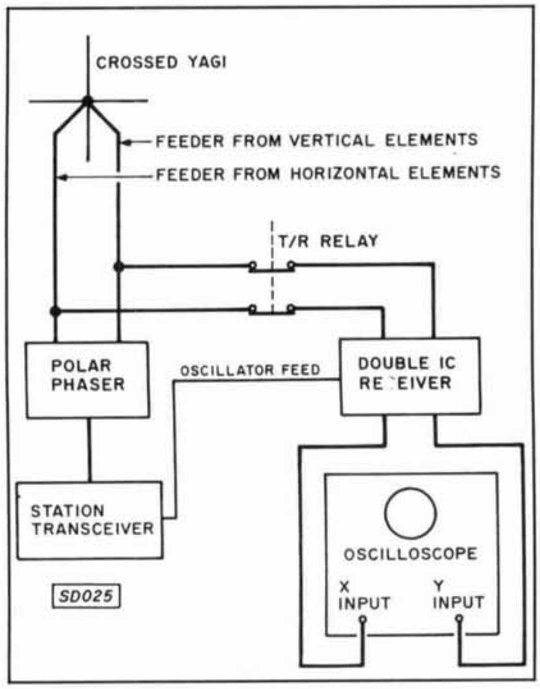


Figure 2. Diagram of possible updated system.

By Steve Powlishen, *K1FO*
816 Summer Hill Road
Madison, Connecticut 06443

432-MHz EME 1990s STYLE

Part 1: The revolution in technology

The requirements of 432-MHz EME* operation were discussed in HAM RADIO Magazine in 1982.[1] But because technology has progressed so significantly in the past eight years, it's time to take a new look at what's needed to make 432-MHz EME contacts. Many prospective EME operators still harbor the image of EME created in the 1950s – that it's a lifetime project to assemble an EME capable station. This isn't the case today. EME-capable arrays are now comparatively small, and it's no longer a complicated task to build an operational 70-cm EME station. Nowadays, the major obstacle to EME operation is fear of failure.

The basics of EME operation have been covered before,[2,3,4] so I won't delve into the topics of moon tracking or specific operating procedures. In part 1, I'll give an update on 432-MHz EME technology. In part 2, I'll discuss general recommendations on readily available equipment. The details of assembling a small, but effective, EME array will be given in cookbook form. This array should put an end to the argument, "I don't have room for an EME array."

Why EME?

Why have most VHF/UHF terrestrial mode weak signal operators wound up on VHF in the first place? They're attracted to the ability to communicate via a challenging mode, away from the guerilla tactics of the HF bands and the pandemonium that can beset FM repeaters. The down side of VHF/UHF weak signal terrestrial work is that the allure of DX can catch hold a little too hard. Because both ionospheric and tropospheric VHF band openings are rare occurrences, you can quickly become a slave to the VHF bands, and wind up sitting close to your radio at all times so you won't miss a rare opening. If you don't believe me, start making a list of the 6-meter friends you haven't seen or heard from since the solar flux crossed 200!

Eventually, before their families abandon them, most VHF DX fanatics are forced to temper their enthusiasm. Daily tropo operation now appears to be at its lowest point in many years. This is due, in part, to the lure of 6-meter F2 DX. It's also because of the attraction of Amateur satellites. Satellites finally provide a mode where you can work DX on a daily basis, in the peace and quiet of the VHF frequencies. Moreover, your off-the-air hours can now be free from the obsession of trying to find a better VHF location. A satellite doesn't care if you're at sea level or on the highest mountain east of the Rockies.

If you're willing to live with the stipulation that you can only operate when you can see the moon, then EME is for you.

Of course, satellites are repeaters. This means that grids, states, and countries don't count for awards requiring natural propagation modes. But, how would you feel about a mode that offers the prospect of worldwide VHF/UHF communications without repeaters or translators, doesn't require a great radio location, and can give you notice months in advance of the best days to operate. If you're willing to live with the stipulation that you can only operate when you can see the moon, then EME is for you. Like satellite operation, EME doesn't require high elevation arrays. Most EME arrays are only 20 feet above ground level. This is an added benefit which often alleviates problems you may have with local antenna restrictions.

*(earth-moon-earth)

Why 432 MHz

Most EME operators start with the 144-MHz band. There are some very logical reasons for starting on 2 meters. There's more 144-MHz tropo activity, so equipment is more readily available. Also, several 144-MHz EME super stations make it very easy for single Yagi tropo stations to make their first EME contact.

In spite of the positive aspects of 144 MHz, there are some very compelling reasons to make 432 MHz your EME band. They are:

- Lower noise level. Most 2-meter EME operators say their major complaint is noise. In urban locations, the noise is primarily manmade. Power lines, appliances, ignitions, electric fences, and many other sources plague the 144-MHz EME operator. There may be so many sources emitting noise simultaneously in densely populated areas, that they combine and appear as random noise which can't be effectively reduced – even by sophisticated noise blankers. My experience has shown that this urban RF noise pollution level is currently up to 30 dB lower on 432 MHz.

  For those 144-MHz operators in rural locations, the noise comes from cosmic sources. (Cosmic noise, also referred to as sky noise, is made up of radio emissions from galaxies as well as individual stars and novas.) Cosmic noise is the primary limiting factor in the receive capability for modern high performance 144-MHz EME stations in rural locations. Cosmic noise at 432 MHz is typically 11 dB lower than on 144 MHz. You simply don't need to consider sky noise for EME operation times at 432 MHz – except for 3 or 4 days a month. Even on these days, the degradation (due to the higher star noise at 432 MHz) is rarely worse than 4 dB, compared with up to 15 dB at 144 MHz. These problem days also occur when the sun is at southern declinations and EME operation is at its lowest point during the monthly lunar cycle.
- Less TVI and RFI. There seems to be less television and general electronic appliance interference at 432 MHz than 144 MHz. There are several reasons why this is true. First, the 432-MHz array has greater directivity. This means that it typically has fewer devices in its main lobe. Second, due to the shorter the wavelength, the field strength decreases more rapidly with the distance from the array. Third, most consumer electronic gadgets are usually less sensitive to 432-MHz RF energy than 144-MHz RF. Also, fewer cable systems operate at 432 MHz – reducing CATV system interference problems.
- Interference to the EME receive system is much lower at 432 MHz. At 144 MHz, many stations are plagued with spurious signals. Leakage from CATV cable, spurious signals from repeaters, and RF emissions from all sorts of consumer electronic devices (including computers, home security systems, scanners, and garage door openers) are usually far worse at 144 MHz than at higher frequencies.
- Arrays are much smaller. Given the same physical array size at 432 MHz, if similar transmit power and receiver performance are available on both bands, the lower sky temperatures and higher antenna gain will result in signals so much stronger that they'll more than overcome the higher 432-MHz EME path loss. Here's another way to look at it. To obtain the same signal strength, the array will be smaller than that required on 144 MHz.

Before I'm besieged with angry letters from 2-meter EME operators, I'll point out the drawbacks of 432-MHz EME operation. Libration* fading becomes significantly worse the higher in frequency you operate. The fading at 432 MHz requires either more operator skill or stronger signals to make EME QSOs. However, on many days, libration fading isn't much of a problem, and the days of low and high libration fading are predictable.

Interference problems also exist on 432 MHz. The primary offenders are shipboard radar and ATV transmitters. UHF television stations can also cause mixing problems with FM broadcast stations if you don't have selective enough preamplifiers. Equipment such as the high power amplifier is less available for 432 MHz, and is often more expensive. Feedline losses are higher, and antenna dimensions are more critical. Simply stated, you're allowed fewer mistakes at 432 MHz if you want to operate a successful EME station. Due to these real and other unsubstantiated reasons, there are fewer stations operating on 432 MHz.

The march of technology

There have been many significant changes in 432-MHz equipment during the last decade. This has led to what I consider a profound change in the profile of the average 432-MHz EME operator. Throughout the 1970s, 432-MHz EME was primarily the realm of the experimenter – someone capable of building all his equipment. He was a builder who was, in many cases, involved in

*Real or apparent very slow oscillation of a satellite as viewed from the larger celestial body it orbits.

the art of designing the system. Today, virtually all the components of a 432-MHz EME station are available commercially. Many articles have detailed easily reproducible equipment designs. This "off-the-shelf" EME station availability has transformed the average 432-MHz EMEer into an operator rather than a constructor. The lowering of the entrance requirements to EME has, in turn, resulted in an explosion of new EME activity. Here are some of the changes that have occurred in 432-MHz equipment.

Antennas

Because a greater number of different antenna designs are usable here than at any other frequency, 432 MHz is a unique EME frequency. At 903 MHz and above, parabolic dishes are truly the only practical EME antennas. (Yes, it has been proven that EME QSOs are possible with Yagi and horn antennas. However, these arrays are of questionable practicality for Amateurs at these frequencies). On the lower frequencies, like 144 MHz, the Yagi is the most suitable antenna. Dishes must be so large (typically, 28 feet or more in diameter) that they aren't feasible for most operators. There are still a handful (but ever declining number) of colinear devotees, but the Yagi is used by over 95 percent of 144-MHz EME operators.

Today, antenna choice at 432 MHz is split about 60 to 40 percent in favor of Yagi arrays. The balance is shifting gradually towards Yagi arrays, as more and more small Yagi stations become operational. Colinear and rhombic arrays have proven usable on 432, but have suffered the same fate that has befallen them at 144 MHz.

There have been minimal improvements in Amateur dish technology at 432 MHz in the last 10 years. Even today, deep dishes are still hard to feed at 432 MHz, while shallow ones pick up too much unwanted Earth noise from feed spillover. A more in depth look at the dish problem reveals that choosing a dish and feed is like any other engineering problem – there's no perfect solution. Instead, you must make the best trade off after considering several factors.

Receive performance is part of the problem. If you have a super quiet receive system, the lower noise pickup of a deep dish lets you take better advantage of the low noise receiver, and gives you further receive performance improvements. On the down side, given the power taper of practical 432 MHz feeds, the forward gain of the deep dish will suffer. Conversely, you can heavily illuminate a shallow dish with ease in order to maximize dish gain. In this situation, receive performance suffers due to excessive noise pickup. The best compromise exists somewhere in the middle, where the receive temperature drops faster than the gain of the dish and provides the optimum f/D given the feed and preamplifier.

Dishes in the 0.45 to 0.55 f/D range remain the best choice. The EIA dual dipole feed is recommended for more shallow dishes; a circular loop driven element over a plane reflector is a good choice for deep dishes. Some minor improvements have been obtained by making modifications to the EIA feed, and through the use of beam-forming rings on dipole feeds. N7ART has achieved performance gains with these kinds of feeds, though others have not. While EME QSOs have been made at 432 MHz with parabolic dishes as small as 10 feet in diameter (20.2 dBi typically), a 15-foot diameter dish (23.7 dBi) is the smallest advisable size to use. WØRAP was one of the first 10 stations to achieve 432-MHz WAS. He's worked 150 different stations using a 16-foot diameter TVRO dish (24.3 dBi)!

Today the EME enthusiast's greatest problem lies in deciding which of the wide selection of good designs to use.

I expect many new Yagi stations to become operational on 432-MHz EME. This will further shift the preference for array type systems towards the Yagi at 432 MHz. It's interesting to look at the tremendous changes in Amateur Yagi technology which occurred in the 1980s and see how they've led to this shift. In 1982, there were only two commercially manufactured Yagi designs suitable for EME use. Although both designs were advanced when they were introduced in the late 1970s, they aren't on a par with today's technology. I'll compare the 1980 Yagis to the 1990 models.

The RIW-19, introduced in 1977, was commonly used in North America until just a few years ago. It was a 5.6-wavelength (13-foot long) design with 17.0-dBi gain. By 1982, the original manufacturer was no longer making this antenna. The commercially produced RIW-19 varied from its original design. This change, intended to improve the driven element match, had the effect of reducing the front-to-back ratio from about 20 to 15 db. The RIW-19 was burdened with a second reflector, which really did nothing to either enhance or degrade the performance of the Yagi. Had the K2RIW Yagi been produced in its original

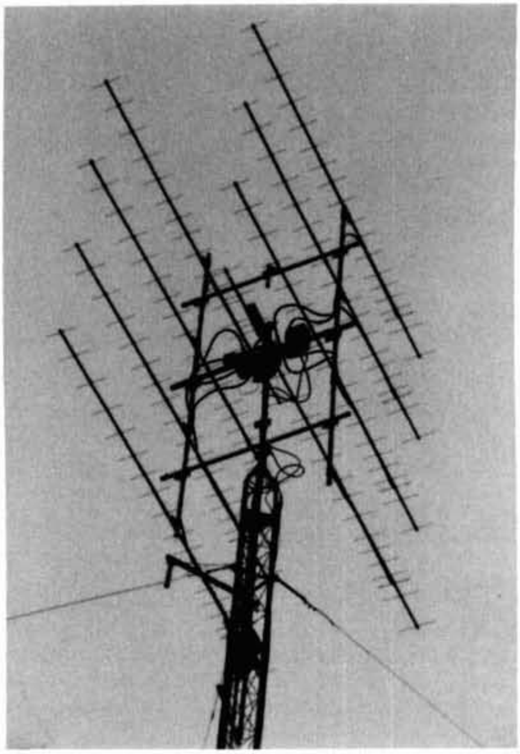


Photo A. The author's 8 × K2RIW-19 Yagi array from the early 1980s has the feedlines run along the booms and preamplifier box mounted away from the power divider, requiring an additional coaxial jumper. These practices, typical of the late 1970s and early 1980s did not result in the highest receive performance.

18-element, 5.2-wavelength, single-reflector form, it would have been an even more significant antenna. Its strictly empirical design had close to the maximum theoretical gain for its boom length in the single reflector form. **Photo A** shows a typical 8 × K2RIW-19 Yagi EME array.

The other commercial Yagi available in 1982 was the F9FT 21-element, 6.5-wavelength (15 feet long) antenna manufactured in France by Tonna. The F9FT design was one of the first to use the combination of progressively wider spacings and shorter elements. Although it's basically a good design and is still popular today, it had its own problems. The Tonna Yagi used an unbalanced feed which, due to the Yagi design, had a very low natural impedance. The Tonna's mechanical construction also used elements which were mounted through a conductive boom. In salt air or other corrosive environments, the elements could loose some of their electrical contact to the boom. This, in turn, would cause the center frequency of the Yagi to drop. These factors make the F9FT Yagi typically measure about 17.2 dBi (about 0.5 dB lower than its theoretical gain) and give it higher noise pickup than it should have. Tonna now sells an improved version which uses insulated elements and a balun feed.

Homebrew Yagis weren't in much better shape at this time. The DL6WU design was published in 1977 and 1982.[5,6] But in the early 1980s, it was still virtually unknown in the United States and hardly in use in Europe. In the United States, builders relied on a limited selection of designs consisting primarily of the 4.2-wavelength (10-foot long) NBS design,[7] the long-boom quagi design, and copies of the K2RIW 19 and 13-element designs. Except for the still unrecognized DL6WU design, other Yagis suffered from less than optimum gain per boom length, inferior patterns, and narrower bandwidths than today's.

Things are quite a bit different now. Today the EME enthusiast's greatest problem lies in deciding which of the wide selection of good designs to use. Three key advancements in Yagi design have created this situation. The first was the development of the DL6WU log taper design Yagi. This Yagi was the first available that could be made in virtually any boom length and still perform well. It meant that 432-MHz operators were no longer tied to the older set boom-length designs, but could pick boom lengths suitable to the total array scheme. The DL6WU design was also the first in which very long Yagis (greater than 7 wavelength) worked properly. As a final benefit, the DL6WU design had a radiation pattern that was significantly cleaner than earlier ones which used fixed element spacings and lengths. The DL6WU design has since become the basis for several commercial Yagi models (KLM, Hy-Gain, M ^ 2 and FlexiYagi).

The computer revolution also had a significant effect on the world of 432-MHz EME. In the mid-1980s, antenna analysis software (NEC, MININEC, and specific Yagi design programs) became available to large numbers of Amateurs. When this software was combined with the truly spectacular advances in microprocessor technology, antenna design capability took a quantum leap.

The new technology led to the evolution of improved versions of the DL6WU design, the development of performance enhancements to existing commercial products,[8,9] and the engineering of a computer-generated and optimized universal Yagi design.[10,11] Not only did these new Yagi designs incorporate vast improvements in the gain per boom length, they also had drastically cleaner radiation patterns than earlier versions. These new Yagis also had immensely wider gain and SWR bandwidths than the older ones. This made them behave much more predictably when stacked in arrays. The improved ability to stack arrays let builders obtain a stacking gain close to the theoretical 3-dB gain measurement when doubling the array size. When combined with very low loss feed systems, low temperature preamplifiers, and higher stacking gain, these new clean pattern arrays resulted in total EME system receive improvements far in excess of the forward gain increases of the new Yagis alone.

The important part of EME antenna technology is total array performance. The 1980s were a time of great progress in array design. At the beginning of that decade, most Amateurs were still clinging to inflated Yagi gain figures. This often led to the use of excessively wide stacking spacings which, in turn, degraded receive performance. Phasing lines were typically RG-8 or RG-8 type foam dielectric cables. The phasing lines were usually arranged so they ran neatly down the Yagi booms and stacking frames to power dividers located at the center of the array. While aesthetically pleasing, this arrangement resulted in excessively long phasing lines, which further degraded receive performance. Quite often, pictures of old EME arrays show the relay and preamplifier box mounted down on the tower behind the flex lines around the rotors—another configuration which deteriorated receive capability.

The performance increases of the 1980s preamplifiers and Yagis demanded better practices. Improved Amateur antenna range measurements, which were confirmed by computer analysis, finally turned antenna designers and builders into honest men and pulled Yagi stacking distances in line. As the individual Yagis were designed with better patterns, better arrays were created. Yagi builders began mounting the phasing lines and power dividers out the back of the array, in or near the plane of the driven elements, in order to create the shortest possible phasing line lengths and obtain correspondingly lower losses. A phasing line war ensued as stations moved up from RG-8 type cables to aluminum and copper-jacketed semiflexible lines. Half-inch Heliax® became a phasing line staple; some stations even used phasing lines as large as 7/8 inches in diameter. In the mid-1980s, DL9KR demonstrated how open wire line could be used practically for phasing lines. When lines were made with a large size wire like no. 8, losses were as low as those obtained with 7/8-inch Heliax. Better still, this wire was very inexpensive and simpler to use than coaxial lines and power dividers with lots of connectors. EME operators learned to move their preamplifiers off the tower and into the center of the array to further reduce phasing line losses.

Preamplifiers

The success of an EME station at 432 MHz and higher frequencies largely rests on the operator's ability to take advantage of the low sky temperatures at these frequencies. A low temperature system depends upon three factors: a low noise preamplifier, very low loss relays and jumper cables (and phasing lines in the case of stacked Yagi arrays), and an array which picks up a minimal amount of sky and earth noise. Progress in 432 MHz and above EME operation was limited by the availability of low noise preamps. In the 1960s, parametric amplifiers were required for the best performance. Unfortunately, they were complicated to build and even harder to tune. As low noise bipolar transistors became available in the early 1970s, a new wave of 432-MHz EME operators joined the ranks of successful UHF EMEers. Although these bipolar amplifiers were inferior to properly working parametric amplifiers, they were much easier to construct and tune up, and could be mounted without difficulty at the array to overcome feedline losses.

The GaAsFET preamplifier was introduced to the Amateur community in 1978.[12] The transistor used in this preamp cost nearly $200. The amplifier itself was hard to tune properly without good test equipment, and even then it was only conditionally stable. The actual noise temperature claimed was 50°K (0.7-dB NF). This may have been somewhat optimistic, but was typically less than half the noise temperature of most bipolar devices used by 432-MHz EME operators at the time. To make matters worse, it was easy to burn out the FET. It wasn't until the early 1980s, that GaAsFET preamplifiers were widely used. Even then, the devices were in the $50 price range and had noise temperatures above 40°K (0.56 dB NF). Good commercial GaAsFET preamps were still a rarity, as the designers of most commercial models took unacceptable liberties. Consequently, the "average" EMEer usually couldn't even get close to obtaining state-of-the-art receive performance.

The year 1990 brought low cost, high performance preamplifiers to the average EMEer. But almost as important as an operator's access to today's devices, is his need to understand noise figure measurement errors, preampifier stability, and proper preamplifier tuning.[13] If you have this information, you'll be able to make the preamplifier work just as well when it's connected to the array as it did on the test bench. This often was not the case in the early days of EME.

There's also more information available on device noise figures and the contribution of input circuit losses. The Mitsubishi MGF-1302 is currently one of the most popular GaAsFET devices. These devices cost just $7.50 each in single lot quantities. The MGF-1302 is capable of noise temperatures below 30 °K (0.4 dB NF) when used in low loss input circuits. The more expensive MGF-1402 ($14) and MGF-1412 ($26) can achieve noise temperatures of less than 25 °K (0.35 dB NF). When used with a low loss cavity input circuit,[14] the MGF-1412 has exhibited noise temperatures (at ambient room temperature) as low as 22 °K (0.3 db NF).* These devices are all single-gate FETs. Dual-gate GaAsFETs are now available for about $1. With noise temperatures over 75 °K at 432 MHz, these dual-gate FETs aren't suitable for first stage preamplifiers in EME systems.

The performance figures above were obtained by measuring the preamplifiers at ambient room temperature. Other measurements were made after cooling the preamplifier. At liquid nitrogen temperature (77 °K), the MGF-1412 can exhibit noise temperatures under 10 °K (0.15 dB NF). Recently, HEMFETs (High Electron Mobility Field Effect Transistors) have become available. At room temperature, these devices aren't any better than the higher quality GaAsFET devices — at least at low frequencies like 432 MHz. When cooled they appear to have performance gains that exceed GaAsFETs, and may approach a noise temperature of 5 °K at 77 °K ambient temperature.

The true benefits of these low noise temperatures are apparent when they are combined with modern low sidelobe arrays. My current system (16 × 3.6-wavelength Yagis) has an array noise pickup of about 32 °K when pointed at a cold sky. Total phasing line and relay losses are about 16 °K. If I had a 44 °K (0.6 dB NF) receive system temperature (typical for a 40 °K first stage), my total system temperature would be 92 °K. If I lower my receive system temperature to 25 °K (MGF-1412 in a cavity circuit), my system temperature is 73 °K. The resultant receive signal-to-noise improvement is 1.0 dB — quite a difference! **Appendix 1** provides information on performing these signal-to-noise ratio calculations.

Transmitter power

The change in Amateur power level limits to 1500 watts output had a much more significant effect at 432 MHz and above than at 144 MHz. At the higher frequencies, final amplifier efficiency is generally poorer. This made the old system of 1-kW DC input power a greater restriction at 432 MHz. In the 1970s, the Eimac 8877 gave a large segment of 144-MHz EME operators the ability to run large amounts of output power easily at 144 MHz. Many 144-MHz United States EME stations applied for and received Special Temporary Authorization (STA) to run high power output. Others assumed that the high power STAs applied to EMEers in general and, in a defacto fashion, to EME stations. Thus at 144 MHz, the 1500-watt output level merely formalized the status quo.

In July 1983, when the FCC adopted the 1500-watt PEP output power regulation, there were few 432-MHz stations capable of running much more than 1000 watts output. The venerable K2RIW parallel kilowatt[15] was the standard amplifier at 432 MHz. The new power regulations provided the impetus for many 432-MHz EME operators to increase their power level. The K2RIW amplifier was originally designed to run 1 kW input (600 to 700 watts output). The amplifier can run 1100 watts (or so) output when 4CX250Rs or 8930s are used. It has also been adapted for use with the Eimac 8874 triode and can easily reach 1200-watts output with the 8874. Use of the Eimac 8938 (the UHF version of the 8877) in an K2RIW-style stripline was described by W3HMU in June 1977.[16] The 8938 can easily make 1500 watts output power at 432 MHz. Unfortunately, the cost of low volume specialty tubes like the 8938 has risen in recent years, putting them out of reach of most Amateurs. The Eimac 3CX800A7 is capable of 1500 watts output from a pair of tubes at 432 MHz.[17] Other UHF tetrodes like the RCA 7213 are often available surplus at reasonable prices. More information on suitable 432-MHz power tubes can be found in QEX.[18]

The higher power levels make EME communication easier than ever before, but super power isn't a requirement for 432-MHz EME. The average 432-MHz EME station runs 800 watts at the array from a K2RIW-type amplifier. Approximately a third of the stations are fortunate enough to

*AA4TJ measured a 17 °K noise temperature at 432 MHz using the NARO hot-cold measurement system. Although the results are very repeatable, measurement accuracy probably isn't much better than 5 °K. Hence the 22 °K temperature is accepted as a worst case measurement.

have large amplifiers capable of approaching 1500 watts at the array. The balance of the stations make do with lower power levels. QSOs are made frequently with power levels as low as 200 watts. Some of the larger stations enjoy QRP operation from time to time. QSOs between the larger stations (K5JL's 28-foot dish and WB0TEM's 32-foot dish, for example) have been made with as little as 10 watts on both ends. With my own station (27.5 dBi gain), I've detected echoes during good conditions using as little as 10 watts at the array. These echoes are quite often copyable with 25 watts at the array. Full power echoes (1250 watts at the array) often result in echoes 14 dB out of the noise (200-Hz bandwidth).

Single Yagi EME

The most spectacular result of all this new technology is the single Yagi EME QSO. Ten years ago, even the most ardent 432-MHz EME fan would have laughed at the prospect of single Yagi QSOs to normal Amateur stations. (Single Yagi stations have made QSOs with the 1000-foot Aricibo dish as far back as 1964.) These QSOs are even more significant when you consider that many of the stations worked by these single Yagi owners aren't Amateur super stations. WD5AGO has heard four single Yagi stations. All of the single Yagi EME stations listed in **Table 1** have made at least one random QSO. All of WD5AGO's QSOs were made during a 3 month period! **Table 1** lists some of the successful 432-MHz EME single Yagi stations as of early 1990.

The more successful single Yagi stations have used antenna elevation rather than simply working stations on the rising or setting moon. It's usually difficult to realize appreciable ground gain at 432 MHz. Elevating the antenna significantly lowers Earth noise and reduces atmospheric attenuation and ionospheric scattering. It also greatly increases the available moon window. N9KC uses manual elevation control. This isn't as inconvenient as it may sound. If the elevation of a single Yagi station is set to the maximum for the moon on a given day, an operator can obtain hours of moon window without needing to make additional elevation adjustment as the moon passes through due south.

Polarity

All forms of space communications – including satellite operation – are affected by the phenomenon of Faraday rotation. Faraday rotation is the actual rotation of the polarization of a radio wave due to the electromagnetic fields in the Earth's ionosphere. This effect can be especially troublesome during years of high solar activity. The amount of Faraday rotation depends upon the frequency of operation. The higher the frequency, the lower the amount of rotation. At 432 MHz, up to 360 degrees of rotation is possible during times of high solar activity. At 144 MHz, seven or more 360-degree rotations are possible, while at 1296 MHz more than 90 degrees of rotation is rare.

There is a second polarization effect known as spatial polarization, which is basically a simple 3-D geometry problem. In its simplest essence, two stations a significant distance apart can have their arrays looking at the moon at substantially different polarity angles – even though they may be the same polarity sense in earthbound terms. In general, the farther apart the two stations are in longitude, the less of a common polarity window they'll have if they're using standard azimuth and elevation movement of the arrays to track the moon.

Station	Number of Different EME Stations Worked	Antenna	Power
I5TDJ	21	9-wavelength DL6WU type	700 watts
WD5AGO	19	Modified Hy-Gain 70-31	600 watts
N9KC	9	Stock KLM 432-30LBX	500 watts
I5CTE	5	9-wavelength DL6WU type	700 watts
W2WD	4	9-wavelength W2WD design Yagi	700 watts
KD5RO	4	Stock 424B (7.5 wavelength)	400 watts
DJ5BV	3	6.6-wavelength F9FT Yagi	500 watts

Table 1. Successful single Yagi 432-MHz EME stations.

The combination of these two effects leads to classic EME QSO problems. Either: (1) neither station hears anything; (2) one station hears the other, the other station hears nothing; (3) the reverse of no. 2; or (4) both stations hear each other, and a contact is made with ease. A change of polarization alignment during a schedule can be even more frustrating. Frequently, one station hears another, but then the situation reverses rapidly. If you're quick on the draw, you may complete a QSO during the transition, but often no contact is made. The mechanics of this situation were first clearly described by K9XY.[19] Several years later, KL7WE offered a more detailed explanation of the problem and how to work around it.[20]

Spatial polarization problems can be eliminated if both stations use polar mounts and the same polarization sense with respect to the polar axis. Polar mounts won't fix Faraday rotation problems, nor will they fix the spatial problem if one of the stations uses an Az-El array mount. Polar mounts also aren't practical for arrays other than parabolic dishes, colinears, or other arrays which can be mounted from the rear.

The use of circular polarity is an obvious way to eliminate both of these polarization effects. Although circular polarization is pretty much the standard at 1296 MHz and above, it's not that practical at 432 MHz and below. When the signals reflect off the moon, the polarity sense changes. Hence stations must transmit in one sense (for example, left-handed) and receive in the other sense (right-handed). If you use a parabolic dish antenna, it's relatively easy to construct a feed which has two sets of feed elements 90 degrees apart. These can then be combined via a four-port hybrid to give one polarization sense on transmit and the other on receive.

Those who use other types of arrays aren't as fortunate. I'm frequently asked about helix arrays. These arrays simply aren't usable for EME, unless you construct two arrays or two sets of elements in opposite polarization senses. Yagi arrays can be built with crossed elements to give any combination of vertical, horizontal, left-hand, or right-hand circular polarization. While this method is commonly used for satellite communications, satellite communication isn't a weak signal mode. Amateur satellite signals are typically 20 db stronger than EME signals. (Most 432-MHz capable stations could easily hear the noise floor of the 435-MHz satellite translators.) The combination of the phasing line and switching circuit losses, plus the pattern distortion that stacking frames introduce when Yagi elements are polarized in the same plane as the frames, will simply destroy the receive performance of such a cross-polarized Yagi array.

The final obstacle to the use of circular polarization at 432 MHz is that the majority of active stations would have to switch to that mode as the 3-dB theoretical (often in practice it seems worse than 3 db) polarization mismatch becomes intolerable. Over the years, several dish stations have tried circular polarization on 432 MHz, only to return to linear polarization.

The most satisfactory solution to the polarization dilemma at 432 MHz has been to use switchable or rotatable polarization. Dish operators generally prefer to use mechanical motion to rotate the polarity of the feed. With 180 degrees of rotation, they have the ability to peak a signal to its polarization peak at any time. Often this polarization peak isn't horizontal or vertical with respect to the Earth. Some stations use two sets of feed elements – one set vertically and the other horizontally. These feed elements are then switched to find the best signals. However, this arrangement suffers up to a 3 dB loss if signals are polarized 45 degrees out.

Yagi and similar arrays have taken a somewhat different path of development. Early on in 432-MHz EME history, arrays like W1JR's colinear[21] used mechanical polarity rotation. Other short Yagi arrays with similar arrangements were contemplated. When the first long Yagis with good performance appeared in the late 1970s (the F9FT-21 and the RIW-19), the thought of building Yagi arrays with polarization rotation faded from the minds of most EMEers. These Yagis, and the better ones which followed, allowed arrays to be built which had enough gain that they could still make EME QSOs – even with substantial polarization misalignment and the resulting signal strength losses. **Photo B** shows NC1I's large and complex K1FO-32 Yagi EME array.

The rush to acquire large Yagi arrays slowed in the late 1980s, as the owners of even the largest fixed polarization arrays discovered there were times when other fixed polarization stations couldn't be worked. Two solutions to the problem emerged. KDØGT came up with the first. He used a moderate number of long Yagis (8 x 10.4-wavelength Yagis) and rotated the booms of the individual antennas to change polarization. KDØGT's system had several

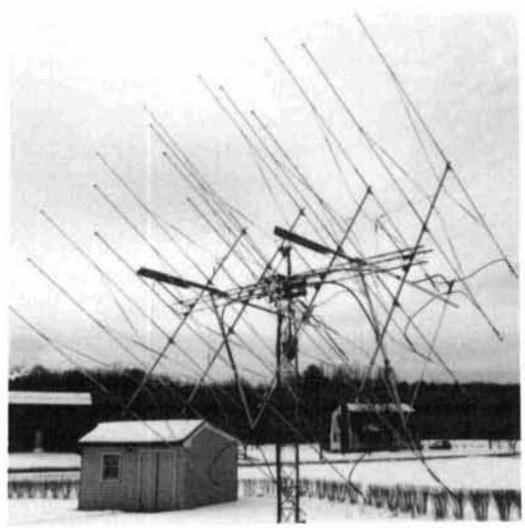


Photo B. NC1I's 16 × 10.4 wavelength (24 foot long) K1FO-32 Yagi array has over 31 dBi of gain and typifies the "make it bigger" approach of the mid-1980s. Note that the 1/2-inch phasing lines and power dividers for the groups of four Yagis are mounted at the rear of the array to minimize losses. The bays of four Yagis are connected to the center power divider through very low loss 7/8-inch Andrew Heliax® coaxial cable.

drawbacks. Because optimum E and H plane spacings aren't identical, you must choose a compromise which will give less than optimum performance in any polarization. Feedlines must be flexible and longer to allow for the polarization movement. This means that their losses will be higher, limiting receive performance. Mechanically, it's difficult to keep the individual Yagis synchronized in rotation. KDØGT's array used four TV-type rotors to drive the eight Yagis via lever arms. It's difficult to get even 90 degrees of rotation with this method — let alone the desired 180 degrees of rotation. A final problem involves the interaction of the stacking frame with the Yagis when the elements align with the frame. KDØGT's experience has been in agreement with computer analysis and antenna range measurements. This information indicated that if the Yagi's polarization plane moves within 30 degrees of parallel to a metal stacking frame element, the pattern of the array becomes severely distorted, again limiting performance. In spite of these seemingly overwhelming obstacles, KDØGT's rotatable Yagis were a success and led others to emulate his work.

The other approach to solving the fixed array problem is to rotate the entire array to obtain variable polarization orientation. The rear mounting of the array required to accommodate the rotation places all of the pattern offending hardware like phasing lines, stacking frames, and rotors behind the array. This allows for maximum performance at any polarity angle. Because the individual Yagis remain in a fixed position relative to each other, you can use semi-rigid or rigid low loss phasing lines to maximize the array's performance. WBØYSG tried this approach in the mid-1980s. He used a mechanically complex but impressive-looking mechanism to rotate an array of sixteen RIW-19 Yagis. KL7WE built a more practical array a few years later. His array used a new generation of high performance Yagis. It consisted of sixteen rear-mounted 2.5-wavelength DL6WU-type Yagis. Although the array is physically quite small and the gain is a modest 25.5 dBi, KL7WE's results have been impressive and adopted by other stations. My 16 × 3.6 wavelength Yagi array with polarity rotation is shown **Photo C**.

These new style arrays make for an interesting subtlety in EME operation. An operator with a large array can be a bit sloppy in his operating because he can make up for it in brute force. A skilled operator using a small dish or a small Yagi array with adjustable polarization, can often keep pace

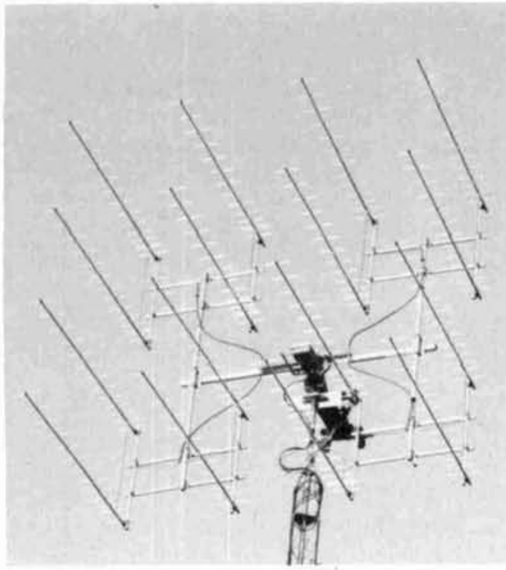


Photo C. The current K1FO EME array uses small rear-mount Yagis to facilitate mechanical polarity rotation. Phasing line losses are kept to a minimum through the use of a combination of open wire and Heliax phasing lines. Note that the preamplifier/relay box is mounted directly to the center power divider. This physically small, but electrically high-performance array is becoming the norm for the 1990s.

with the big boys. He must, however, pay attention to the direction in which he points his array on transmit. He must also constantly scan with his polarization for signals on receive. To do so, he must use the current received polarization and the predicted spatial polarization to determine the current amount of Faraday rotation. The operator can then work backwards to establish the best transmit polarization (remembering the nonreciprocal polarity conditions which frequently occur). Reference 20 describes a straightforward method using nomographs to facilitate this polarity prediction.

An EME operator who's used to operating with a fixed polarization array may be very surprised when he first experiences the benefits of polarity control. An operator who typically digs signals out of the noise, may feel like he's cheating when he's able to simply peak the signals up and carry on in relative ease. The problems of cross polarization have been known for almost 30 years. In spite of this, the rediscovery by the masses (dish users were often accused of hiding the truth) of this nearly lost knowledge (along with the modern solutions to the problem) has become one of the more significant EME events of the late 1980s.

Modes of operation

The great increase in EME signal strengths has generated still another notable change in 432-MHz EME operation. At the start of the 1980s, most QSOs were minimal CW contacts which used the standard TMO reporting system and were usually facilitated by prearranged schedules. Today the majority of QSOs are made on random operation by calling CQ. Most QSOs use the standard RST reporting system and exchange quite a bit of other information. This information exchange has progressed from simple reports and cordialities to information on polarization, fading, general conditions, word of other stations heard or worked that day, equipment used, and grid squares. EME capability has become a must have for the big multi-operator stations in the VHF contests if they want to work a few extra multipliers.

CW is still the dominant transmission mode, but SSB QSOs are now common on any activity weekend. Several European stations who only operate SSB, have appeared in recent years. Today schedules and the TMO reporting system are reserved for working new stations with minimal equipment. Experiments with other modes like packet communications and slow scan television have been tried, but haven't been very successful. The primary obstacle to using these modes is libration fading. The rate and depth of the fades severely distorts the signals, even when strengths may appear to be adequate. SSB operation has proven more usable than it first appeared it would be. It's typical for SSB communication to require 10 to 13 dB more signal strength than CW slow Morse code in order to communicate. The information exchange rate is so fast on SSB that it's possible to work around the libration fades by simply repeating the information several times. Probability dictates that with enough repeats, a signal strength peak will come along and get the information through.

Several EME operators have considered Digital Signal Processing (DSP). W3IWI has claimed successful Amateur DSP-based EME communications. Theoretically, 20-dB signal-to-noise gains are feasible, making EME possible with tiny arrays and low power. DSP requires stringent frequency and time coordination. The probability that DSP will be accepted by the EME community isn't very high, unless someone comes up with a system which allows operators to make random QSOs.

Conclusion

For several decades, the goal of the EMEer may have been to simply make an EME QSO. As we enter the 1990s, EME operation is becoming a given for the advanced VHF/UHF operator. These days, the goals of the 432-MHz EME operator are a bit loftier. They include attaining Worked All States (WAS), pursuing DXCC with country totals crossing the halfway point, making regular SSB QSOs, working the most different EME stations, winning the EME contest, or simply putting the biggest signal on the band. Yes, EME operation has changed from a colossal technical project into a mode which is just a lot of fun. ■

Appendix 1. Noise temperature and signal-to-noise calculations.

The concept of noise temperature is very useful in determining how the changes in a system will improve receive performance.

Total system temperature:
$Ts = Ta + Tl + Tr$

Where Ta = the array temperature
(that is, Earth and sky noise reception)
Tl = the phasing line loss, relay loss, etc.
Tr = the noise temperature of the receiver

The change in receive signal to noise is the 10 * log10 of the ratio of the total noise temperature. For example:

System 1: Ta = 50°K, Tl = 30°K, Tr = 50°K, therefore Ts = 130°K
System 2: Ta = 32°K, Tl = 15°K, Tr = 25°K, therefore Ts = 72°K

The relative signal-to-noise ratio between the two systems is:

$10 * \log10 (130/72) = 2.6$ dB

Assuming that the gains of array 1 and array 2 are identical, array 2 will hear signals 2.6 dB stronger than array 1. The difference in temperatures can be obtained from using the proper spacings to change Ta from 50°K to 32°K, changing the receiver from a 0.7-dB noise figure to a 0.36-dB noise figure, and changing the phasing lines from RG-8 to 1/2-inch Heliax. Without making the array any larger, you've almost doubled the receive performance. As you can see, even small improvements are important to EME systems.

References

1. Joe Reisert, W1JR, "Requirements and Recommendations for 70-CM EME," *HAM RADIO*, June 1982, pages 12-18.
2. *The Radio Amateur's Handbook*, 67th Edition, American Radio Relay League, Newington, Connecticut, 1990 Chapter 23, page 23-30.
3. Joe Reisert, W1JR, "VHF/UHF World: Minimum Requirements for 2-meter EME Part 1," *HAM RADIO*, August 1987, page 39.
4. Joe Reisert, W1JR, "VHF/UHF World: Minimum Requirements for 2-meter EME Part 2," *HAM RADIO*, September, 1987, page 65.
5. Gunter Hoch, DL6WU, "More Gain with Yagi Antennas," *VHF Communications*, April 1977. VHF Communications is a West German Publication available from TimeKit, P.O. Box 22277, Cleveland, Ohio 44122. Back issues are available.
6. Gunter Hoch, DL6WU, "Extremely Long Yagi Antennas," *VHF Communications*, March 1982.
7. "WØEYE 15-Element 432-MHz Yagi," *The Radio Amateurs VHF Manual*, 3rd edition, American Radio Relay League, 1972, page 210.
8. Steve Powlishen, K1FO, "High Performance Yagis for 432 MHz, *HAM RADIO*, July, 1987, page 8."
9. Steve Powlishen, K1FO, "Improved High Performance Yagis for 432 MHz," *HAM RADIO*, May 1989, page 9.
10. Steve Powlishen, K1FO, "An Optimum Design for 432-MHz Yagis," *QST*, December 1987, page 20.
11. *The Radio Amateur's Handbook*, 67th Edition, 1990, Chapter 33, Newington, Connecticut, Chapter 33, page 33-16.
12. Shigeru Sando, JH1BRY, "Very Low-Noise GaAsFET Preamp for 432 MHz," *HAM RADIO*, April 1978, page 22.
13. Raaner Bertelsmeier, DJ9BV, "Low Noise GaAsFET Preamps for EME: Construction and Measurement Problems," *DUBUS*, Volume 17, No. 4/1988. *DUBUS* is a West German Publication available through KAØHPK, P.O. Box 126, St. Mary of the Woods, Indiana 44876.
14. "WA7CJO Cavity Preamplifier," *432 MHz and Above EME News*, Volume 17, No. 4, published by Al Katz, K2UYH, April 1989.
15. "High Efficiency 432-MHz Kilowatt Amplifier," *The Radio Amateur's VHF Manual*, 3rd edition, American Radio Relay League, 1972, page 299. This information originally appeared in *QST*, April and May 1972.
16. Tony Souza, W3HMU, "432-MHz Power Amplifier Using Stripline Techniques," *HAM RADIO*, June 1977, page 10.
17. Russ Miller, N7ART, "The Audry II UHF Amplifier, *VHF/UHF and Above*, March 1988, page 6. *VHF/UHF and Above* is published by KAØHPK, P.O. Box 126, St. Mary of the Woods, Indiana 44876.
18. Steve Powlishen, K1FO, "Characteristics of Some Common Power Tubes at 432 MHz," *QEX*, No. 81, November 1988, page 3.
19. C.H. Hustig, K9XY, "A Closer Look at Faraday Rotation," *432 MHz and Above EME News* (K2UYH, publisher), November 1984.
20. Tim Pettis, KL7WE, "Spatial Polarization and Faraday Rotation," *Proceedings of the 22nd Conference of the Central States VHF Society*, American Radio Relay League, Newington, Connecticut, 1988.
21. Joe Reisert, W1JR, "VHF Antenna Arrays for High Performance," *QST*, December 1974, page 38.

By Yardley Beers, *WØJF*
740 Willowbrook Road
Boulder, Colorado 80302

THE 5/4-WAVELENGTH DIPOLE: A REVIVAL

Design data for long-wire antenna with matching transformer

Here are some single-wire antennas which have approximately 3-dB gain over half-wave dipoles. They are known as 5/4-wavelength dipoles. These antennas are especially useful in instances (like portable operation) where multi-element beams aren't practical, but where you do have a wide choice of antenna locations. I've found them easier to transport when making overseas trips with only carry-on baggage. In the past, when Amateurs used open-wire feeders, such antennas (then known as double-extended Zepps) were popular.

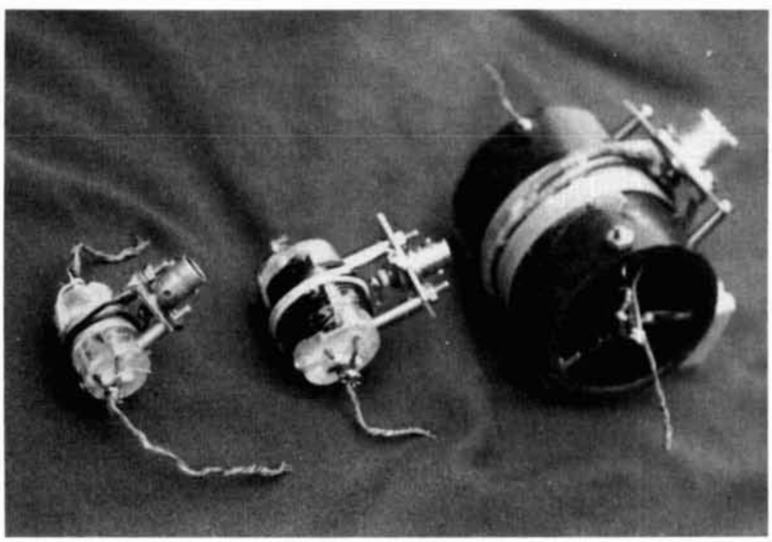


Photo A. The two smaller transformers are for 28 MHz, for 25 and 100 watts, respectively. They are wound on sections of plastic rods, which also serve as compression-type center insulators. The largest transformer is wound on tubing, with a compression-type insulator inside. The tubing is supported from the antenna by radial wires. This is a mechanical model that illustrates a construction method which can be used with larger transformers. It has never been connected to an antenna. Its approximate size is appropriate for a transformer for 14 MHz and 1500 watts.

Now that the use of 50-ohm coaxial line has become standard practice, these antennas have been almost forgotten. However, the 5/8-wavelength vertical antennas derived from them are well known. A recent rendition of the 5/4-wavelength dipole appeared in a July 1989 CQ article by Schultz.[1] However, he still uses a section of open-wire line, and introduces a transformer to make a transition to coax—a cumbersome arrangement.

I'll describe the design and construction of compact transformers which match the impedances of these dipoles to 50-ohm coax, as illustrated in **Photo A**. In some cases, they may be combined with the center insulator. Practical details for construction of transformers made for the 21, 28, and 50-MHz bands are given in **Table 1**.

Properties of 5/4-wavelength dipoles

Because of end effects, the physical length of a half-wave dipole is slightly shorter than the half wavelength in free space. The reduction in length depends upon the ratio of the length of the conductor to its diameter. The effect is small for a wire halfwave dipole at HF and can be compensated for by making the physical length 5 percent shorter than the free-space half wavelength. I assume that in feet, not in percent, a 5/4 wavelength dipole must be shortened by an equal amount. The physical length in feet, S, is given by:

$$S = 1218/F \quad \textbf{(1)}$$

Where F is the frequency in MHz.

Before I built any transformers, I determined theoretically approximate values of the radiation resistance and reactance.

I have a general program, which uses the common assumption of a sinusoidal current distribution, for calculating the radiation resistances of linear antennas of arbitrary lengths with various types of loading when in empty space. When applied to this type of antenna, the program gave a value of about 210 ohms. Using the methods discussed in an earlier article,[2] I found the total reactance to be twice the characteristic impedance, as defined there.

Using my latest data, I estimated the reactance to be 1000 ohms, capacitive. I've done some experimenting which supports the validity of these theoretical estimates.

Because of the high radiation resistances, these antennas are very efficient. Even if the transformers are wound with thin wire on small diameter forms, the Qs must be at least 100. In such cases, the loss resistances must not be larger than about 1000/100 = 10 ohms, and the efficiencies must be at least 210/(210 + 10) = 0.95, or 95 percent.

The antennas I've built have broad bandwidths, though not as large as that of a half-wave dipole. This is demonstrated by **Figure 1**, which gives plots of SWRs for a 28-MHz 5/4-wavelength dipole and for a half-wave dipole as functions of frequency.

There are two plots for the 5/4-wavelength antenna. One places it horizontally about 20 feet above ground. The other plot is a V with the apex about 20 feet above ground, as might be the case in portable operation, with the center supported by a rope thrown over the branch of a tree.

In the V configuration, the frequency of minimum VSWR was shifted lower—although the VSWR remained reasonably low over an appreciable part of the band. To save weight, I don't use hard end insulators on portable antennas, but depend upon the rope (usually Teflon®) to supply the insulation. I make small loops at the ends. When operating portable, it's simple to raise the resonant frequencies by twisting the ends back on themselves a few inches. This makes it easy to bring the resonance into the band when your antenna is in the V configuration.

The pattern is symmetrical and consists of a principal maximum at 90 degrees to the wire, nulls at about 54 degrees to the wire, and secondary maxima (about 10 dB down from the principal one) at about 32 degrees to the wire. *The ARRL Antenna Book*[3] and other sources give polar plots. The rectangular plot shown in **Figure 2** is more useful when discussing the details. This plot shows only one quadrant of the pattern. On the same graph, you'll see the pattern of a half-wave dipole with its scale adjusted to correspond to the same radiated power. The

Frequency (MHz)	Power (watts)	Diameter (inches)	Spacer (inches)
21.2	25	1.00	5/8
28.3	25	0.75	3/8
28.3	100	1.00	5/8
51	25	0.75	1/2

Secondary turns	Secondary wire no.	L (μH)	Reactance (ohms)
13	24	6.57	875
12	24	3.97	706
13	20	4.04	718
8	24	2.15	675

Notes:
1. All wire plastic coated.
2. Primaries wound with no. 20 plastic-coated wire. The 21.2-MHz primary has 3 turns widely spaced. The other primaries have 2 turns closely spaced.

Table 1. Transformer data.

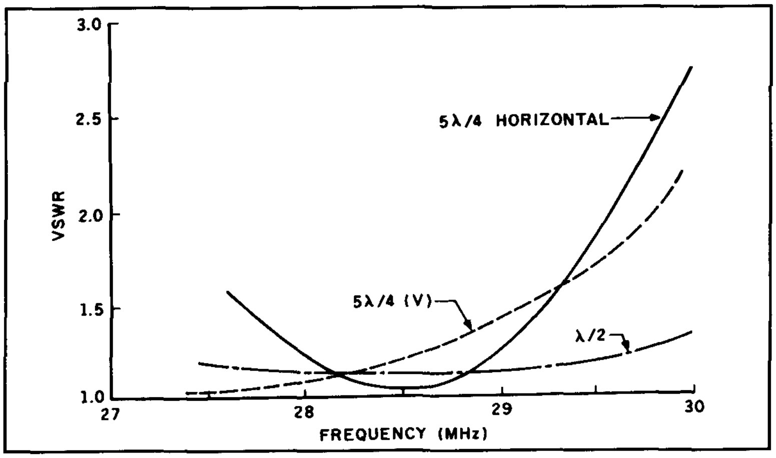


Figure 1. Plots of standing-wave ratio versus frequency for a 5/4-wavelength dipole in horizontal and V configurations, and for a horizontal 1/2-wavelength dipole. The calibration of the meter used is untrustworthy at high SWRs and the true absolute values may be larger.

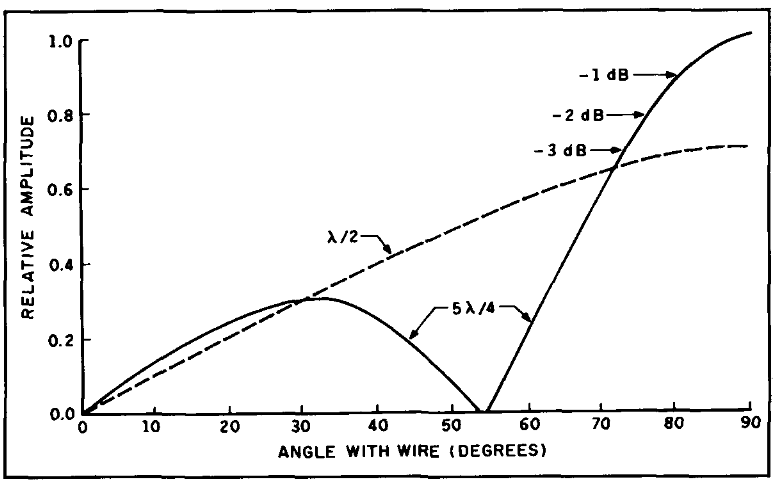


Figure 2. Plots of the radiation patterns of 5/4 and 1/2 wavelength dipoles. Because of symmetry in the patterns, only one quadrant is shown. The scales are adjusted to correspond to equal total radiated powers.

total beam width of the major lobe is about 18 degrees at 1-dB down, 27 degrees at 2-dB down, and 34 degrees at 3-db down. For approximately 37 degrees, it gives a greater signal strength than a half-wave dipole. At angles below 32 degrees, it's slightly better than a half-wave dipole.

Duplicating the antennas

If you're only interested in duplicating my antennas, you need little information other than that contained in **Table 1**. Consequently, you can skip most of the rest of the article. But expect to do some trimming of the number of turns. It's important that you read the section on adjusting the coupling. If you want to understand the significance of the power ratings, read the section giving construction details. If you wish to build an antenna for another situation, read the entire article.

General design procedure

You should be able to get a good match of the impedance of 50-ohm coax to an antenna with a radiation resistance of 210 ohms and a reactance of 1000 ohms using a 4:1 balun and two 500-ohm loading coils connected to either side. As this is a lot of hardware to suspend from an antenna, I wanted to build transformers to replace this combination.

Before I built the first transformer, I had to select a wire size and diameter. In the section "construction details," I explain how I made my choice. My first transformer was for 28 Mhz with a power level of 25 watts.

Measurements

A reactance of 1000 ohms at 28 MHz corresponds to that of a capacitance of 5 pF. I recognized that with the compression type insulators I use, there's shunt capacitance due to the wires on the insulators. With the limited measurement facilities I have available, I measured this to be roughly 1.5 pF. I wound a coil that I expected to resonate at 6.5 pF on one of the homemade insulators.

Next, I connected a 5-pF capacitor to the terminals of the coil. This was wound on a homemade compression insulator constructed from a piece of plastic rod. I took off a couple of turns to produce resonance at 28 MHz, as observed with a dip meter. Then, I connected a dummy load of two 100-ohm resistors in series with the 5-pF capacitor. At the same time, I wound a primary coil of a few turns on top of the secondary. I connected the primary to a standing-wave detector and a low powered signal source. I took off some turns until I got a low VSWR reading. I found that the frequency at which VSWR minimum occurred depended somewhat on the coupling. I worked to reduce the VSWR before setting the minimum at the frequency I wanted.

Finally, I connected the transformer to the antenna. The transformer needed just a small amount of trimming to give a VSWR minimum near the desired frequency. My original estimates of the resistance and reactance, although crude, turned out to be nearly correct!

Later, I built a second transformer for 28 MHz for a power level of 100 watts, and transformers for 50 MHz and 21 MHz (see details in **Table 1**). I measured the inductance and computed the reactance for each.

To measure the inductance, I determine the resonant frequencies using a dip meter when the secondary is connected to known capacitances selected to put the resonance between 4 and 10 MHz. I then compute the average values. These frequencies are low enough that the effect of the shunting capacitance is small.

I use a calibrated variable capacitor to make my measurements. However, you can use fixed capacitors and the values marked on them instead. Use several of them, preferably different models, and take the average value.

Reactance and capacitance values

Table 1 shows that the reactance of the inductors is larger at lower frequencies than at higher ones. This effect is the result of the shunting capacitance. By using a mathematical procedure known as "least squares,"[4] I found the best estimates for the reactance of the antenna, as corrected for the shunt capacitance, and the value of the capacitance. These are, respectively, 990 $\pm$170 ohms and 1.4 $\pm$0.9 pF. They are in excellent agreement with the theoretical estimate for the reactance and the direct measurement for the capacitance. The stated errors are for one standard deviation* and represent random errors.

Strictly speaking, these values pertain to antennas made of no. 12 wire 20 feet above ground. However, reactance varies very little with height and wire diameter. Thus, the variations you are likely to experience should be smaller than the stated errors. With these qualifications, you can use these values with confidence when designing transformers for other situations.

*A statistic used as a measure of dispersion in a distribution.

Effects of reactance

In **Table 1**, the average ratio of numbers of secondary turns to primary turns is about 5:1. If reactive effects were unimportant, you'd expect the ratio to be 2:1 with a load resistance of 210 ohms. While working with the dummy load, I found no indication of an impedance match with such a turns ratio. Therefore, reactive effects are significant. When they are present, there's an ambiguity. The impedance produced by 210 ohms in series with 1000 ohms reactance can also be produced by a resistance of 4972 ohms in parallel with a reactance of 1044 ohms. If you were to consider just the resistances, you'd expect a turns ratio of 10:1. The experimental value is between 2:1 and 10:1. This impedance-matching problem can't be solved simply by estimating a turns ratio from knowledge of the load resistance.

A brief review

To summarize, you should first select wire sizes and form diameters as outlined in section on construction details. Because it's easier to remove turns than add them, wind a secondary coil which gives a reactance slightly larger than 1000 ohms (when corrected for the shunting capacitance) using standard formulas, calculators, or trial and error. Then, wind a primary coil with about one-fifth the number of turns on the center of the secondary.

Install the transformer in the antenna and measure the SWR as a function of frequency. You should find some evidence of a minimum. Probably, it will be lower in frequency than you wish. Because the frequency of minimum VSWR depends a bit on the primary coupling, adjust the coupling to give a low VSWR according to the method discussed in the following section. Finally, adjust the number of secondary turns to bring the frequency of minimum VSWR to the desired value.

Adjusting the coupling

Looking at **Table 1**, you'll note that the primary coils consist of just two or three turns. When the coils are this small, you may have some problems adjusting the coupling. On your first attempt, two turns may seem to be too small and three turns too large.

There are some tricks you should keep in mind. Remember that the fringing field on the outside of the coil runs in the opposite direction to that inside the coil. This means that if you arrange the leads to the primary to include an appreciable area outside the cross section of the secondary, you can decrease the coupling.

In one case, I had to use larger spacers between the coil form and the connector to increase the exterior area. You can also tighten the coupling by twisting the leads to the primary together. In principle, you could tighten the coupling by having a loop of wire in the fringing field wound in the direction opposite to that of the main primary coil, but I've never tried this trick. With the 21-MHz transformer, I loosened the coupling by winding the outer turns of the primary close to the secondary end turns.

Construction details

If you're designing your own transformer rather than trying to duplicate one of those listed in **Table 1**, you must first choose the frequency and power level at which you're going to operate. These decisions lead indirectly to a determination of the transformer diameter.

Core size

If the physical size is small enough, you can combine the transformer with the center insulator (see the two smaller transformers in **Photo A**). In other cases, as suggested by the largest transformer in the photograph, you'll wind the transformer on a piece of tubing of suitable diameter, to be supported from the antenna wires by a group of radial wires with the center insulator inside.

Choosing wire sizes

I selected the wire sizes by first computing the current using the standard relationship – power is equal to the resistance times the square of the current – with a resistance of 50 ohms in the primary. Because the resistance in the secondary is about 200 ohms, its current is half as large. Then, I chose wire sizes using the copper wire table in *The ARRL Handbook*. I considered three power levels:

- **25 watts.** Appropriate for the Uniden HR 2510 and Radio Shack HTX-100. (Minimum wire sizes: no. 22 for the primary, no. 24 for the secondary.)
- **100 watts.** Appropriate for most common transceivers operating "barefoot." (No. 20 wire for the primary, no. 22 for the secondary.)
- **1500 watts.** Appropriate for linear amplifiers. (No. 14 wire for the primary and no. 16 wire for the secondary.)

These choices are very conservative. I assume key-down CW conditions. The table is also conservative. It implies closed multilayer windings, while these transformers are in open air and subject to excellent ventilation. Actually, I have used 25-watt transformers at 100 watts for short times without ill effects. It's probable that 25-watt transformers could be used safely with SSB at the 100-watt level.

The wire size you choose determines the number of turns per inch. At this point, it's necessary to proceed by trial and error. Assume a diameter and, either by experimentation or by using standard formulas or calculators, find the number of turns to obtain the desired inductance. If the size is reasonable, accept this diameter. If not, assume another diameter, and repeat the process until you get a reasonable set of dimensions.

Combining transformer and center insulator

There are practical size limitations for combining the transformer with the center insulator. You'll need to drill four 1/8-inch holes more or less parallel to the axis of the plastic core for the antenna wire loops. The length of common drill bits prevents this dimension from exceeding about 1.5 inches. The diameter must be 3/4 inch or larger. Because I don't have a drill press, I spoiled some samples because I couldn't guide the drill with sufficient accuracy. On the other hand, it may be difficult to obtain rods much larger than 1 inch in diameter. Even if you do, the transformer may become heavier than the combined weight of some tubing and a separate center insulator.

Concluding remarks

The 5/4-wavelength dipole, giving a gain of about 3 dB over a half-wave dipole, is an antenna that's simple to construct and erect. I've shown how to build compact transformers to match its impedance to 50-ohm coaxial cable. Such an antenna is useful for those who want antenna gain but can't use a multi-element beam. It should find favor with those who have lightweight portable equipment. This antenna should also appeal to those few Amateurs who own large tracts of rural land, and wish to erect effective 1.8 and 3.5-MHz antennas. ■

References

1. John J. Schultz, W4FA/J40DX, "The DEZI Dipole for 10 Meters," *CQ Magazine*, July 1989, page 18.
2. Yardley Beers, W0JF, "Designing Trap Antennas: A New Approach," *HAM RADIO*, August 1987, page 60.
3. *The ARRL Antenna Book*, edited by Gerald Hall, K1TD, ARRL, Newington, Connecticut, 1988, page 8-4.
4. Yardley Beers, *Introduction to the Theory of Error*, Addison-Wesley Publishing Company, Reading, Massachusetts, 2nd edition, 1956, page 38.

By Bryan Bergeron, *NU1N*
30 Gardner Road, Apt. 1G
Brookline, Massachusetts 02146

DIGITAL SIGNAL PROCESSING

Working in the frequency domain

Though most of us are comfortable working in the time domain, when it comes to digital signal processing (DSP), much of the work is done in the frequency domain. There are certain motivations for, and constraints associated with, performing digital signal processing in the frequency domain. I'll discuss these motivations and constraints, and review the Fourier Transform and its descendants.

Frequency domain processing

Like the oscilloscope and the spectrum analyzer, digital time domain and frequency domain signal processing take two very different perspectives of the same phenomena (see **Figure 1**). Time-domain waveforms viewed by an oscilloscope and frequency-domain spectra viewed through a spectrum analyzer are transforms of each other. One view doesn't carry any *more* data about a signal than the other. Rather, each provides a different way to think about and work with the same signal.

Frequency domain processing of a signal has two distinct advantages over processing in the time domain: tractability and efficiency. Often a problem that's virtually unsolvable in the time domain can be deciphered easily in the frequency domain. Consider the challenge of equipping a RADAR system with real-time pattern recognition capabilities. While this is a sizable engineering feat in the time-domain, it becomes a trivial problem in the frequency domain. In the frequency domain, a signal from an aircraft or other object of interest can be looked at independent of the size, orientation, or position of the object[1]

For cases that can be solved in either domain, working in the frequency domain is almost always simpler. For example, you can easily determine the frequency response of a microphone by sampling its output (using a white noise source) every few milliseconds and

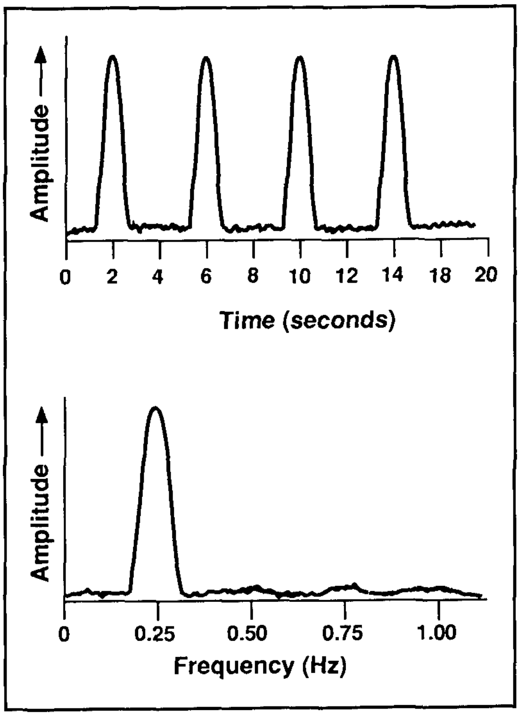


Figure 1. How the Fourier Transform relates time to frequency. In the time domain (top), a signal is composed of short bursts regularly spaced at one per every four seconds (0.25 Hz). When transferred via the Fourier Transform from the time domain to the frequency domain (bottom), the same signal appears as a response at 0.25 Hz.

then manipulating the signal in the frequency domain. While you could obtain the same results working with a digital filter defined in the time domain, frequency domain work is more straightforward and conceptually clean. It is also more efficient computationally.

Here's an illustration of the potential advantages of working in the frequency domain. Think about the DSP challenges associated with constructing adaptive filters[2]* and in filtering unwanted noise

*Adaptive filters are filters that cancel or minimize noise and interference by dynamically updating the filter coefficients to adapt to the characteristics of the interference.

from a signal. Adaptive filters are commonly implemented using digital technology because of the inherent stability and mathematical tractability of the algorithms used for the computation of the filter coefficients. Although the algorithms are straightforward when implemented in the time domain, the performance of these algorithms is typically less than that of equivalent frequency domain algorithms.[3] The more computationally efficient frequency domain implementations of adaptive filters are often called frequency domain adaptive filters, or block adaptive filters.[4]

Frequency domain work lends itself to the identification and elimination of noise and other artifacts, especially when these undesirable signals are significantly higher or lower in frequency than the desired signal. A 60-Hz noise in a communications signal (bandwidth 300 to 3000 Hz) or 2-kHz instrumentation noise superimposed on a 200-Hz signal are examples of undesirable signals. Think of the procedure for removing high or low frequency noise from a signal as a simple multiplication process in the frequency domain.[5] For example, if the 60-Hz noise is distinct and separate from a 300 to 3000-Hz signal spectrum, then a simple rectangle function can zero all data values below 300 Hz. That is, data values between 300 and 3000 Hz are multiplied by 1; all other data values are multiplied by zero (see **Figure 2**).

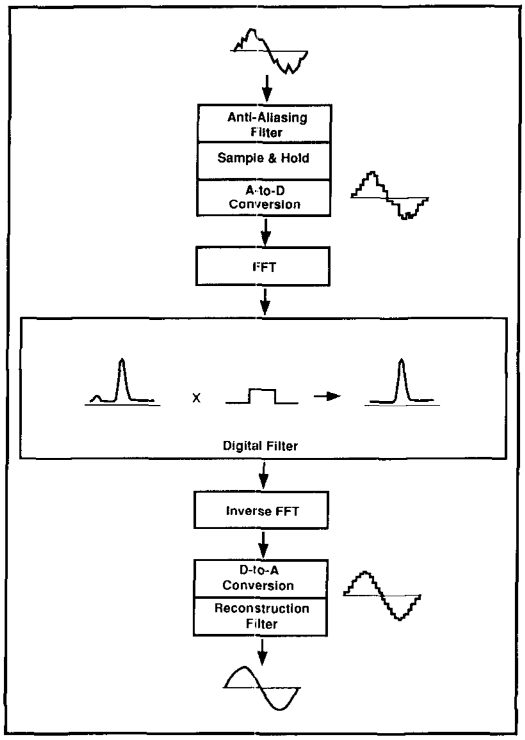


Figure 2. An example of Digital Signal Processing in the frequency domain.

Look at the top of **Figure 2**. It shows that the original noisy analog signal is first preprocessed by a high pass (anti-aliasing) filter. Next the digitizing hardware samples the analog signal and maps the values onto digital values at regular time intervals. A Fast Fourier Transform, (FFT – more on this later), or its equivalent, is then used to move the signal into the frequency domain. The digital filter shown in the middle of **Figure 2** (in this case, a simple rectangular function) removes the unwanted signal (represented by the small peak to the left of the main signal peak). Once the filter function has been applied, the signal is converted back to the time domain by performing an inverse FFT on the data. After digital to analog (D/A) conversion and filtering, the desired signal, free of noise (shown at the bottom of the figure), is available for futher processing or direct use.

If the signal data and the noise aren't clearly distinguishable (as is normally the case), you must select a judicious cutoff, based on the known characteristics of the unwanted noise and the desired signal. Also, the digital filter function you choose must gradually attenuate the unwanted spectra in the frequency domain; the spectra can't suddenly drop off to zero. Such a sudden cutoff would result in false accentuation of frequencies corresponding to the cutoff point in the frequency domain. This idea is developed in the discussion of windowing found in the section on frequency domain DSP considerations.

The Fourier Transform

You can't work effectively with DSP in the frequency domain without a good conceptual understanding of the Fourier Transform (named after the 18th century mathematician Jean Baptiste Joseph Fourier). By relating time to frequency, this transform is the basis of frequency domain processing as it is known today. For the mathematically inclined, the Fourier Transform is defined by the following equations:[6]

$$x(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} X(j\omega)e^{j\omega t}dt \qquad (1)$$

$$X(j\omega) = \int_{-\infty}^{\infty} x(t)e^{-j\omega t}dt \qquad (2)$$

where $j = \sqrt{-1}$ and $\omega = 2\pi f$. Notice that the transformation from the frequency domain to the time domain (**Equation 1**), and from the time domain to the frequency domain (**Equation 2**), integrate over the limits from $-\infty$ to $+\infty$. This results in the mathematically correct concept of negative frequency, which has no basis in physical reality.

In practice, the complex exponential, $e \pm j\omega t$, is usually replaced with the trigonometric expression:

$$e^{-j\omega t} = \cos \omega t \pm j \sin \omega t$$

Although mathematically elegant and conceptually beneficial, the basic Fourier Transform is of limited practical value when working with digital computers because it assumes that the data to be transformed are continuous. Actually, analog data are sampled and digitized into a machine-readable form at discrete intervals. To allow high speed computers to handle frequency-time domain transformation calculations, the Fourier Transform, which uses the infinitesimal dt, was modified into the Discrete Fourier Transform (DFT) which expects either a quantized continuous signal or a signal of limited duration. Mathematically, the DFT appears as:[6]

$$x_n = \sum_{k=0}^{N-1} X_k \exp\left[j \frac{2\pi nk}{N}\right] \qquad (3)$$

$$X_n = \frac{1}{N}\sum_{k=0}^{N-1} x_k \exp\left[-j \frac{2\pi nk}{N}\right] \qquad (4)$$

Notice that the integral has been replaced with summation over N discrete data points. The frequency to time transform in **Equation 3** also differs from the time to frequency transform in **Equation 4** as the result of a change in the sign of imaginary j and the scaling multiplier 1/N.

Creating a workable computer program to perform these calculations is relatively straightforward. However, because a merely workable program is seldom good enough for real applications, researchers have spent many years increasing the computational efficiency and memory requirements of DFT algorithms. There is now a wide variety of rapid and efficient methods for computing the DFT. The original Fast Fourier Transform (FFT) and the more recent Fast Hartley Transform (FHT) are two popular ones.

The Fast Fourier Transform

The Fast Fourier Transform may be the best known method for computing the DFT rapidly. The FFT takes advantage of the redundant calculations within the DFT — a primary reason for the increased speed of the FFT over the basic DFT. As its basic premise, the FFT and related algorithms sort data using a data-pairing permutation process (sometimes referred to as the "butterfly" because of the appearance of the associated data flow diagram) until data is separated into pairs. The Fourier transform calculation on these data pairs is rapid. It is computationally more expensive to compute a 32-point DFT than it is to compute 16 two-point DFTs.

A large part of the data-pairing permutation in the FFT algorithm is concerned with a bit reversal procedure which scrambles the order of the output data creating a mirror image of the input. The speed of this bit reversal (or reshuffling of data) defines, to a great degree, the efficiency of a given FFT algorithm. (In cases where execution speed is critical, this bit reversal can be accomplished in ASSEMBLER.) To increase the speed of the actual Fourier transform, you can substitute a trigonometric look-up table for the calculation of trigonometric functions supplied by the host language.

The FFT algorithm has a major restriction. For the data-pairing permutation to function properly, the number of discrete data input values (N) must be an integral power of 2 (2, 4, 8, 16, 32, 64, 128, and so on). Because the FFT expects 2^n data points, zero filling or padding is commonly used to increase the number of data points to the next higher power of 2; that is, from 45 to 64 or from 120 to 128 data points. Unfortunately zero filling can result in phantom responses. These phantoms can be reduced by windowing the data. This means you variably attenuate the first and last few input data values to reduce sharpness of the drop to the zero-padded area. Another problem associated with zero filling is related to the increased storage and computational requirements imposed by the added data, which add nothing to the information content of the signal.

It may seem that you have to go to a lot of trouble to achieve an increase in execution speed, but considerable time can be saved by substituting the FFT for the DFT. For example, the time required to compute a DFT is proportional to N^2, while the time required to compute the FFT is proportional to $N \times \mathrm{Log}_2 N$ — thanks to the data-pairing permutation. For large data sets, this can amount to a significant savings in computer resources (see **Table 1**).

Figure 3 provides a BASIC implementation of an FFT subroutine (compatible with BASICA for the IBM-PC). It is based, in part, on a listing by Brook and Wynne.[7] The subroutine assumes that the arrays for holding real data (AR) and imaginary data (AI) have been defined with dimension N. Changing the sign of ID from plus to minus allows the inverse function to be performed on the data array. For example, ID = +1

Samples (N)	DFT	FFT
8	64	24
16	256	64
32	1024	160
64	4096	384
128	16384	896
256	65536	2048
512	262144	4608
1024	1048576	10240

Table 1. A comparison of the number of computations involved in calculating the DFT and the FFT of a signal. Notice that as the sample size (N) increases, FFT superiority becomes more significant. That is, with a 1-K sample size (1024 elements), the DFT/FFT ratio is approximately 100:1.

for time to frequency transformation and ID = −1 for frequency to time transformation. Lines 100 to 130 of the subroutine perform the scaling function when the transformation is from the time to the frequency domain. Lines 140 to 230 perform the data pairing and lines 250 to 420 are responsible for the actual Fourier transform. As noted previously, the data-pairing permutation can be coded in ASSEMBLER to maximize computational efficiency. As you can see, this algorithm produces the same number of data output points as data input points.

```
100 IF ID > 0 THEN FOR J = 1 TO N
110      AR (J) = AR (J) / N
120      AI (J) = AI (J) / N
130 NEXT
140 NHLF = N/2 : NM1 = N-1 : J = 1
150 FOR L = 1 TO NM1
160      IF (L >= J) THEN 190
170      T = AR (J) : AR (J) = AR (L) : AR (L) = T
180      TX = AI (J) : AI (J) = AI (L) : AI (L) = TX
190      K = NHLF
200      IF (K >=J) THEN 230
210      J = J - K : K = K / 2
220      GOTO 200
230      J = J + K
240 NEXT  L
250 FOR M1 = 1 TO M
260      UR = 1.0 : UI = 0 0
270      ME = 2^M1 : K = ME / 2
280      CON = PI / K
290      FOR J = 1 TO K
300           FOR L = J TO N STEP ME
310                LPK = L + K
320                TR = AR (LPK) * UR - AI (LPK) * UI
330                TI = AR (LPK) * UI + AI (LPK) * UR
340                AR (LPK) = AR (L) - TR
350                AI (LPK) = AI (L) - TI
360                AR (L) = AR (L) + TR
370                AI (L) = AI (L) + TI
380           NEXT L
390           UR = COS (CON * J)
400           UI = - SIN (CON * J) * ID
410      NEXT J
420 NEXT M1
```

Figure 3. A simple FFT subroutine in BASIC, compatible with BASICA for the IBM-PC and clones. This code can be easily extended to include plotting functions to produce graphs like those shown in *Figures 4-7*. For time critical applications, lookup tables can be substituted for the SIN and COS functions supplied by your BASIC interpreter.

Notice also how the computationally expensive evaluation of the complex exponential has been replaced with an equivalent trigonometric expression (lines 390 to 400). Replacing the COS and SIN evaluations with a look-up table is an easy way to improve the performance of this subroutine significantly, without resorting to working in ASSEMBLER. The tradeoffs associated with the faster speeds provided by a look-up table include the cost of RAM required to hold the look-up table elements and decreased accuracy of the transform function (because the values returned for each SIN and COS evaluation are limited by the bit length and angle resolution of the table).

Because you can't fully appreciate the operation of the FFT algorithm without illustration, I've included graphs of the actual input and output data from an FFT program similar to the one in **Figure 3**, implemented in MacForth on the Apple Macintosh (see **Figures 4** through **7**). Notice that, for each figure, there are three values plotted for the signal in the frequency domain. The real and imaginary components correspond to the values in the AR and AI data arrays used **Figure 3**. The magnitude component, which corresponds to the familiar power spectra of the signal, is proportional to the absolute value of the square of the real and imaginary components.

The Fast Hartley Transform

The FFT has been a dependable workhorse for Frequency Domain Digital Signal Processing (FDDSP) since the mid-sixties. However, the demands of modern DSP applications and the move from mainframe systems to microcomputers have spurred the development of more efficient algorithms. One of the more notable descendants of the FFT is the Fast Hartley Transform (FHT), based on the continuous transform introduced by R.V.L. Hartley in 1942.[8] Like the FFT, the FHT maps a signal from the time domain into the frequency domain (and vice versa). However, where the FFT maps a real function of time into a complex function of frequency, the FHT maps a real function of time into a real function of frequency. Mathematically, the FHT appears as:

$$X(t) = \sum_{t=0}^{N-1} H(f)\, cas\left[\frac{2\pi ft}{N}\right] \qquad (5)$$

$$H(f) = \frac{1}{N} \sum_{t=0}^{N-1} F(t) cas\left[\frac{2\pi ft}{N}\right] \qquad (6)$$

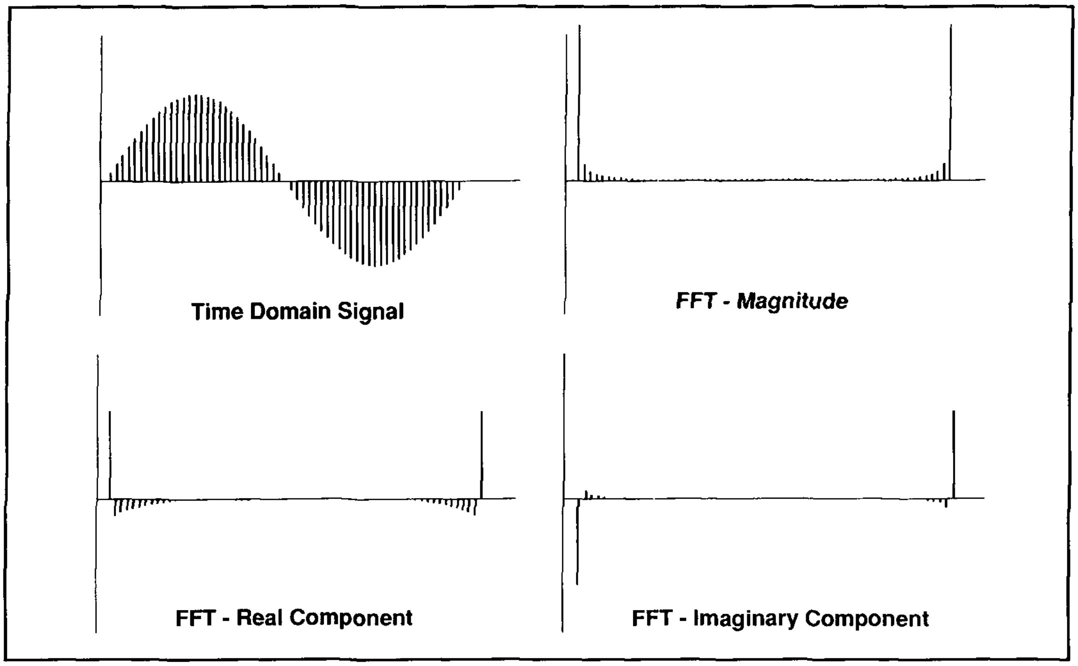


Figure 4. A time domain sinusoidal signal sample with an integer number of complete cycles, and nearly equal amplitude values at either end of the sample (top, left). The magnitude or power spectra (top, right), as well as the real (bottom, left) and imaginary (bottom, right) components of the FFT are also illustrated. Note the relative purity of the magnitude plot. Most of the signal energy is concentrated in a few spectral lines. Note also that in this figure, as well as the following three, the FFT plots contain both positive (left side of each frequency domain plot) and negative (right side of each frequency domain plot) frequency components. For practical purposes, you can ignore the right half of each plot.

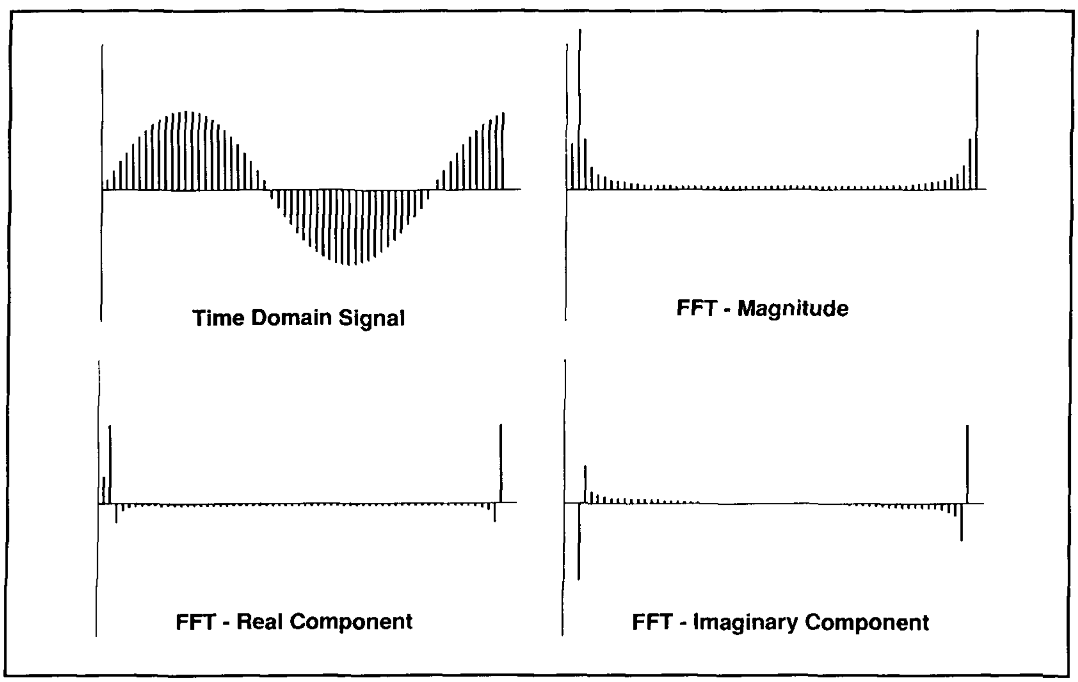


Figure 5. A sinusoidal signal sample in the time domain in which the sampling interval and frequency are such that irregular points in the waveform have been captured (top, left). Even though this signal is of the same amplitude and purity as the sinusoidal signal in *Figure 4*, notice the relative impurity of the magnitude plot; *i.e.*, there is now a considerable amount of energy distributed throughout the frequency domain plot. The solution is to either employ a windowing function, or to adjust the sampling interval so an integer number of complete cycles are captured.

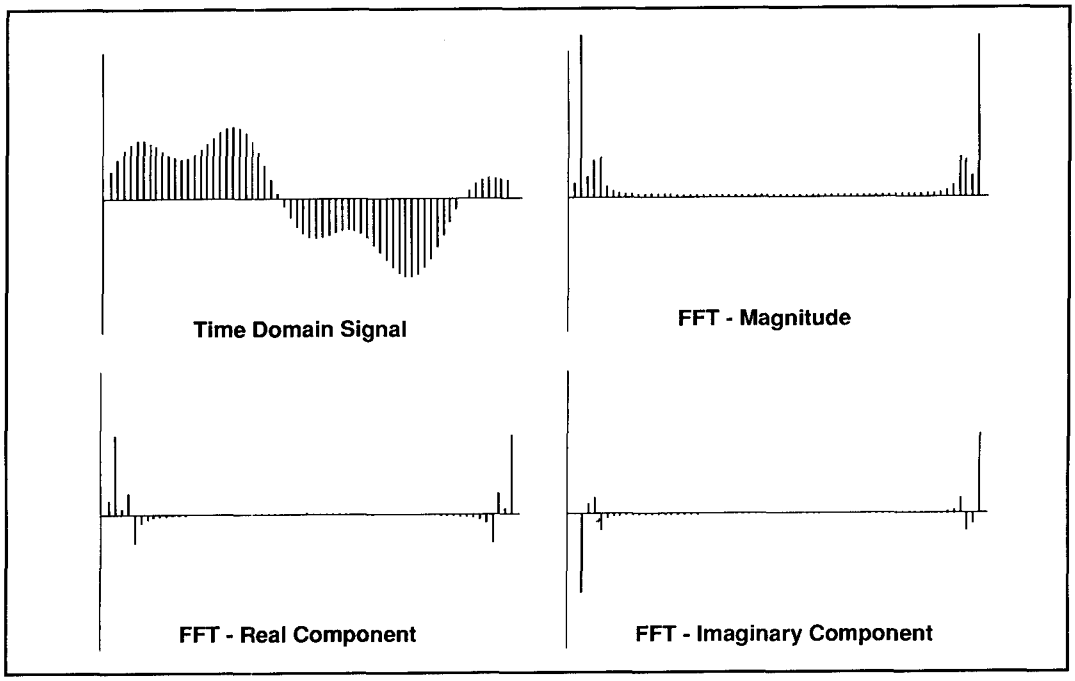


Figure 6. In this example, there are two sinusoidal signals, one double the frequency and one quarter of the amplitude of the other (top, left). Notice the extra responses in the frequency domain in the magnitude (top, right), real (bottom, left), and imaginary (bottom, right) components.

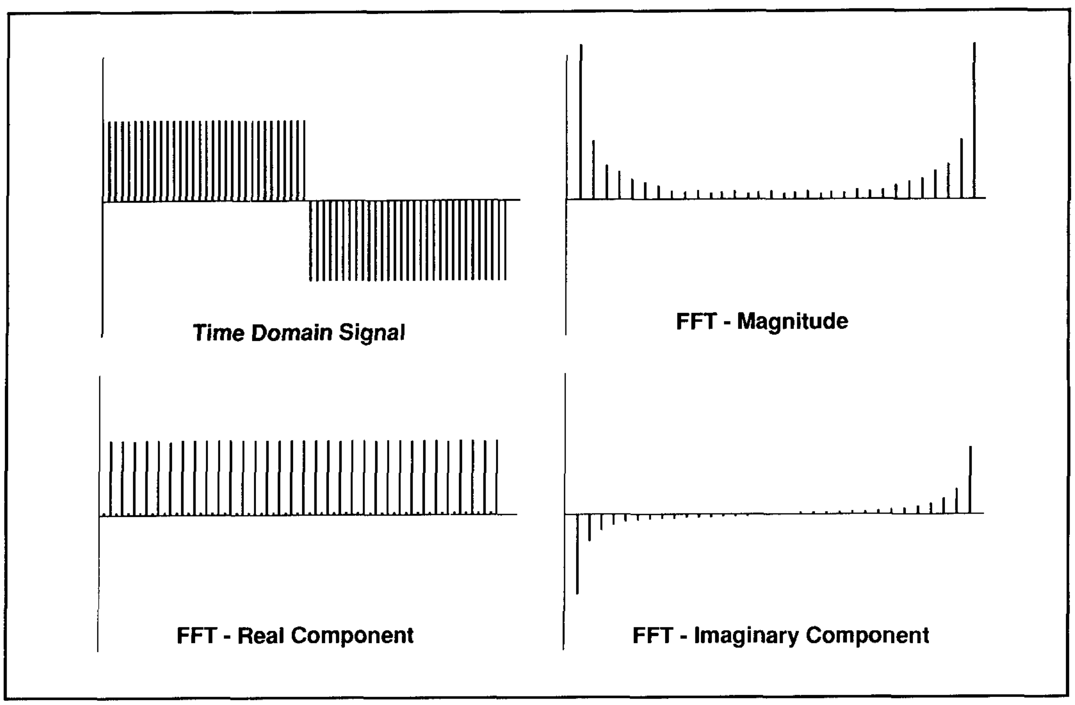


Figure 7. The time domain signal in this example is one complete cycle of a simple square-wave signal (top, left). When compared with a similar sinusoidal signal in *Figure 4*, you'll note there is a relatively large amount of spectral energy distributed above the main signal peak (top, right). This is to be expected, because square waves are composed of a large number of harmonically related sine waves.

where **cas**, in both the frequency to time **(Equation 5)** and time to frequency transform **(Equation 6)**, is equivalent to the cosine and sine of the expression in the brackets. That is, cas (β) = cosine (β) + sine (β). Notice the similarity of **Equations 5** and **6** with those that describe the FFT. The main difference is the substitution of the real function cas(2πft) for the complex exponential term in the FFT.

Computationally, working with real numbers (instead of real and imaginary numbers) is very advantageous. For each arithmetic operation required to compute the FHT, six operations are required to compute the FFT. Four operations are necessary for each complex multiplication or division, and two operations are required for complex addition or subtraction.[8] Compared with the FFT, the FHT has associated memory savings in addition to computational savings. FFT calcuation requires the use of complex numbers. Because complex numbers are composed of two distinct parts (real and imaginary), they require twice as much computer storage space as real numbers.

Consequently, the FHT requires only about half as much working memory as the FFT, because complex data arrays require twice as much space as real data arrays.[9] The FHT is an especially attractive alternative to the FFT when you have a large volume of data to work with, as in digital image processing. The FHT, while more efficient than the FFT, has considerably more code associated with its implementation than the FFT. If you are interested in coding examples of the FHT, see the excellent text by Bracewell.[10]

Frequency domain DSP considerations

The FFT and its derivatives constitute the core software tools used for virtually all FDDSP applications. Like other software tools, they can easily be misused if the operator doesn't understand the underlying assumptions of their design. To reap the greatest benefit from any FDDSP system, you have to understand the characteristics of the signal to be processed and also be aware of the capabilities and limitations of the software and hardware components of your DSP system. Some of the more pertinent aspects of DSP in the frequency domain are outlined in more detail in the sections that follow.

Windowing

In most FDDSP systems, signal samples are collected into blocks of length 2^n and then processed by some type of FFT algorithm. In FDDSP it is important to assume that each successive discrete sample represents part of a continuous signal which repeats indefinitely what is in the sample; that is, the signal is periodic. If the sampling frequency and sampling interval are selected so that complete integer number of cycles are captured in each sample (as in **Figure 4**), then the sample boundaries will be of nearly equal amplitude, and the transition from one sample to the next will be smooth. If, on the other hand, the sampling frequency and interval are such that irregular points in the waveform cycle are captured, there will be high frequency artifacts in the transformed data (as in **Figure 5**). The effect will be most pronounced when the sample contains an exactly odd number of half cycles, because the discontinuity at sample boundaries will be maximum.[11]

It's obvious that one condition under which the FFT works best is when the data to be transformed smoothly approaches 0 at both ends of its range **(Figure 4)**. If, however, the actual data do not conform to the ideal, you can force them into an acceptable form by multiplying them by a *window function* before calculating the transform. A window function effectively multiplies data values near the center of the sample by unity, data near the ends of the sample by 0, and data between the center and ends by some intermediate value. The nature of these intermediate multiplicative factors defines the nature of the window.

The triangular window is a simple and fast window function, where the multiplication factor decreases linearly and symmetrically toward both ends of the sample. Another popular window function is the Hanning function (see **Figure 8**), which provides better artifact reduction, at the expense of computational efficiency. This function is defined as:

$$w\{i\} = 0.5(1-(\cos(2\pi/N))$$

where i = the sample point number and N = the total number of samples.[12] For more information on window functions and their uses, see the work by Press.[13]

Sampling jitter

It is a basic (but commonly overlooked) assumption of FFT work that the signal is sampled at *regular* time intervals – every 10 milliseconds, for example. Sampling jitter is the distortion of the sampled signal due to variations in the sampling interval. This jitter, which increases the noise floor of a signal, affects higher frequencies more than lower ones. Sampling jitter is most common in systems that rely on software triggering of the analog to digital (A/D) conversion process, rather than the more stable and reliable hardware triggering methods.[11] Software-based triggering systems that vary only a few microseconds between samples can add significantly to the noise floor of a system.

Quantization error

Like sampling jitter, quantization error raises the noise floor in a FDDSP system. This is commonly referred to as quantization noise. Like sampling jitter, quantization error is a function of how the data is acquired and processed, before the actual digital signal processing. Quantization error results when the actual, instantaneous value of a continuous signal is mapped onto the nearest integer value supported by the A/D conversion hardware. For instance, quantization error can occur when both 12.157 and 12.234-volt signals are mapped to 12.2 volts by an A/D converter. This error can be minimized by using an A/D converter with greater resolution. For example, you could use a 12-bit digitizer in place of an eight-bit unit. A compromise must always be made between quantization error (noise) and the increased cost, speed penalty, memory requirements, and computational load imposed by a higher resolution A/D converter.

Sampling frequency

Although extrapolation procedures have been developed for the FFT to accurately determine the frequency of signals higher than the Nyquist frequency,[14] it's generally accepted that the sampling frequency must be at least twice the frequency of the signal to be sampled. This means there should be at least two samples per cycle of the highest frequency contained in the signal. It's often

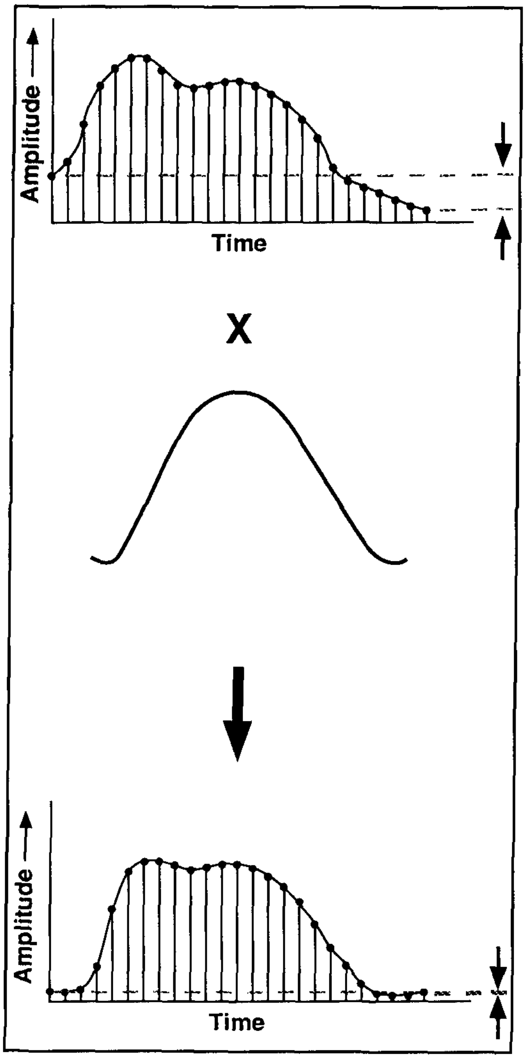


Figure 8. An example of how windowing, when applied to time domain signals prior to FFT processing, can help minimize artifacts due to discontinuities at the ends of the sample. In this example, the original signal (top) is preprocessed with the Hanning function (depicted graphically, center), to equalize signal amplitudes at both ends of the sample (bottom). *Notice that one side effect of windowing is that* data points at both ends of the sample are thrown away. The consequences of this data loss are described in the text.

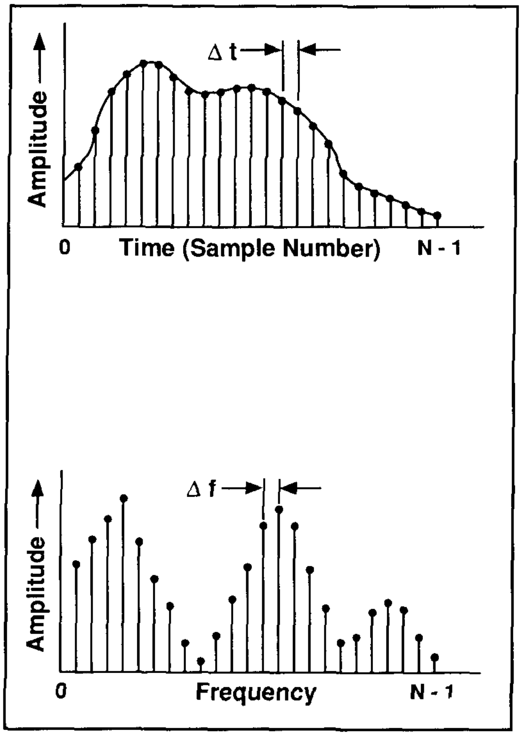


Figure 9. *Spectral resolution versus sampling time for the* Discrete Fourier Transform (DFT). The spectral resolution (Δf) is equal to $(N \times \Delta t)^{-1}$, where N is the number of samples taken of the signal in the time domain, and Δt equal to the sampling time in seconds. For example, if the sampling time in the time domain (top) is 10 ms, and the number of samples is 100, then the spectral resolution (bottom) will be $(100 \times 0.010)^{-1}$, or 1 Hz.

necessary to use a low-pass filter front end to any A/D converter to assure that only frequencies which can be handled adequately are passed on to the converter. Otherwise, aliasing (the folding down of undersampled signals) can result.

Resolution

The spectral resolution of an FDDSP system is closely related to the sampling frequency, sampling jitter, and windowing. In general, the frequency resolution is about equal to the reciprocal of the sampling interval in the time domain (see **Figure 9**). The smaller the sampling interval, the higher the frequency resolution. Sampling jitter effectively decreases the resolution of a system, because the certainty of the sampling interval, and therefore the frequency interval, is diminished. The noise associated with sampling jitter also diminishes the effective resolution of system, especially higher frequency signals.

Windowing also decreases the effective spectral resolution of a system. While reducing the number of possible artifacts, windowing throws out or gives less weight to the sampled data at both ends. As **Figure 9** illustrates, the FFT and its descendants produce one output data point for each input data point. Throwing out data in the time domain, in effect, spreads the frequency domain signal by a proportional amount. For example, by using windowing to decrease the effective number of input data points by 10 percent, you decrease the frequency resolution by approximately 10 percent.

Aperture time

Aperture time, like quantization error, is largely a function of the A/D hardware used in signal acquisition. The aperture time of an A/D converter — the time during which an analog signal is actually sampled before being digitized — can be likened to the shutter speed of a camera. When the camera shutter is open, light falls on the film emulsion exposing silver halide crystals to light energy. For a given shutter speed, slowly moving (low frequency) objects may be exposed clearly and accurately, while very fast objects (high frequency) might appear as blurs on the developed film. In photography, the solution is to use a faster shutter speed — assuming the film has enough latitude to work at the higher speed. It may be necessary to use a faster film with less resolution to capture a clear image of the faster objects.

The photograph analogy is useful if you think of the A/D resolution as the film resolution, the A/D conversion time as the film speed, and the aperture time as the shutter speed. High frequency signals can be digitized accurately only with a relatively short aperture time. But to use the short aperture time, the sample-and-hold circuit within the A/D conversion hardware must be capable of acquiring the signal in a relatively brief period of time. Of course, the A/D conversion hardware must also be capable of digitizing the signal before the next sample time. A high resolution A/D converter, like a 32-bit system (the photographic equivalent of low speed, high resolution film), will generally support a lower maximum sampling frequency than a low resolution converter, like an eight-bit system (high speed, low resolution film), assuming the converters are in the same price/performance range.

For many DSP applications, working in frequency domain is not only more efficient than working in the time domain, but the only means of arriving at a solution to a particular problem.

Summary

For many DSP applications, working in the frequency domain is not only more efficient than working in the time domain, but the only means of arriving at a solution to a particular problem. The Fourier Transform and its more computer compatible descendants, including the FFT and FHT, serve as the basis for the vast majority of operations in the frequency domain. While powerful algorithms for the FFT and FHT are easy to implement on desktop computer platforms, these frequency domain tools must be used with caution. Intelligent use requires knowledge of the signal to be processed and the features and limitations of the available DSP hardware and software.

With the rapid evolution and introduction of inexpensive DSP software environments, knowledge of the inner workings of both time and frequency domain signal processing is becoming a necessity. There are electronic circuit design prototyping techniques available today that place powerful DSP tools, like the FFT, in the hands of anyone with a personal computer (see **Figure 10**). To use these tools effectively, you must understand the assumptions and tradeoffs made by the software system designer. For instance, if you use a look-up table to increase the computational efficiency of the supplied FFT algorithm, what effect does this have on the accuracy of the trans-

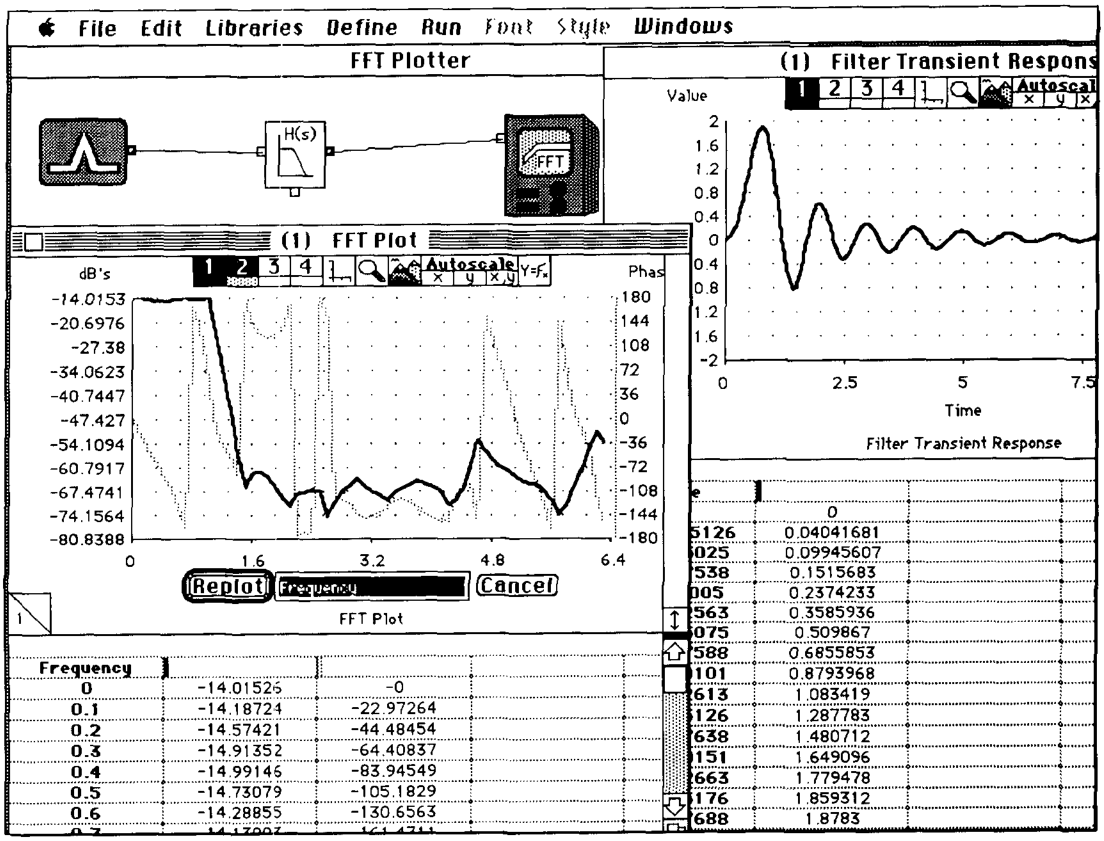


Figure 10. An example of the many, increasingly popular microcomputer-based DSP prototyping environments available to engineers. In this Apple Macintosh program (Extend™, from Imagine That!), you see a simple impulse function, followed by a low-pass filter and an FFT plotter (top, left panel). The icons represent code modules, and the lines connecting them represent data-flow paths. The time domain plot appears in the right hand panel, partially obscured by the FFT plot to the left and center of the display. Tools like this let engineers to design and debug complex DSP systems in hours instead of weeks.

formed waveform? A knowledgeable user is a powerful user.

In the next part of this series, I'll discuss artificial intelligence (AI) techniques that have been applied to digital signal processing in both the time and frequency domains. ■

References

1. D. Brown and S. Silverman, "Flexible High Speed Image Processing with Non-von Neumann Design," *Computer Technology Review*, 1986, 8(16): pages 91-94.
2. S. Haykin, *Introduction to Adaptive Filters*, Macmillan, New York, 1984.
3. S.D. Stearns, "Fundamentals of Adaptive Signal Processing," *Advanced Topics in Signal Processing*, Lim and Oppenheim, editors, Prentice Hall, Englewood Cliffs, New Jersey, 1988.
4. M.J. Dentino, J. McCool, and B. Widrow, "Adaptive Filtering in the Frequency Domain," *Procedure IEEE*, 1978, 66(Dec): pages 1658-1659.
5. E.E. Aubanel and K.B. Oldham, "Fourier Smoothing Without the Fast Fourier Transform," *Byte*, February 1985, pages 207-218.
6. D. Brook and R.J. Wynne, "Frequency Characterization," *Signal Processing: Principles and Applications*, Edward Arnold, London, 1988.
7. D. Brook and R.J. Wynne, *Signal Processing: Principles and Applications*, Edward Arnold, London, 1988.
8. M.A. O'Neil, "Faster Than Fast Fourier," *Byte*, April 1988, pages 293-300.
9. R.N. Bracewell, *The Fourier Transform and Its Applications*, McGraw-Hill Book Company, New York, 1986.
10. R.N. Bracewell, *The Hartley Transform*, Oxford University Press, Oxford, 1986.
11. A. Sewards, "Introduction to DSP," *Electronics and Wireless World*, 1988, 94(1630): pages 741-746.
12. H.J. Hutchings, "Interfacing and Signal Processing with C," *Electronics and Wireless World*, 1988, 94(1631): pages 948-956.
13. W.H. Press, B.P. Flannery, S.A. Teukolsky, and W.T. Vettering, "Fourier Transform Spectral Methods," *Numerical Recipes in C*, Press Syndicate of the University of Cambridge, Cambridge, 1988.
14. S.E. Georgeoura, "Fast Fourier Transforms of Sampled Waveforms," *Electronics and Wireless World*, 1988, 94(1633): pages 1122-1126.

Bibliography

1. P. Sewla, "Building a Digital Filter: FIR Filter Features Guarantee Phase Linearity," *Ham Radio*, April 1989, pages 9-17.
2. B.P. Bergeron, NU1N, "Digital Signal Processing: Part 1—The Fundamentals," *Ham Radio*, April 1990.

By Allen R. Wooten, *WD4EUI*
106 Belair Road, S.E.
Huntsville, Alabama 35802

INTERFACE YOUR COMPUTER TO THE "POOR MAN'S SPECTRUM ANALYZER"

Add features found in expensive commercial units to your "Poor Man's Spectrum Analyzer"

Since the publication of the "Low-Cost Spectrum Analyzer with Kilobuck Features" by Robert Richardson, W4UCH, in the September 1986 issue of *HAM RADIO*,[1] there have been several follow-up articles about modifying and improving this interesting project.13[2,3,4] I purchased the tuners and receiver board from WA2PZO* and built my own version. I combined the VHF "cable ready" tuner and the VHF/UHF tuner (the VHF section isn't used) in a single enclosure and now have a spectrum analyzer that covers from 1 to 900 MHz in two ranges.

Background

I added a sawtooth generator to produce the frequency sweep described by Joe Carr in the September 1987 issue of *HAM RADIO*.[3] This provided a good oscilloscope display of the frequency spectrum. A turns-counting dial, attached to the multi-turn frequency-control potentiometer, let me set the center frequency accurately after calibrating the dial. The more expensive spectrum analyzers offer additional features like calibrated display (amplitude and frequency), programmable start and stop frequencies, and the capability to plot the spectrum display.

I wanted to add these features to my unit, so I interfaced the "Poor Man's Spectrum Analyzer" to my IBM AT compatible computer and wrote a control program in *Microsoft*® Quick Basic.* The interface requires a digitizer board that's mounted inside the computer, a frequency control board mounted with the spectrum analyzer, and an interconnecting cable.

The computer controls user-selected start and stop frequencies through the parallel printer port. After the amplitude versus frequency data has been acquired, it's plotted on the screen. The "print screen" command then plots the data on paper via the printer. **Figure 1** shows my spectrum analyzer and the computer interface.

The program lets you select the start frequency in two bands (1 to 500 MHz and 393 to 900 MHz) for use with the two tuners available from WA2PZO. You choose the start frequency, and the program establishes it using a 12-bit D/A converter (DAC). The frequency sweep uses an 8-bit DAC to obtain high resolution over a limited frequency range above the start frequency. The maximum permissible stop frequency is calculated based on the start frequency, and then displayed. After you select the desired stop frequency, the program makes one sweep and displays the plot on the monitor screen.

* WA2PZO, Science Workshop, Box 393, Bethpage, New York 11714.

* IBM AT is a registered trademark of IBM Corporation. Quick BASIC and GW-BASIC are registered trademarks of Microsoft Corporation.

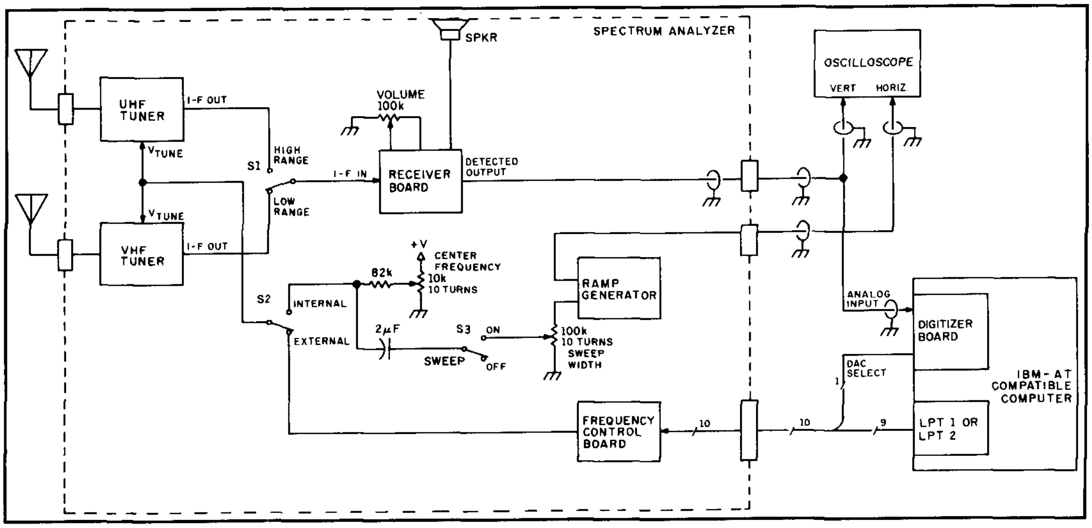


Figure 1. Spectrum analyzer.

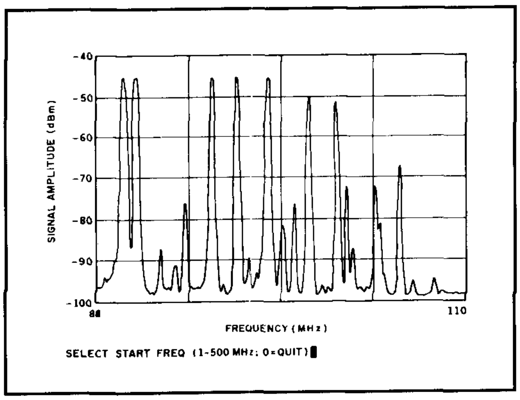


Figure 2. Spectrum plot using the VHF tuner.

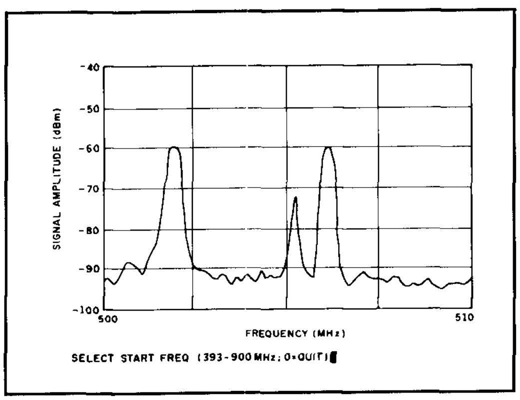


Figure 3. Spectrum plot using the UHF tuner.

Figure 2 shows a typical plot from the *spectrum analyzer program using the* VHF tuner. This sample shows the FM broadcast band with a part of the aircraft band as received in the Huntsville, Alabama area. **Figure 3** is a typical plot using the UHF tuner. In this plot, you can clearly see the video, color, and audio carriers of television channel 19.

Digitizer board

The schematic for the digitizer board is shown in **Figure 4**. Integrated circuit U3 is an 8-bit A/D converter that requires no special logic for interfacing with the computer data bus. The output latches of the A/D converter have high-impedance outputs that connect directly to the computer's data bus. The A/D converter only outputs data when addressed by the computer. Potentiometer R1 sets the reference voltage for the converter and thus determines the converter's full-scale setting. U1 and U2 are address decoders wired so an address of 22F hexadecimal (559 decimal) will start the conversion.

DAC operation

While in operation, the computer treats the digitizer board as just another memory location. The board only digitizes the analog input when memory location 22F is addressed. U4 is used as a switch to select the appropriate DAC on the frequency control board when the computer sends frequency control data. When the computer addresses memory location 22B hexadecimal (555 decimal), the DAC-select line is a logic high. When the computer addresses memory location 22D hexadecimal (557 decimal), the DAC-select line is a logic low. More about the DAC-select line later.

Digitizer construction

The digitizer board is mounted inside the computer in one of the expansion slots. Each expansion slot has two receptacles for the edge connectors on the expansion boards. The digitizer board requires just one edge connector and uses only the receptacle

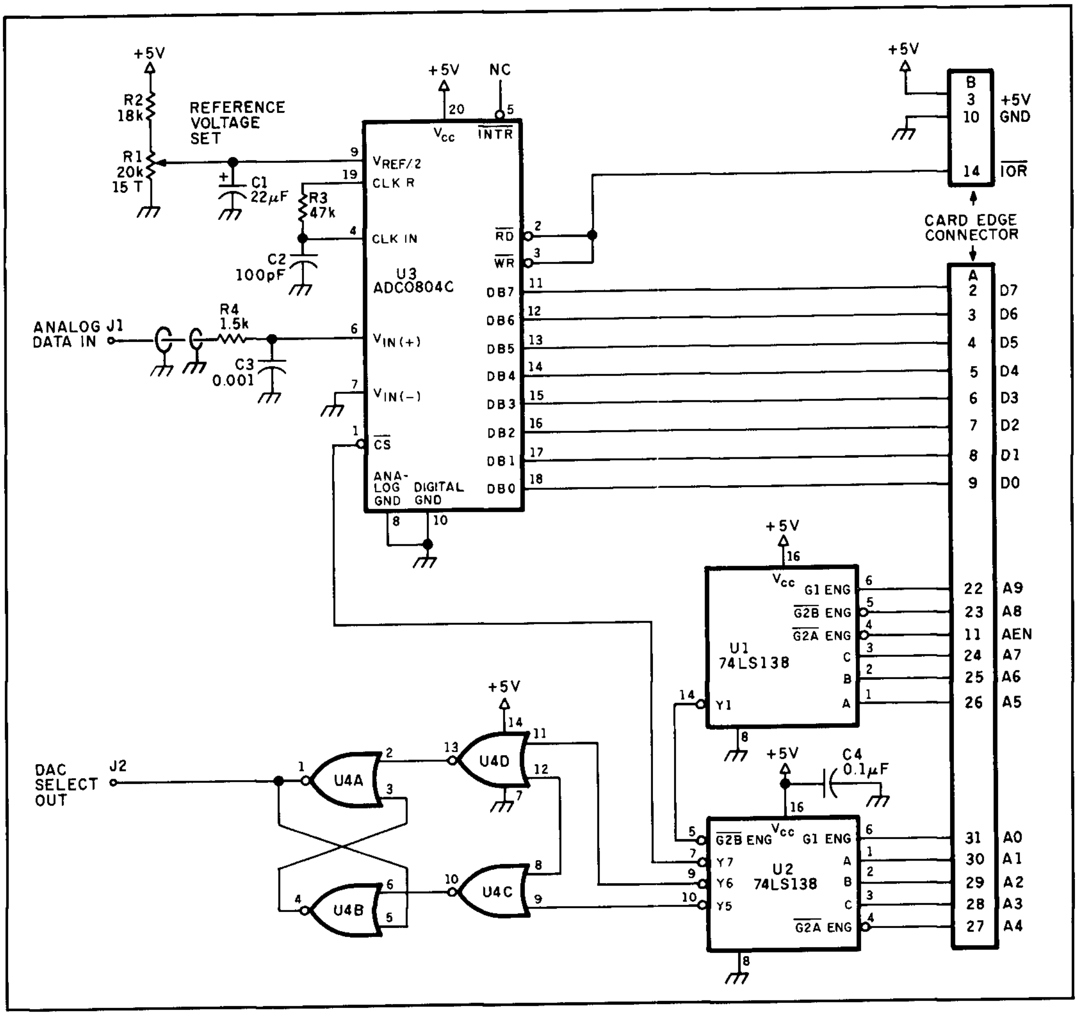


Figure 4. Schematic for the digitizer board.

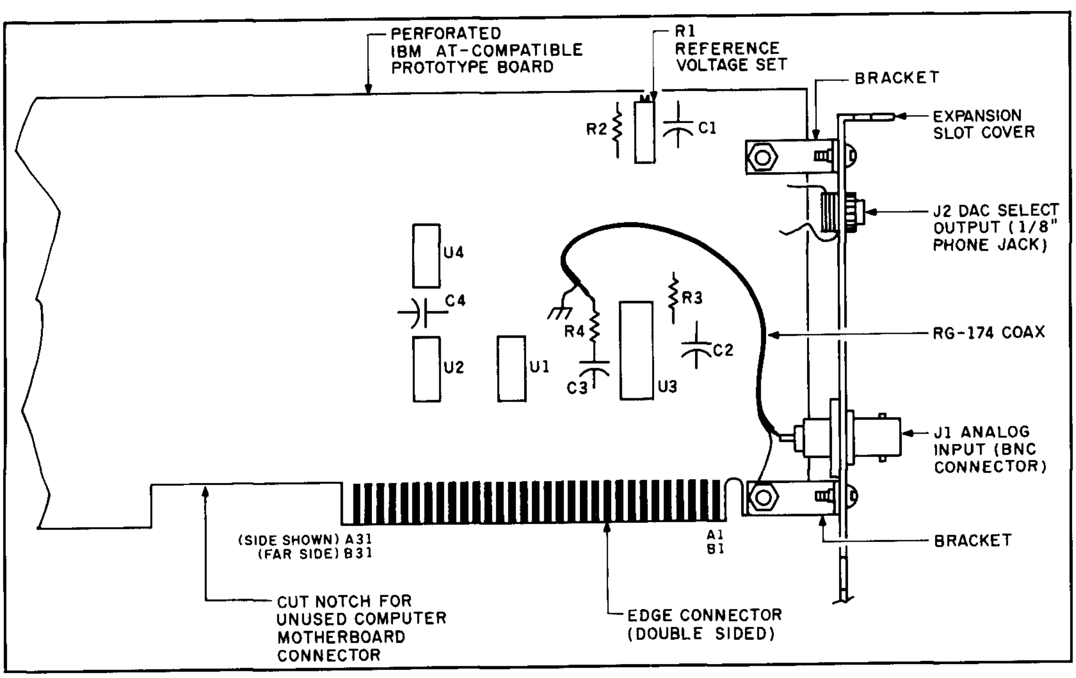


Figure 5. Layout of the digitizer board.

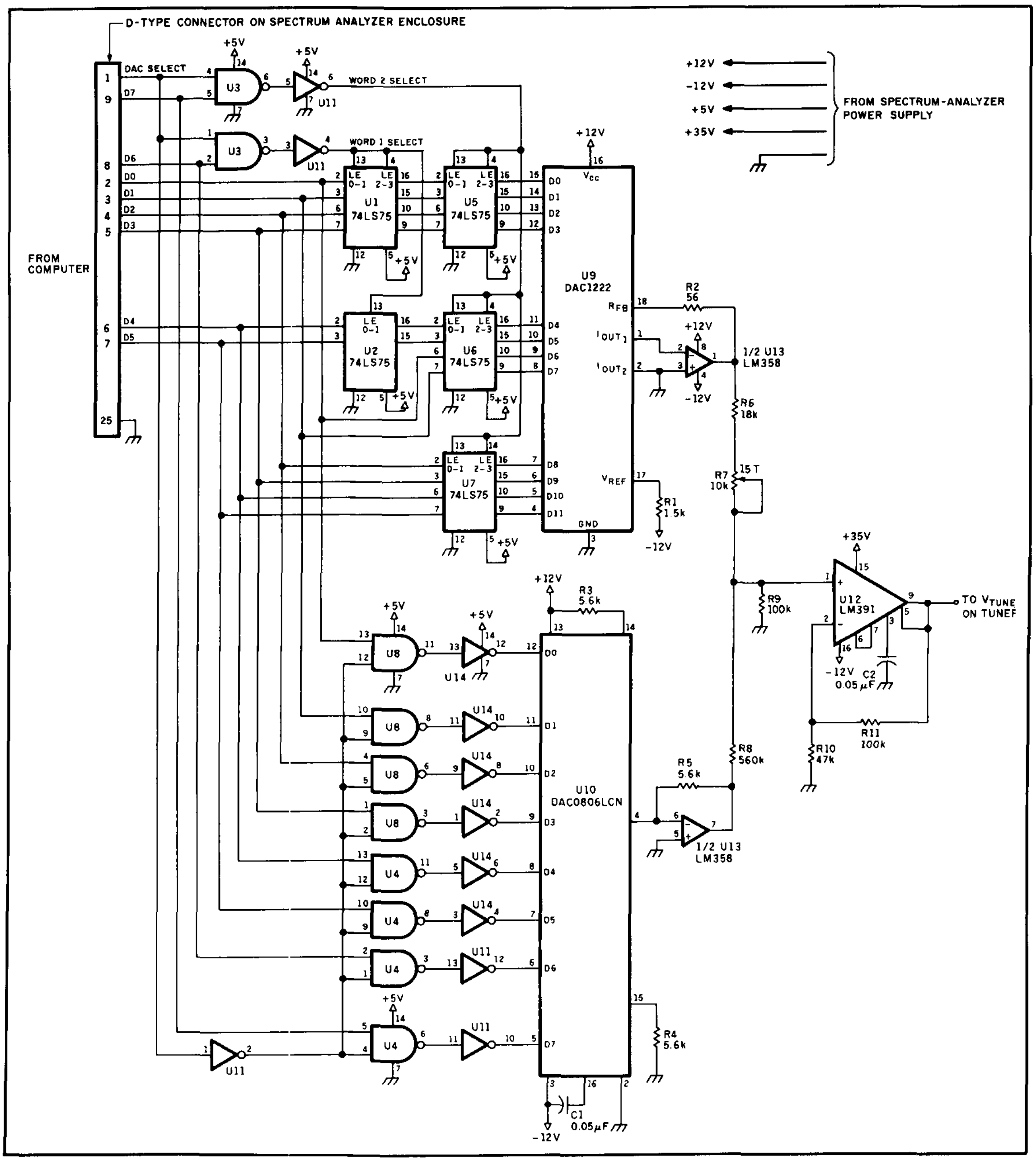


Figure 6. Schematic for the frequency control board.

at the rear of the computer.

I built the board on a perforated prototype board made especially for the computer. These boards are available from most electronics hobby stores and mail-order suppliers for about $25. I used point-to-point wiring. All ICs are mounted in sockets to facilitate repair.

The component layout isn't critical. **Figure 5** shows the layout and edge-connector pinouts of the digitizer board. Note that the edge connector is double sided. In the schematic, pin numbers A1 through A31 refer to edge-connector pins on the component side of the board. Pin numbers B1 through B31 refer to edge-connector pins on the solder side. I made a cutout in the board to clear the unused expansion slot receptacle. R1 is mounted near the top of the board. With the computer cover removed, you can adjust R1 while the board is mounted inside the case.

Installing the digitizer board

The analog input and DAC-select line are the only connections external to the computer. I attached the slot cover from the slot in which the digitizer board is installed to the rear end of the board (see **Figure 5**).

Two small brackets stand the cover off from the board so the cover fits properly against the computer rear frame when the digitizer board is in the machine. I mounted a BNC connector on the slot cover for the analog input and a 1/8-inch phone jack on the slot cover for the DAC-select signal.

Frequency control board

The schematic for the frequency control board is shown in **Figure 6**. The circuit uses two DACs. A 12-bit DAC decodes the start frequency; an 8-bit DAC sweeps the frequency from the start to the stop frequency.

System logic

Both DACs are controlled through the computer's parallel-printer port. The DAC-select signal generated on the digitizer board chooses the DAC to be addressed by the printer port. When the DAC-select line is a logic high, the 12-bit DAC is selected. Conversely, when the DAC-select line is a logic low, the 8-bit DAC is selected.

The parallel-printer port is an 8-bit port. To control the 12-bit DAC, two 8-bit words – Word 1 and Word 2 – are sent from the computer. Each word contains six bits of data. The remaining two bits identify whether Word 1 or Word 2 is being sent. Bits D0 through D5 from the parallel-printer port contain the data. Bits D6 and D7 are the word identifiers. The identifier for Word 1 is a logic 0,1 for bits D6 and D7, respectively; the identifier for Word 2 is a logic 1,0 for bits D6 and D7, respectively.

The 6-bit data of Word 1 from the parallel printer port are temporarily stored in latches U1 and U2. The data are latched under command of the Word 1 latch signal generated by U3 and U11. The 6-bit data stored in U1 and U2 are then stored in latches U5 and U6. The 6-bit data of Word 2 from the parallel-printer port are stored in latches U6 and U7 under command of the Word 2 latch signal generated by U3 and U11. Thus the 12-bit data required by the 12-bit DAC are stored in latches U5, U6, and U7, following two consecutive outputs from the parallel-printer port. The 12-bit DAC (U9) will then output a current proportional to the 12-bit word present at its inputs. U13 converts U9's output current to an output voltage.

The DAC-select signal is inverted and applied to one input of NAND gates U4 and U8. When the DAC-select signal is low, these gates are activated and the 8-bit data from the parallel-printer port are applied to 8-bit DAC U10's inputs. The computer generates a series of 8-bit words and outputs them to U10 to generate a ramp at U10's output. U13 converts U10's output current to an output voltage.

The voltage-converted outputs of the two DACs are summed by non-inverting amplifier U12. (U12 is an audio power driver intended for use in high-fidelity audio amplifier designs.) Because it can operate with high power supply voltages, I used it in this application as a high-voltage op amp.

The tuners used in the spectrum analyzer require tuning voltages ranging from zero to about 30 volts to cover their wide frequency range. U12's output can produce the required tuning voltages and drive the tuners directly. The op amp doesn't need to deliver a large current because tuning is accomplished inside the tuners with reverse-biased variable-capacitance diodes. Potentiometer R7 is used to calibrate the start frequency.

Control board construction

I built the frequency control board on a perfboard using point-to-point wiring. All integrated circuits are mounted in sockets to facilitate repair. Once again, the board layout isn't critical. The board is mounted inside the spectrum analyzer enclosure. I added switch S2 in series with the tuning inputs of the two tuners so I could select manual (internal) or computer (external) control. See **Figure 1** for details.

Interconnecting cable

The computer and the frequency control board in the spectrum analyzer are connected via a 10-conductor cable. I mounted a 25-pin D-type connector on the spectrum analyzer enclosure for connection to the cable. One end of the cable has a 25-pin D-type connector compatible with the one on the spectrum analyzer. The computer end of the cable has two connectors. One is a 25-pin D-type compatible with the computer's parallel-printer port. The other connector is a 1/8-inch phone plug. This plug carries the DAC-select signal and fits into jack J2 on the computer slot cover mounted on the digitizer board. The pinouts for the D-type connectors on the spectrum analyzer and computer end of the cable are shown in **Figure 7**.

Computer program

I originally wrote the computer program in Microsoft Quick BASIC for use with an EGA monitor so I could tap into the high-resolution graphics capability. I've included line numbers on all lines of the code (Quick BASIC doesn't require line numbers for all lines) so the program can run with the more common BASIC versions like Microsoft GW-BASIC.

The program runs much slower in GW-BASIC. This means the frequency sweep generated by the computer will be slower, and you may miss some of the intermittent signals you want to "catch" in the frequency spectrum. For the most part, the speed of the BASIC program or the computer isn't a critical parameter.

I first wrote the program for use with parallel-printer port 2 (LPT2), with the spectrum analyzer connected to LPT2 and the printer connected to parallel-printer port 1 (LPT1). Using this arrangement eliminates the need to switch connections between the printer and spectrum analyzer at the parallel-printer port when I want a hard copy of the spectrum plot. I also have a version for use with LPT1 if you have only one parallel printer port.

The computer program contains about 300 lines of code and is too long to be included with this article, but I'll send you a copy of the program. My address appears at the beginning of the article. A self-addressed stamped envelope would be appreciated. I can also supply the program on a diskette for $5. Please state whether you want the LPT1 or LPT2 version.

A flowchart for the computer program is shown in **Figure 8**. The program begins with the establishment of four arrays. Arrays A and B contain the digitized data. The digitized spectrum analyzer data are stored immediately upon acquisition in Array A. Array B stores the digitized data after it's scaled to an equivalent dBm input to the spectrum analyzer. Arrays A and B both contain a maximum of 256 elements, with each element corresponding to a sample of spectrum analyzer's detected output. Because an 8-bit DAC is used for the fre-

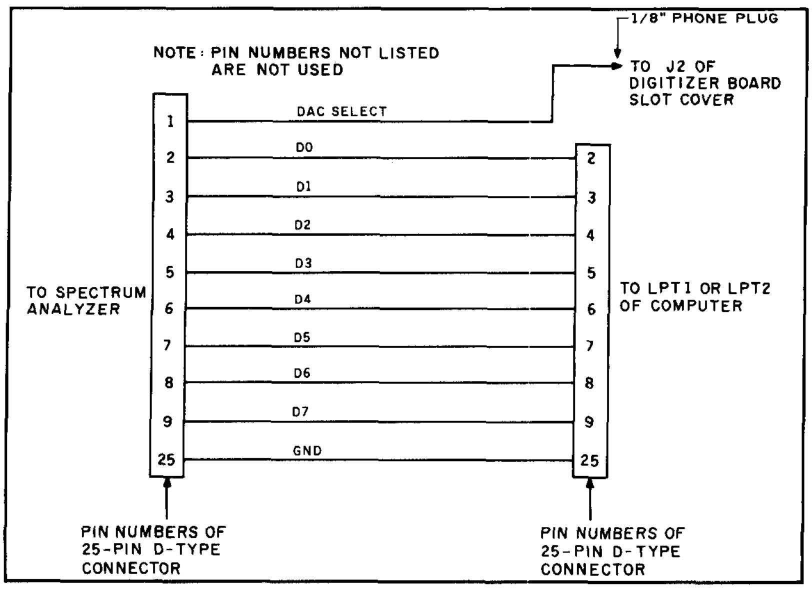


Figure 7. Interconnecting cable.

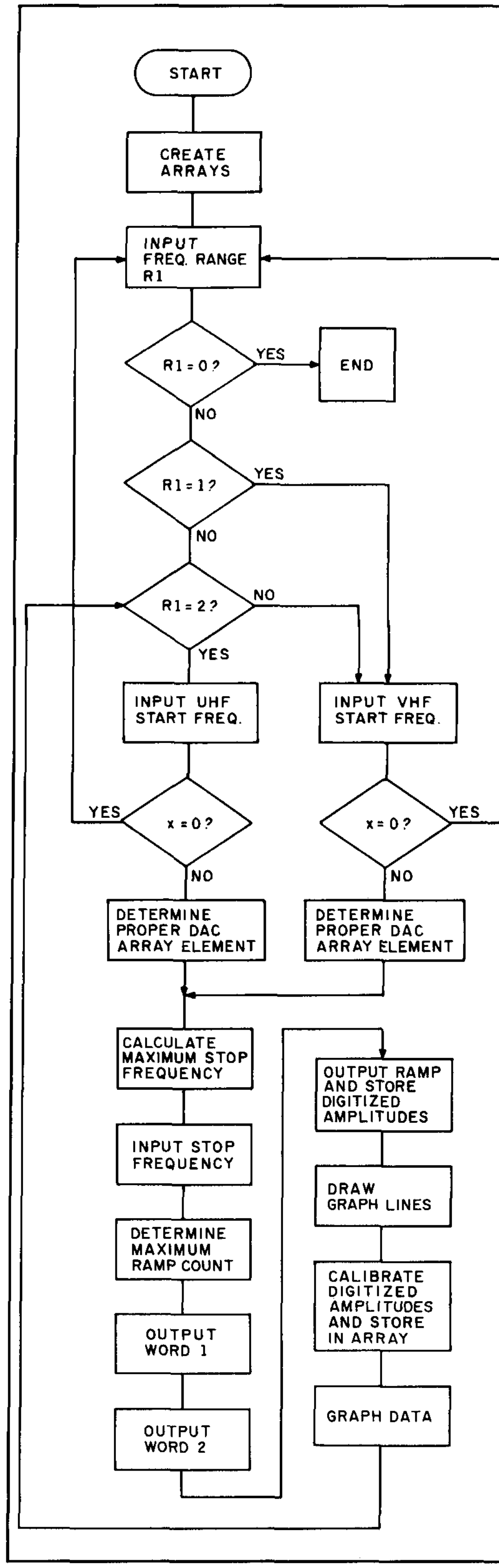


Figure 8. Flowchart for the computer program.

quency sweep, a maximum of 256 samples is possible (that is, 2 to the 8th power). Therefore, 256 array elements are required. Arrays C and D have 4098 elements each. These arrays contain the data for the two words that are sent to the 12-bit DAC for establishing the start frequency. The data for Word 1 are stored in Array C and the data for Word 2 in Array D. The program automatically creates the data for arrays C and D and stores them in the proper locations within these arrays before the program continues. The program then prompts you for the desired frequency range. There are three possibilities: 0 to quit, 1 for the VHF tuner (1 to 500 MHz), and 2 for the UHF range (393 to 900 MHz). You'll then be prompted for the desired start frequency.

After you've input the start frequency, the look-up tables in the program are used to determine the elements of Array C and Array D corresponding to the desired start frequency. The program includes two look-up tables — one for the VHF tuner and the other for the UHF tuner. The tuning curves for the two tuners aren't linear across the wide tuning ranges. **Figure 9** shows a plot of the tuning voltage versus input frequency for the VHF tuner. As you can see, the curve is very nonlinear. The UHF tuner has a similar tuning voltage versus input frequency characteristic. I've divided the tuning curve of each tuner into several small linear segments and stored equations for these segments in the look-up tables. The program selects the proper segment based on the selected start frequency, and determines the one element out of 4098 that corresponds to the frequency nearest the one selected. The element number (1 to 4098) is stored in variable f1.

Next, the maximum allowable stop frequency, based on the selected start frequency, is calculated and displayed. This frequency is the upper limit of the permissible frequency sweep, based on the start frequency selected. (Remember that the frequency sweep is generated by an 8-bit DAC.) The computer outputs a series of 8-bit words ranging from 000 to a maximum of 256 to generate a ramp which is applied to the tuners' input. A count of 256 is the maximum value available with 8 bits and thus represents the largest ramp that can be produced. This means a count of 256 will represent the maximum available frequency span.

Because the tuning curves of the tuners are nonlinear, a maximum frequency span corresponding to a count of 256 isn't constant across the tuners' frequency range. Therefore, the maximum allowable stop frequency must be calculated for each start frequency selected.

The program prompts you for the desired stop frequency. After this frequency is input, the program converts it to the nearest number from 000 to 256 corresponding to that frequency. That number is stored in variable f2. The program then outputs data to the frequency control board telling it to set up the start frequency using the 12-bit DAC. The appropriate element of Array C, corresponding to Word 1, is sent first. This is followed by the appropriate element of Array D, corresponding to Word 2.

After a brief delay routine is completed, the frequency ramp output is generated, and the detected output amplitude of the spectrum analyzer is sampled by the digitizer board and stored in Array A. The detected output is sampled and the data stored for each element of the ramp. The ramp can contain a maximum of 256 elements (that is, the computer counts up to a maximum of 256), so the maximum number of digitized samples equals 256.

Next, the program creates the display screen by drawing the border and amplitude lines and labeling the axes. The digitized samples stored in Array A are then "calibrated" and converted to dBm (decibel referenced to 1 mW) using another set of look-up tables. I created the look-up tables for the VHF and UHF tuners from measurements I made of the tuners using a laboratory-grade signal generator and oscilloscope. The calibrated amplitudes are stored in Array B.

Finally, the data stored in Array B are graphed on the screen, and the program returns to the first prompt for a start frequency. At this point, you can output the graph to the printer using the "print screen" command if you wish.

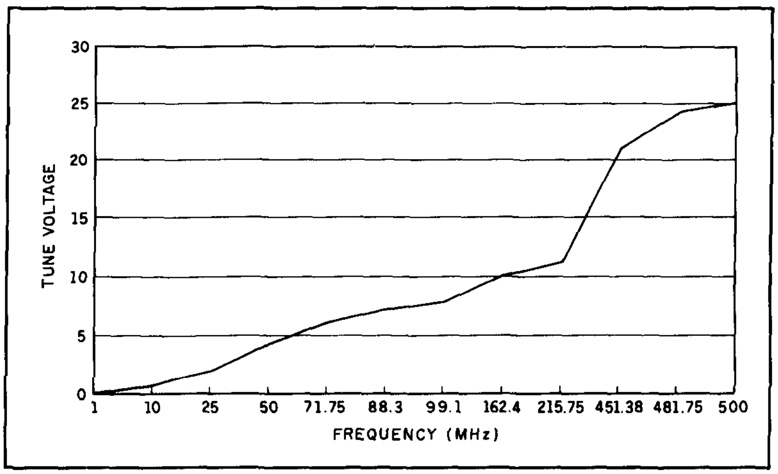


Figure 9. VHF-tuner curve.

Installation and alignment

Once you've assembled the digitizer board, carefully inspect it for any wiring errors, shorts, or bad solder joints. Install the digitizer board in the expansion slot desired. Adjust potentiometer R1 on the digitizer board to a point approximately midway between the extremes of its range.

Digitizer alignment

To align the digitizer board, you must apply +2.5 volts DC to its analog input. I used a 12-volt DC power supply and a potentiometer connected as a voltage divider between the power supply output and ground. The potentiometer's wiper was connected to the BNC connector on the digitizer board through a coaxial cable. An oscilloscope was used to measure the voltage at the wiper precisely. To avoid loading down the circuit, use a high-impedance device like an oscilloscope or digital voltmeter (DVM) to make this measurement. Verify that the voltage is about +2.5 volts before connecting this voltage to the computer.

Computer interface

Turn on the computer and load the spectrum analyzer program into BASIC. Connect the test voltage to the BNC-input connector on the digitizer board and readjust the voltage for precisely +2.5 volts DC.

Run the spectrum analyzer program. The words "PLEASE WAIT" should appear on the screen. The computer is now filling Arrays C and D with the proper data for the 12-bit DAC. This takes a little time, especially if you're using GW-BASIC. The wait message appears to let you know that the computer is working.

In a few moments, the screen prompts you to select the desired frequency range. Type a 1 to choose the VHF range. The computer now prompts you for the desired start frequency. Type 100. The computer asks you for the desired stop frequency. Type 132, and a graph will appear on the screen. After a pause, a horizontal line corresponding to the input voltage level will be plotted on the graph. Below the graph, you'll again see a computer prompt for the desired start frequency.

The test voltage of +2.5 volts DC corresponds to a −60 dBm input to the VHF tuner. If the line that was plotted on the graph isn't drawn at the −60 dBm mark, readjust R1 on the digitizer board and again select 100 as the start frequency and 132 as the stop frequency. Observe the new plot. Repeat this operation until the line is plotted at the −60 dBm mark.

Frequency-control board alignment

Remove the test voltage from the input to the digitizer board and connect the spectrum analyzer detected output to the digitizer input board. Connect the 10-conductor cable from the spectrum analyzer to the computer parallel-printer port. Next, connect the DAC-select line to the digitizer board's 1/8-inch phone jack. Configure the spectrum analyzer so the frequency control board controls the VHF tuner's frequency, and the output of the VHF tuner is connected to the receiver board. With my spectrum analyzer, all I have to do is select low range (VHF) and external control with the two toggle switches, S1 and S2, and reconnect the antenna to the VHF tuner (see **Figure 1**). Adjust R7 on the frequency-control board to the center of its range.

Perform the frequency-control board alignment using a known frequency as an input to the spectrum analyzer. Try the frequency of an FM commercial radio station. At the prompt for select start frequency, type in the frequency of a local FM station. (I used 89.3 MHz.)

At the select stop frequency prompt, type in the maximum stop frequency that the computer calculated and displayed. After a pause, you should see a spectrum plot displayed on the screen. When the plot is complete, listen to the audio output from the spectrum analyzer. The spectrum analyzer will be "resting" on the start frequency that you selected.

If your R7 setting is correct, you should hear the radio station in the speaker. If you don't hear the proper station, readjust R7 on the frequency-control board. Align the frequency-control board using several different stations as your start frequency. Adjust R7 so you can hear the majority of them to some extent.

The 12-bit DAC used for the start frequency control generates only discrete voltages. It won't be able to position the start frequency precisely on all selected start frequencies. Consequently, you won't hear all the FM stations that you use for calibration clearly. If R7 is properly adjusted, the frequencies 89.3 MHz and 96.1 MHz are easy to hear when used as start frequencies. Use these frequencies for calibration, if they are available in your area.

As an alternative, you can use a signal generator as the signal source, adjusting R7 so the spike for the generator frequency is plotted at the left of the graph. If you choose this approach, you should also try several different frequencies. Adjust R7 for the best compromise.

That's all you have to do to align the digitizer and frequency-control boards. Now, at the start frequency prompt, type 0. The computer asks you to select the desired frequency range. Type 0 to end the program.

Conclusions

I've found that these modifications provide an interesting versatility to WA2PZO's spectrum analyzer[4] While the computer interface doesn't make the "Poor Man's Spectrum Analyzer" equal in performance to more expensive laboratory grade equipment, it does allow some reasonably accurate measurements. ■

References

1. Robert M. Richardson, W4UCH, "Low Cost Spectrum Analyzer with Kilobuck Features," *HAM RADIO*, September, 1986, pages 82-90.
2. Joe Carr, K4IPV, "Building the 'Poor Man's Spectrum Analyzer,' " *HAM RADIO*, March 1987, pages 99-104.
3. Joe Carr, K4IPV, "Revisiting the 'Poor Man's Spectrum Analyzer,' " *HAM RADIO*, September 1987, pages 49-52.
4. Murray Barlowe, WA2PZO, "Add a Digital Readout to the 'Poor Man's Spectrum Analyzer,' " *HAM RADIO*, September 1988, pages 84-93.

Parts List: Digitizer Board

C1	*22-μF tantalum*
C2	*100-pF disc ceramic*
C3	*0.001-μF disc ceramic*
C4	*0.01-μF disc ceramic*
(All capacitors 12 volts DC or higher)	
J1	*BNC-type connector (chassis mount)*
J2	*1/8-inch phone jack (chassis mount)*
R1	*20-k, 15-turn potentiometer*
R2	*18 k*
R3	*47 k*
R4	*1.5 k*
(All fixed resistors 1/4-watt, 10 percent)	
U1	*ADC0804C (8-bit analog-to-digital converter)*
U2	*74LS02 (quad 2-input NOR gate)*
U3,U4	*74LS138 (1-of-8 decoder/demultiplexer)*

Parts List: Frequency Control Board

C1,C2	*0.05 μF disc ceramic (All capacitors 35 volts DC or higher)*
R1	*1.5 k*
R2	*56*
R3,R4,R5	*5.6 k*
R6	*18 k*
R7	*10-k, 15-turn potentiometer*
R8	*560 k*
R9	*100 k*
R10	*47 k*
R11	*100 k*
(All fixed resistors 1/4-watt, 10 percent)	
U1,U2,U5,U6,U7	*74LS75 (4-bit bi-stable latch)*
U3,U4,U8	*74LS00 (quad 2-input NAND gate)*
U9	*DAC1222 (12-bit binary multiplying D/A converter)*
U10	*DAC0806LCN (8-bit D/A converter)*
U11,U14	*74LS04 (hex inverter)*
U12	*LM391 (audio power driver)*
U13	*LM358 (low power dual op amp)*

By Peter J. Bertini, *K1ZJH*
20 Patsun Road
Somers, Connecticut 06071

TRANSMITTER IMD MEASUREMENTS ON UTILITY-GRADE SPECTRUM ANALYZERS

Measuring SSB transmitter IMD characteristics on inexpensive spectrum analyzers

Utility grade spectrum analyzers are becoming commonplace in the Amateur workshop. They typically cost under $2000 (under $500 for one particularly good kit!*), and let you see transmitter harmonics, mixer spurs, and other anomalies spread many kHz across the spectrum.

Background

Spectrum analyzers are useful for measuring transmitter intermodulation distortion (IMD) performance. But instruments with this capability cost many thousands of dollars, and aren't available to most Amateurs. Many hams make do with the shop scope, using trapezoid or similar patterns to see when a gross maladjustment has been made. Scopes work in the time domain, and have a display which provides a linear representation of the input signals.

Spectrum analyzers work in the frequency domain and use a log display calibrated in dB. Both instruments (scope and analyzer) will do the job. But while a log display can easily handle a 60-dB signal differential, it represents a voltage ratio of 1,000:1. Obviously, the spectrum analyzer will show problems long before they are noticeable on your shop scope.

* A&A Engineering, 2521 West LaPalma, Unit K, Anaheim, California 92801. Phone: (714)952-2114.

SSB analyzers

Spectrum analyzers must resolve signals in narrow swept bandwidths for conventional IMD measurements. The filters used in an SSB analyzer are quite different from those used in a communications receiver. While the desired rectangular shape factor offered by most modern SSB filters is good for receivers, the filters won't work in analyzers.

A conventional SSB filter produces severe phase discontinuities on its slopes when swept rapidly across a carrier. The end result is not unlike that produced by pulse-type noise interference. The signal is distorted and broadened.

SSB filters for analyzers have carefully designed shape factors matched to the sweep rates of the analyzer. They are expensive and beyond the reach of most home experimenters. Not only is the analyzer itself expensive, but SSB IMD capabilities are often an equally expensive option on most professional models.

Voltage-controlled oscillator (VCO) noise limit performance

Another limiting factor is VCO noise and jitter. Spectrum analyzers use swept local oscillators operating in the VHF-to-microwave range. The VCO noise characteristics of most of these oscillators, even in comparatively expensive analyzers, would result in reciprocal mixing products which would render narrow resolution displays useless.

Conventional IMD measurements

The conventional method of making transmitter IMD measurements for spectral displays uses two clean, unharmonically related tones injected into the transmitter microphone input. When the tones are within the SSB filter bandpass (as determined by the beat-frequency oscillator, or BFO, and filter bandpass), and set for equal amplitude, the result is two RF carriers separated by the difference in tone frequencies.

It's important to note one fact. The maximum frequency spread between the two tones is limited by the transmitter's SSB filter bandwidth. This, in turn, limits you to a specific grade of analyzer. But don't despair! If you have a utility grade spectrum analyzer with 15-kHz filters and 50 dB of dynamic range, which can cover the bands of interest, you can make valid IMD measurements on your SSB transmitter.

I use an A&A Engineering spectrum analyzer that meets these requirements, and costs well under $1000. Let me show you how.

Test procedure

If you are an Amateur who homebrews his own transmitters and has a need for IMD measurements, or if you have no qualms about tearing into a commercial product to make these tests, this procedure is for you. My method involves taking a different tack than others. For the most part, IMD products generated ahead of the SSB filter can only result in the generation of in-band products. In most cases these can be disregarded.

I eliminate the filter stage and start the process from there. By using two calibrated signal generators or crystal oscillators, I'm able to generate and combine the signals and inject an IF signal at the filter output. If the oscillators are set 50-kHz apart, you should observe two carriers at the transmitter output separated by the same amount. With the carriers separated by 50 kHz, the IMD products are easily resolved on the inexpensive analyzer.

Caution: disconnect filter output

Be sure to disconnect the filter output. It's important to do so for two reasons. First, you'll eliminate any residual carrier feedthrough. Second, the filter presents a low impedance termination within its bandwidth and a high impedance outside it. Leaving the filter in line might lead to an unwanted imbalance of the injected IF carriers.

Equipment needed

I use two Clemens SG-83 generators with calibrated step attenuators and variable outputs. You can also use fixed frequency crystal oscillators, but you must provide some method of setting individual and combined output levels. The generator outputs are set to produce the rated transmitter output power and equalize the two RF carrier levels. I use the combiner shown in DeMaw and Hayward's book[1] to combine and isolate the generator outputs (see **Figure 1**). If you don't do this, one generator may modulate the other and cause products that will fall in the IMD frequencies.

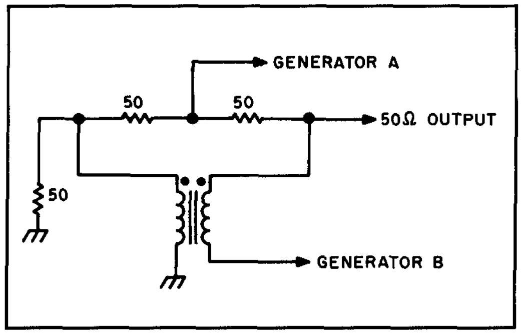


Figure 1. Combiner adds signal-generator outputs. Each generator "sees" the other's signal in-phase and canceled, preventing cross modulation. Transformer is ten turns of no. 30 wire, bifilar wound on FT-23-43 ferrite toroid.

Obviously, my method will work at any point in the transmitter's RF chain. You can view mixers, bandpass filters, IF amplifiers, and other sources which generate IMD. Gain distribution is important in transmitters and receivers. That innocent looking IF stage may be generating all of the IMD, well before gain compression occurs in the PA stage.

Additional generator amplification

At some point, you may find it necessary to use additional amplification (at the transmitter predriver, for instance) to achieve rated transmitter power. DeMaw and Hayward's book[1] gives examples of fixed-gain feedback amplifiers suitable for this purpose. Use "strong" devices, like 2N3866s, to avoid IMD generation in the test setup — even for signal levels under 0 dBm.

Figure 2 shows a suggested amplifier circuit. When making changes, I always check the combiner/amplifier output directly on the analyzer to ensure that IMD products aren't being generated by the test setup. Then I look at the transmitter output. **Figure 3** shows my layout.

Alternative points for injection

My method lends itself to most modern solid-state designs. They usually use broadband bandpass filters between stages, and can handle the 50-kHz signal spread easily. If your's can't, use a point further down the transmitter chain. Hybrid or all tube-type transmitters employing tunable preselectors and relatively sharp IF interstage transformers, may be a little trickier to set up.

Set the preselector to straddle-tune the center frequency. Carefully straddle the generators on either side of the IF frequency, and use the minimum frequency spread that still gives good resolution on the analyzer. It helps to select the highest frequency covered by the transmitter. (A sharp preselector that won't pass a 50-kHz spread on 80 meters will probably do so at 28 MHz.) You'll find that 455-kHz IFs are too low in frequency. The arithmetical selectivity offered by any tuned circuits in this range won't pass signals with a 25-kHz spread.

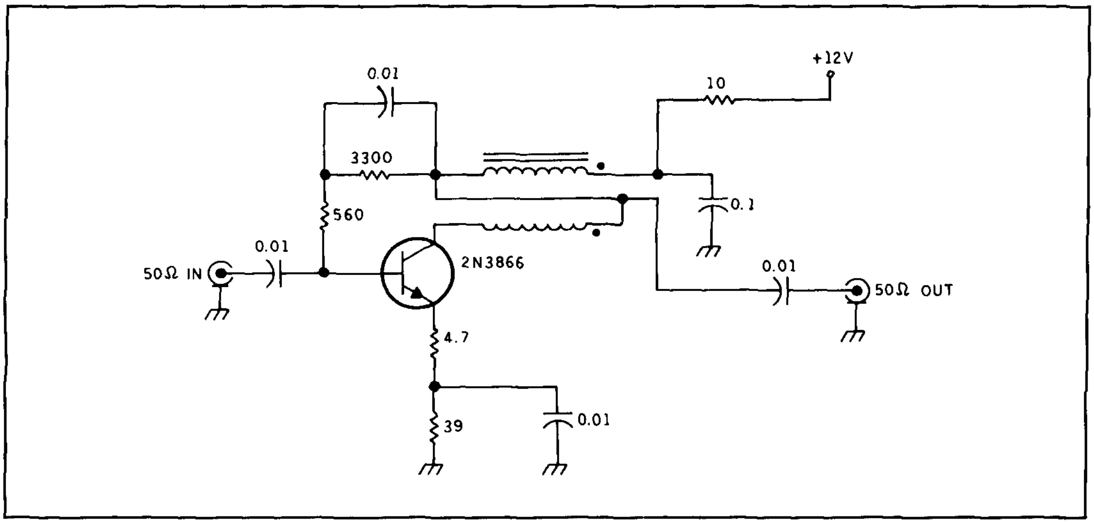


Figure 2. Fixed gain amplifier stage boosts combined generator output where needed. Stages may be cascaded for more gain. Transformer is ten turns no. 30 wire, bifilar wound on FT-23-43 ferrite toroid.

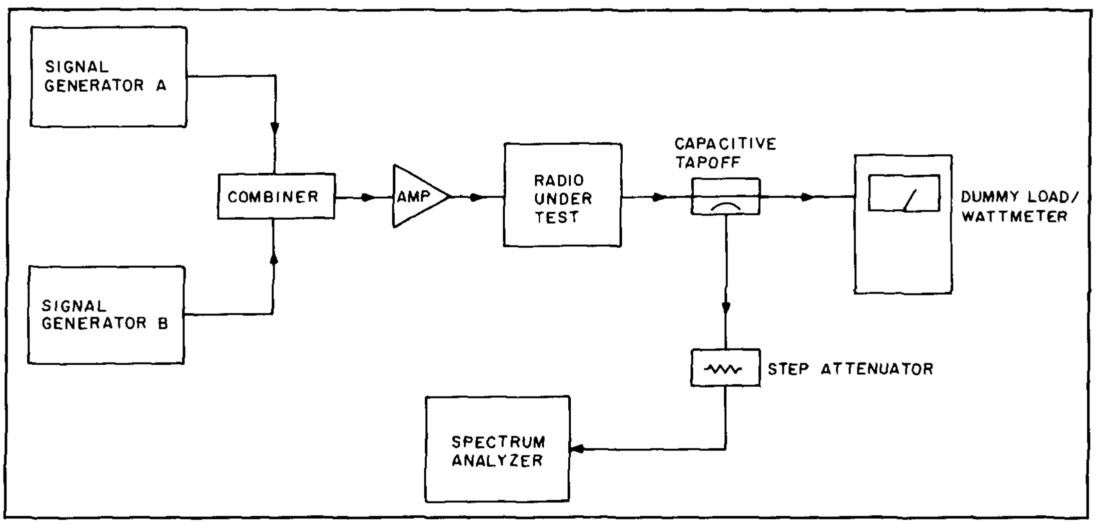


Figure 3. Test setup used by author. Output from two Clemens signal generators is combined (*Figure 1*) and amplified (*Figure 2*) as needed. Transmitter output is sampled via capacitive tapoff, about 30-dB loss, and further attenuated by Tektronix step attenuator before spectrum analyzer.

Because most IMD generation occurs in the transmitter PA and driver stages, it does no harm to find an injection point at the highest frequency – even if it means going to a predriver stage at 28 MHz.

With most solid-state transmitters, the interstage coupling is of low impedance, and the 50-ohm output from the combiner or amplifier will drive most stages without additional matching. For tube transmitters or high impedance points, try using link coupling into interstage transformers. Or use a simple matching circuit to achieve adequate injection voltage.

Measuring IMD products

The generators are set to produce equal-amplitude RF carriers in much the same way as the audio method. Because the transmitter power is being divided equally into two carriers, many manufacturers rate the IMD products 6-dB down from that shown on the display. This is done by shifting the carrier display down 6-dB from the 0-dB graticule while measuring the IMD products from the 0-dB line.

Single-tone output would be at the 0-dB graticule. This is also the reference point for harmonic and carrier-suppression measurements. The ARRL lab uses this method. A more stringent analysis (sometimes called the Collins method) doesn't allow the 6-dB offset; the IMD products are referenced to the peak of either carrier.

Some inexpensive analyzers have limited dynamic range (50 dB or less), so take care not to overload them. I use a capacitive tapoff between the transmitter and dummy load, followed by a step attenuator to keep signals well below the point where gain compression occurs in my analyzer. IMD products are generally no better than 35 dB down. The analyzer's dynamic range is rarely a limiting factor.

A third method

Though I haven't tried it, it might be possible to hardwire a jumper around the filter (assuming the balanced modulator has particularly good carrier rejection) and insert a 25-kHz audio signal directly into the balanced modulator's AF input. This would produce a DSB signal with carriers separated by twice the audio frequency – in this case, 50 kHz. This scheme might be worth trying if you don't have access to RF generators, or if the radio IF or predriver stages are overly selective. ■

Reference

1. Doug DeMaw, W1FB, and Wes Hayward, W7ZOI, *Solid State Design for the Radio Amateur*, ARRL, 1977.

By Jacob Makhinson, *N6NWP*
1100 N. Sunset Canyon Drive
Burbank, California 91504

HIGH DYNAMIC RANGE RECEIVER

This receiver's modular design encourages experimentation

How often have you been interrupted by a loud, raspy signal clobbering a good portion of the band during a QSO with a weak station? Your radio's front end may be the cause.

On one hand, receiver sensitivity has reached a level where further refinements would be of little benefit – especially for Amateurs residing in densely populated areas. On the other hand, the current levels of intermodulation, compression, and desensitization could still be improved. The techniques for upgrading these receiver parameters have been known and used for years. I've found a way to improve the receiver dynamic range – the most important receiver parameter for an Amateur residing in an urban area.

I had several reasons for designing this receiver. Aside from an experimenter's curiosity, I had a long-standing dream to own a receiver with an "uncrunchable" front end. I'm sure this feeling is shared by many urban dwellers whose receivers are subject to never-ending assaults from strong local stations.

System considerations

I wanted to design a receiver for the 14-MHz band capable of handling strong signals several kilohertz away from a desired weak DX station. **Figure 1** shows the block diagram.

I found simplicity in a single-conversion approach. A low-loss input bandpass filter protects the mixer from strong out-of-band signals. Traditional LC-type capacitor tuning in the VFO reduces far-out phase noise. Additional phase noise reduction is achieved by using a frequency divider on the VFO output.

My mixer circuit uses a double-balanced MOSFET mixer for excellent dynamic range. The mixer is followed by a diplexer to ensure that the mixer IF port is properly terminated at all frequencies of interest. A high dynamic range post-mixer amplifier compensates for mixer loss.

A four-stage IF amplifier using discrete transistors provides high gain coupled with low noise level and adequate AGC range. The AGC is IF-derived with full "hang" action. A pair of matched crystal filters yields an extremely steep-skirted IF frequency response and helps to confine the IF-generated noise to the portion of the spectrum that contains information. The product detector is a diode double-balanced mixer for best IMD performance.

The audio amplifier is preceded by a set of low and high-pass filters to minimize 60-Hz hum and enhance the overall signal-to-noise characteristic of the receiver.

I began my design with the mixer – the most important part of a high-performance receiver. I managed to attain a third-order input intercept point of +43 dBm with a conversion loss of 8 dB in the mixer. An RF amplifier would have improved the receiver noise figure, but I decided to feed the signal from the antenna directly into the mixer (through the input BPF).

The RF amplifier would have required a third-order output intercept point of better than +43 dBm and a very low noise figure to prevent degradation of the spurious-free dynamic range (SFDR). To preserve high levels of 1-dB compression point and 1-dB desensitization point it's prudent to have as little gain as possible before the mixer. However, the levels of atmospheric and manmade noise in urban areas are of such magnitude that the use of an RF amplifier results in only a marginal improvement on the 14-MHz band.

Because the mixer has a relatively high conversion loss, I've used a post-mixer amplifier to avoid overall noise figure deterioration. To preserve the dynamic range of the front end, the third-order input intercept point of the post-mixer amplifier (36.5 dBm) exceeds the third-order output intercept point of the mixer.

The IF amplifier provides most of the gain in the receiver (in excess of 90 dB), while two crystal filters provide the required selectivity and out-of-band noise suppression. The audio amplifier provides the additional gain

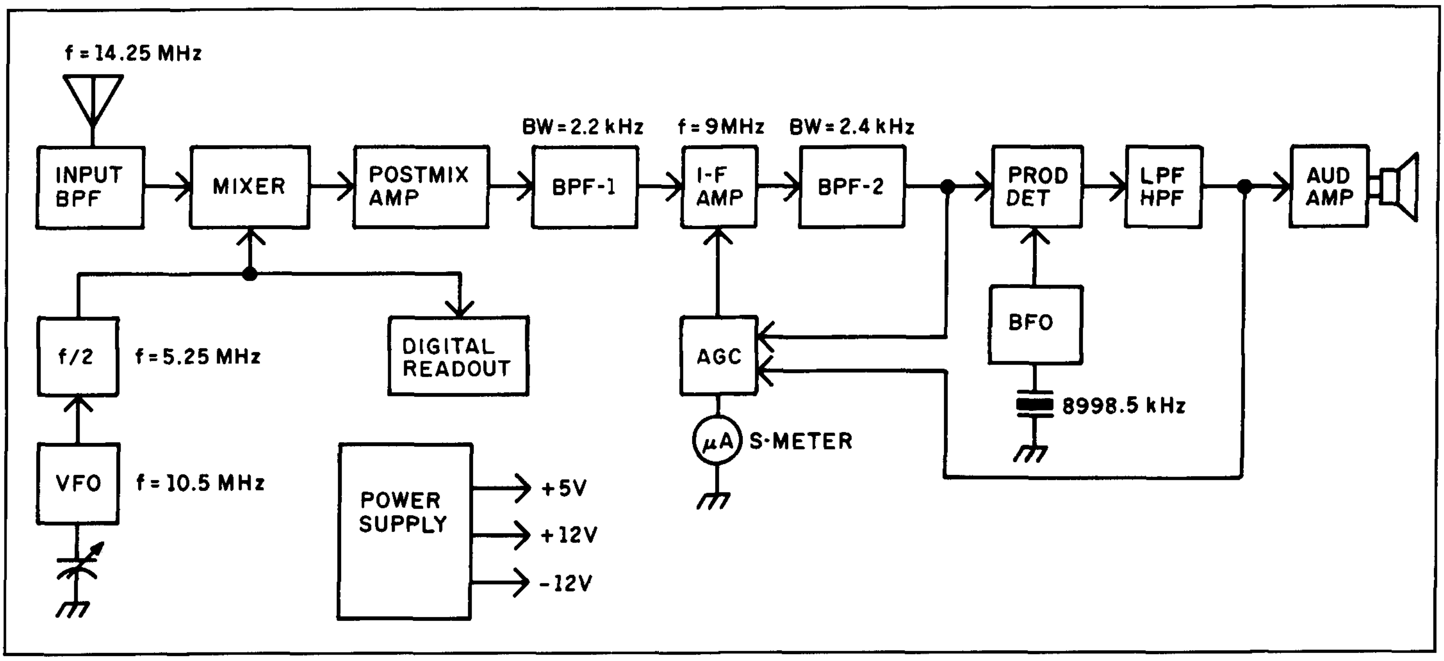
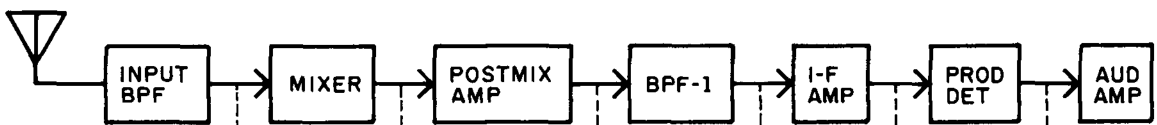


Figure 1. Block diagram of receiver. Circuit description and design goals for each block are covered in text.



	INPUT BPF	MIXER	POSTMIX AMP	BPF-1	I-F AMP	PROD DET	AUD AMP
GAIN AV (dB)	-0.5	-8	10.5	-3	92	-6	30
NOISE FIGURE NF (dB)	0.5	8	5	3	5	—	—
THIRD-ORDER INPUT INTERCEPT IP3in (dBm)	43.5	43	36.5	—	—	—	—
1dB COMP POINT Pc (dBm)	19	18.5	18.5	—	—	—	—
1dB DESENSITIZATION POINT Pd (dBm)	22	21.5	—	—	—	—	—

Figure 2. Gain distribution and noise figures through receiver.

(up to 30 dB) needed to drive the speaker.

You can use the data presented in **Figure 2** to find the receiver sensitivity and its dynamic range. Detailed calculations are given in the **Appendix**. The results of these calculations are summarized below:

Minimum Discernable Signal: MDS = −126 dBm, or 0.12 μV rms
Sensitivity at 10 dB S/N: Ps = −116 dBm, or 0.36 μV rms
Spurious Free Dynamic Range: SFDR = 113 dB
Blocking Dynamic Range: BDR = 148 dB

Circuit Description

Input band-pass filter. The preselector filter (**Figure 3**) is a half-octave Bessel type[1] chosen to keep second or higher-order products out of the front end pass-band and to provide steep attenuation beyond the cutoff points. The center frequency of the filter is: Fm = 14.2 MHz. The −3 dB points are: Fl = 12.35 MHz and Fh = 16.35 MHz.

I selected a five-pole maximally flat delay characteristic and calculated the component values from normalized tabulated values[2] The filter's insertion loss doesn't exceed 0.5 dB (see **Photo A**). The source and load impedances are: Rs = Rl = 50 ohms.

Capacitors with 5-percent tolerance are recommended. I used ceramic capacitors picked from a 10-percent tolerance stock in my prototype.

Mixer. I took extra care with the mixer design because this stage is primarily what determines the receiver's dynamic range. I conducted an extensive survey to find a mixer capable of providing the best dynamic range. For those interested in this subject, I recommend **References 3** through **9**. Dr. Ulrich L. Rohde's article, "Performance Capability of active mixers,"[7] deserves special attention. Rohde claims that mixers using field-effect transistors (FETS) without operating voltage (acting as switches) can provide a +42 dBm input intercept point.

A transceiver designed by V. Drozdov, RA3AO[3] which is currently the premier Soviet home-built transceiver, sports exceptionally high blocking dynamic range. It uses a passive mixer with active devices.

Once I chose the mixer scheme, I needed to select an active device. Siliconix manufactures several matched devices in quad packages which are unduly neglected by Amateur Radio builders. Among them are the U350 double-balanced mixer[6] (matched JFETs in a quad package) and the Si8901[5] (a monolithic quad DMOS FET designed for

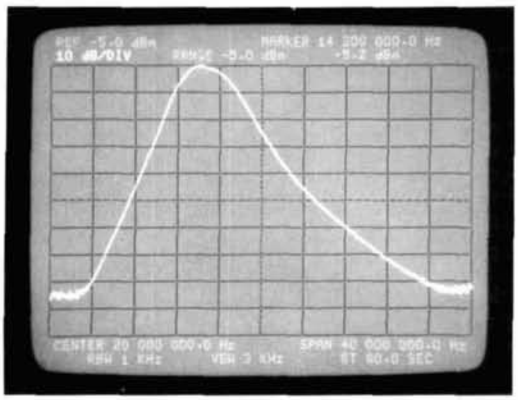


Photo A. Input bandpass filter.

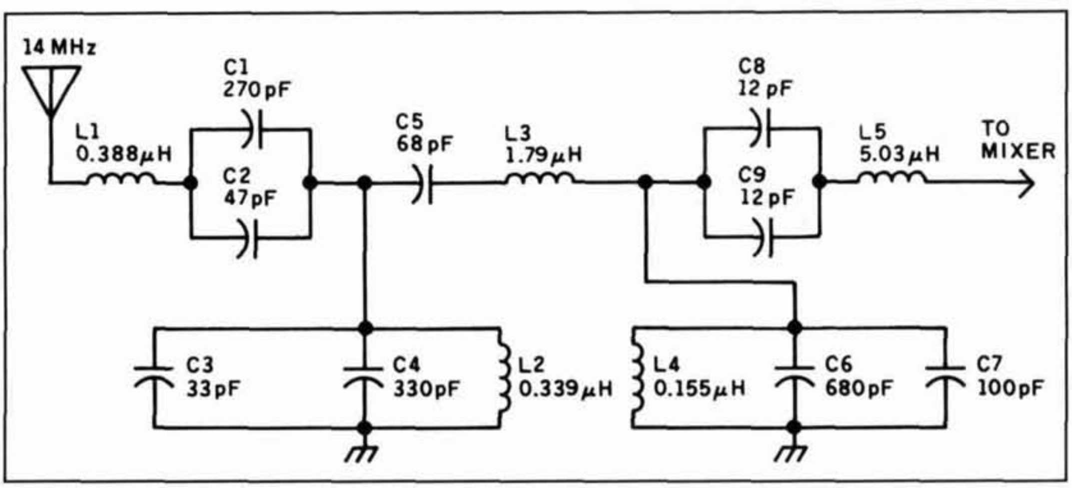


Figure 3. Fixed tuned input bandpass filter for 20 meters. If needed, alignment is best done using spectrum analyzer with tracking generator. A scope and sweep generator may also be used.

demodulator/balanced mixer applications). I tried both devices in a passive mixer configuration with excellent results. The Si8901 is my device of choice because of its easier local oscillator port drive requirements. Siliconix Application Note AN85-2[5] provides detailed information needed for the design. The functional block diagram of the Si8901 mixer is given in **Figure 4**. The design boils down to the proper termination of all three ports – signal, local oscillator, and IF – and the configuration of the gate drive.

The schematic diagram of the mixer is shown in **Figure 5**. Transmission line transformers match the source (BPF) to the signal port and the load (diplexer) to the IF port. The source and the load impedances are: Rs = Rl = 50 ohms.

Analysis of the equivalent circuit of a commutation mixer shows that there's a compromise between conversion loss and intermodulation distortion (IMD). The impedance transformation performed by transformers T1, T2, and T3 achieves a low level of IMD while keeping the conversion loss at an acceptable level. The performance of the commutation mixer depends greatly on the waveform at the local oscillator (LO) port. In the case of a sinusoidal waveform, very high levels of LO drive are required to ensure proper mixing action. It can be shown that the level of IMD products is a function of the rise and fall times of the LO waveform. Very high levels of LO voltage minimize the rise and fall time and, at least theoretically, an idealized square-wave drive should provide an infinite improvement in IMD performance. Therefore, the signal from the VFO is fed into a flip-flop (U1B) dividing the frequency by two and providing 50-percent duty cycle square waves at its complimentary outputs. Because the square waves aren't ideal and the MOSFET match within the package isn't perfect, an offset bias adjustment (R18) ensures that the switches operate in a 50-percent duty cycle. The input of the flip-flop is biased at the midpoint of the power supply voltage to obtain a waveform with a 50-percent duty cycle. The output of the second half of the flip-flop is connected to the digital frequency counter. Voltage regulator U4 provides a 10-volt supply for the flip-flop.

The third-order intercept measurement results are shown in **Photo B**. To find the output intercept point, use **Equation 1**:

$$IP3out = P + 1/2\,(IMR) \qquad \textbf{(1)}$$

where IP3out is the third-order output intercept point in dBm, P is the level of the fundamental frequency, and IMR is the ratio between the levels of the fundamental frequency and third-order products in dB.

Diplexer. Double-balanced mixers are sensitive to IF-port mismatches![10] A reactive load at the IF port can cause an increase in conversion loss, as well as degrade the third-order intercept response. I used a bandpass-type diplexer consisting of L3, C15, C16, L4, C17, C18, R20, and R21 (see **Figure 5**). The desired frequency (9 MHz) is passed

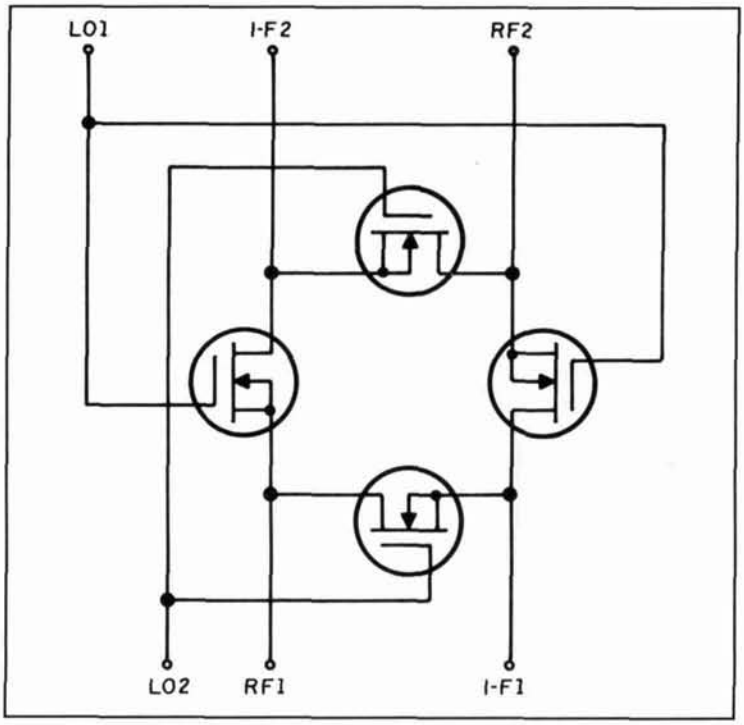


Figure 4. Functional block diagram of the Siliconix Si8901 passive commutative mixer.

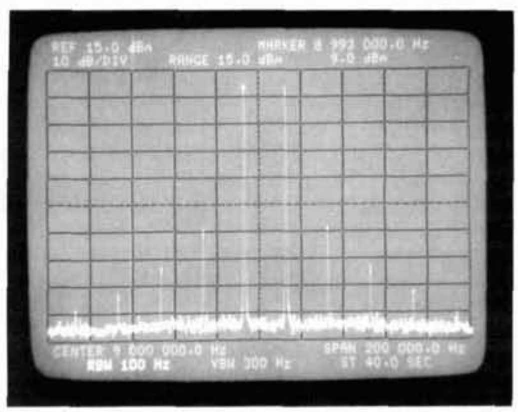


Photo B. Mixer third-order intercept point.
IP3out = P + 1/2 (IMR)
P (FUNDAMENTAL) = +9.0 dBm; IMD = −44 dBm
IMR = (P − IMD) dB = +9.0 − (−44) = 53 dB
IP3out = +9.0 + 1/2 (53) = +35 dBm
IF mixer loss Av = −8.0 dB,
IP3in = IP3out − Av = 35 − (−8.0) = + 43 dBm

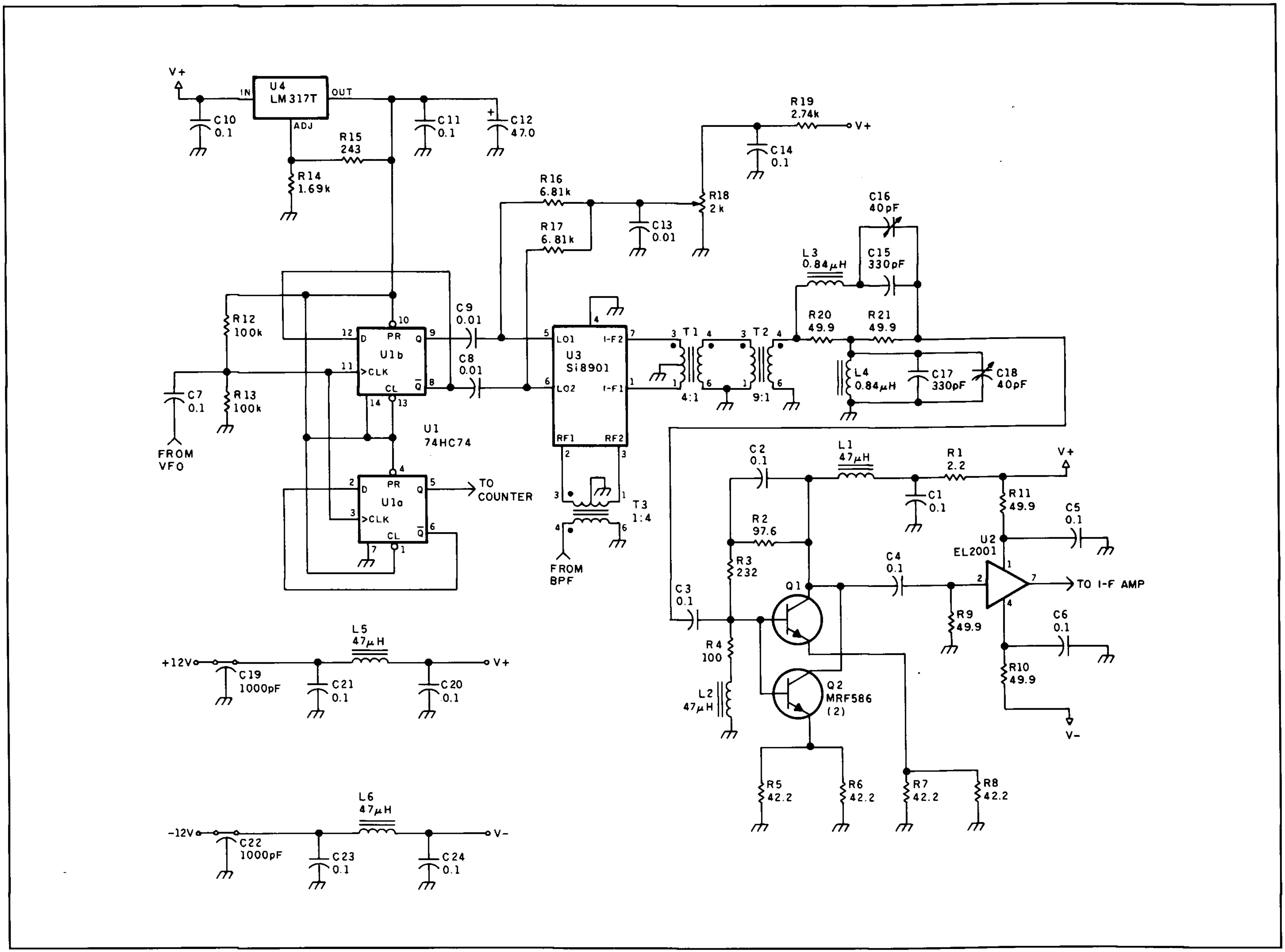


Figure 5. Mixer and post mixer amplifier. IC U1A and B divide the VFO frequency by two, reducing phase noise while providing an ideal squarewave LO injection for the mixer and a digital output for the display counter. Proper mixer termination at image and spurious frequencies is provided by a diplexer.

through the network with minimum attenuation, but the out-of-band signals are terminated over a considerable frequency range. Resistor R20 presents a 50-ohm impedance to the secondary winding of T2, while R21 presents a 50-ohm impedance to the input of the following stage.

Use of a diplexer provides an improved third-order intercept point over a resistive pad of 3 or 6 dB. This would add to the mixer loss.

Post-mixer amplifier. Because the mixer has conversion loss, it must be followed by a low-noise post amplifier to preserve the system noise figure. The criteria for the design is as follows: The third-order input intercept point of the amplifier should exceed the output intercept point of the mixer (35 dBm).

I chose a bipolar post-mixer amplifier with negative feedback. The third-order intercept requirements are met by biasing the transistors for high current. The amplifier is shown in **Figure 5** and uses a pair of MRF 586s biased for 90 mA each. The emitter resistors and biasing components set the gain — Av = 10.5 dB — and the third-order input intercept point — IP3in = 36.5 dBm (IP3out = 47 dBm). The third-order intercept measurement results are shown in **Photo C**. Detailed information on negative feedback amplifier design is given in **Reference 11**.

Because heavy negative feedback is employed in the amplifier, the input impedance is a function of the output impedance. If the post-mixer amplifier were followed directly by a crystal filter, any variation in the impedance of the filter would be reflected back through the amplifier to the IF port of the mixer because the diplexer is transparent at the IF frequency.

Instead of using the traditional resistive pad separating the post-mixer amplifier from the crystal filter, I decided to try a resistive termination (R9) which terminates the IF port of the mixer at all times, and a buffer (U2) which separates the highly reactive input impedance of the crystal filter from the post-mixer amplifier load.*

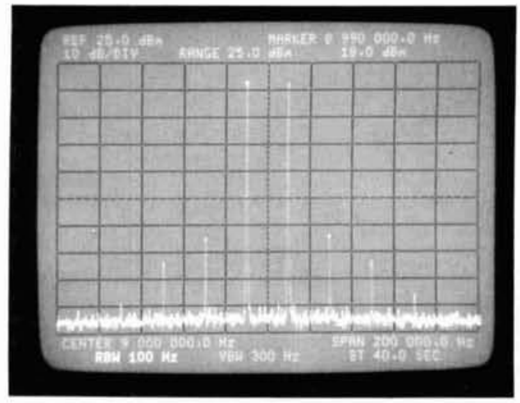


Photo C. Post-mixer amplifier third-order intercept point.
IP3out = P + 1/2 (IMR)
P (Fundamental) = +18 dBm; IMD = −40 dBm
IMR = (P − IMD) dB = +18 − (−40) = 58 dB
IP3out = +18 + 1/2 (58) = +47 dBm
IF amplifier gain Av = 10.5 dB,
IP3in = IP3out − Av = 47 − 10.5 = 36.5 dBm

Variable frequency oscillator (VFO). The VFO is another stage that requires much attention. A poorly designed oscillator may ruin an otherwise good receiver. A VFO with excessive phase noise may render any improvements in the dynamic range of the front end useless. Oscillator phase noise can be divided into two different regions:

- close-in phase noise (within the bandwidth of the SSB signal), and
- far-out phase noise (beyond the bandwidth).[12]

Excessively high levels of close-in phase noise may reduce a receiver's ability to separate closely spaced signals. In fact, a combination of two highly selective filters provides out-of-band rejection in excess of 100 dB at 2.0 kHz (the accuracy of this measurement is limited by the dynamic range of the spectrum analyzer). This would necessitate the use of a VFO with −133 dBc/Hz phase noise at 2.0 kHz offset, if the filter performance is not to be degraded.

A VFO with a high level of close-in phase noise may lead to a phenomenon called "reciprocal mixing." This phenomenon occurs when the noise sidebands of a VFO mix with a strong off-channel signal to produce an IF signal.

An excessively high level of far-out phase noise may degrade the dynamic range of the receiver. A signal entering a mixer is modulated by the phase noise of the VFO. At this point, the signal appears to have at least as much phase noise as the VFO. Thus, the noise floor is raised, leading to the degradation of the dynamic range.

*The author uses a conventional RF amplifier design for the post mixer which employs resistive feedback techniques for setting gain and impedance. These elements introduce additional noise. Because no RF amplifier is used in this design, the noise figure of the post mixer amplifier plays a large part in determining the receiver MDS floor. More technically inclined readers may wish to investigate the use of post mixer amplifier designs using transformer feedback techniques. These designs offer many advantages over ones which use only voltage and current feedback, including improved noise figures. For more information, you might like to read Ulrich Rohde's articles "Recent Developments in Circuits and Techniques for High-Frequency Communications Receivers," *Ham Radio*, April 1980, pages 20–25, and "High Dynamic Range Receiver Input States," *Ham Radio*, October 1975, pages 26–31. Ed.

The phase noise governed dynamic range can be found using **Equation 2:**[13]

$$PNDR = -Pn + 10 \log BWn \qquad (2)$$

where Pn is the VFO phase noise spectral density in dBc/Hz and BWn is the IF noise bandwidth in Hz. The dynamic range limited by phase noise should be at least equal to or better than the dynamic range limited by the IMD (SFDR) if the receiver dynamic range is not to be degraded. Thus, if the PNDR = SFDR, the required far-out phase noise level can be found from:

$$Pn = -SFDR - 10 \log BWn\ Pn = -113 - 33 = -146 \text{ dBc/Hz} \qquad (3)$$

To satisfy these requirements, I designed a high-quality LC-type VFO. I used a Hartley oscillator (described in detail in **References 11** and **14**). **Figure 6** shows the schematic.

I took extra care when building tank coil L1. While I recognize that iron-core inductors should be avoided in the interest of long-term stability, I made a compromise keeping the following considerations in mind:

- reduction of VFO noise sidebands is of primary concern (high Q),
- the VFO is part of a receiver only,
- a typical ham shack has a reasonably constant year-round temperature.

My objective in designing the coil was to obtain a high Q while maintaining acceptable temperature stability. Amidon no. 6 is the most stable powered-iron material for the HF band.[15] Theoretically, it should be possible to obtain a Q over 300 using proper cores and wire. Twenty-two turns of no. 16 AWG on a T80-6 core provided a Q of 285. I took precautions to ensure adequate long-term drift. Low-noise JFET Q1 is biased so the drain current is almost independent of the ambient temperature. Oscillator Q1 and buffer Q2 are powered from a regulated power supply (+6.5 volts). Capacitors in parallel (C3, C4, C5, and C6) are used instead of a single capacitor to reduce component heating by AC currents.[16] I recom-

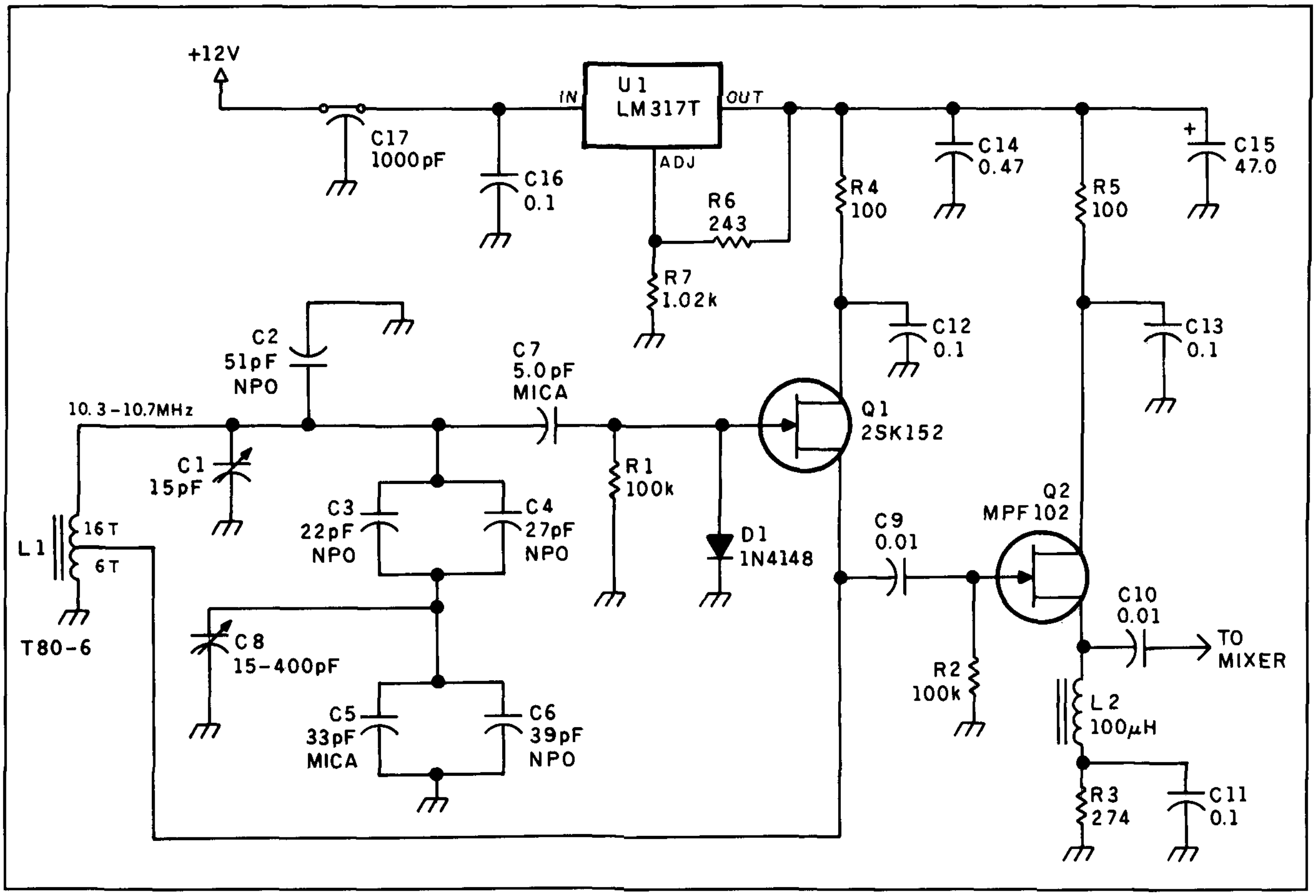


Figure 6. A 10-MHz VFO provides tuning range across the 20-meter voice band. Positive temperature coefficient and NPO capacitors combine to reduce thermal drift. Interstage coupling reduces oscillator and buffer phase noise. Note that the VFO frequency is divided by two (*Figure 5*) reducing phase noise and halving drift.

mend you use zero-temperature coefficient (NPO) ceramic capacitors for the VFO, although one or more capacitors should have a negative drift (silver mica, polystyrene) to compensate for the positive temperature drift of the magnetic core. Very light coupling between the tank and the JFET through capacitor C7 preserves the Q of the tank circuit. I avoided the traditional light coupling between the oscillator and the buffer (C9) because it could lead to an increase in VFO phase noise.[3] Use a good quality variable capacitor for C8, preferably a double-bearing type.

The frequency of the signal at the buffer output (Q2) is between 10.3 and 10.7 MHz. The buffer is followed by flip-flop U1B (**Figure 5**) which divides the frequency by two. Several objectives are met by this frequency division:

- A 50-percent duty cycle square wave ensures the optimum performance of the Si8901 commutative mixer.
- Frequency division reduces the VFO noise sidebands.
- The phase noise of the signal at the output of the frequency divider is improved if the magnitude of the signal at the output of the divider exceeds the magnitude of the signal at the input of the divider. Indeed, the "dBc" in the phase noise units of "dBc/Hz" means that you are measuring the power of noise relative to the power of the carrier. Any increase in carrier power would lead to phase noise improvement (providing that the noise level hasn't increased).

Photo D shows the VFO signal spectrum. The close-in phase noise measured at 2.0-kHz offset from the carrier is −140 dBc/Hz and the far-out phase noise measured at 50-kHz offset is −146 dBc/Hz, satisfying the noise level requirements.

IF amplifier. Although the noise figure of the IF amplifier isn't as critical as it is in the front end of the receiver, I decided to build an amplifier using discrete MOSFET transistors. The amplifier schematic diagram is shown in **Figure 7**. The four stages of amplification provide gain in excess of 90 dB, AGC range over 100 dB, and a noise figure below 5 dB. Crystal filters BPF-1 and BPF-2 provide the IF selectivity. BPF-1, an eight-pole crystal filter has a −3 dB bandwidth of 2.2 kHz and provides most of the selectivity. The second filter, BPF-2, is also an eight-pole filter with a −3 dB bandwidth of 2.4 kHz intended to reduce the wideband noise generated by the IF amplifier and improve the overall IF selectivity. This filter need not be as exotic as the one used up front, but it is desirable to match the center frequency.

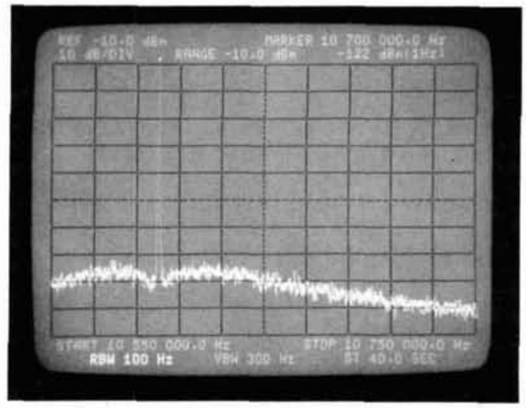


Photo D. VFO spectrum.
VFO frequency: Fvfo = 10.6 MHz
VFO spectral noise density at:

4 kHz from carrier −116 dBm (1 Hz)
20 kHz from carrier −111 dBm (1 Hz)
100 kHz from carrier −122 dBm (1 Hz)

Phase noise at the divider output:*
2 kHz from carrier −140 dBc (1 Hz)
10 kHz from carrier −135 dBc (1 Hz)
50 kHz from carrier −146 dBc (1 Hz)
***Note: Phase noise at the divider output in "dBc (1Hz)" is the sum of the spectral noise density in "dBm (1 Hz)" at the divider input and the divider output voltage swing in "dBm" (10 volts p-p or +24 dBm).**

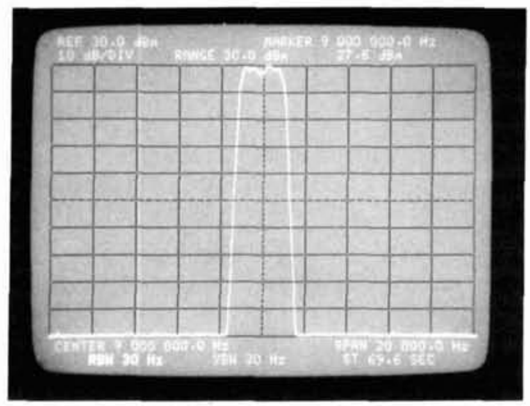


Photo E. IF amplifier band-pass filters.

The frequency response of two cascaded crystal filters is shown in **Photo E**. Band-pass filters BPF-1 and BPF-2 have 500-ohm input and output terminations (R1, R2, R14, and R27). Variable coils L1, L2, and L3 provide adequate voltage swing and additional selectivity. Resistors R3, R9, and R18 limit the gain of the stages. U1 serves as a buffer stage between amplifier Q4 and the product detector. Resistor R24 helps discourage parasitic oscillations. Resistor R23 sets the overall gain of the IF strip. The stages use

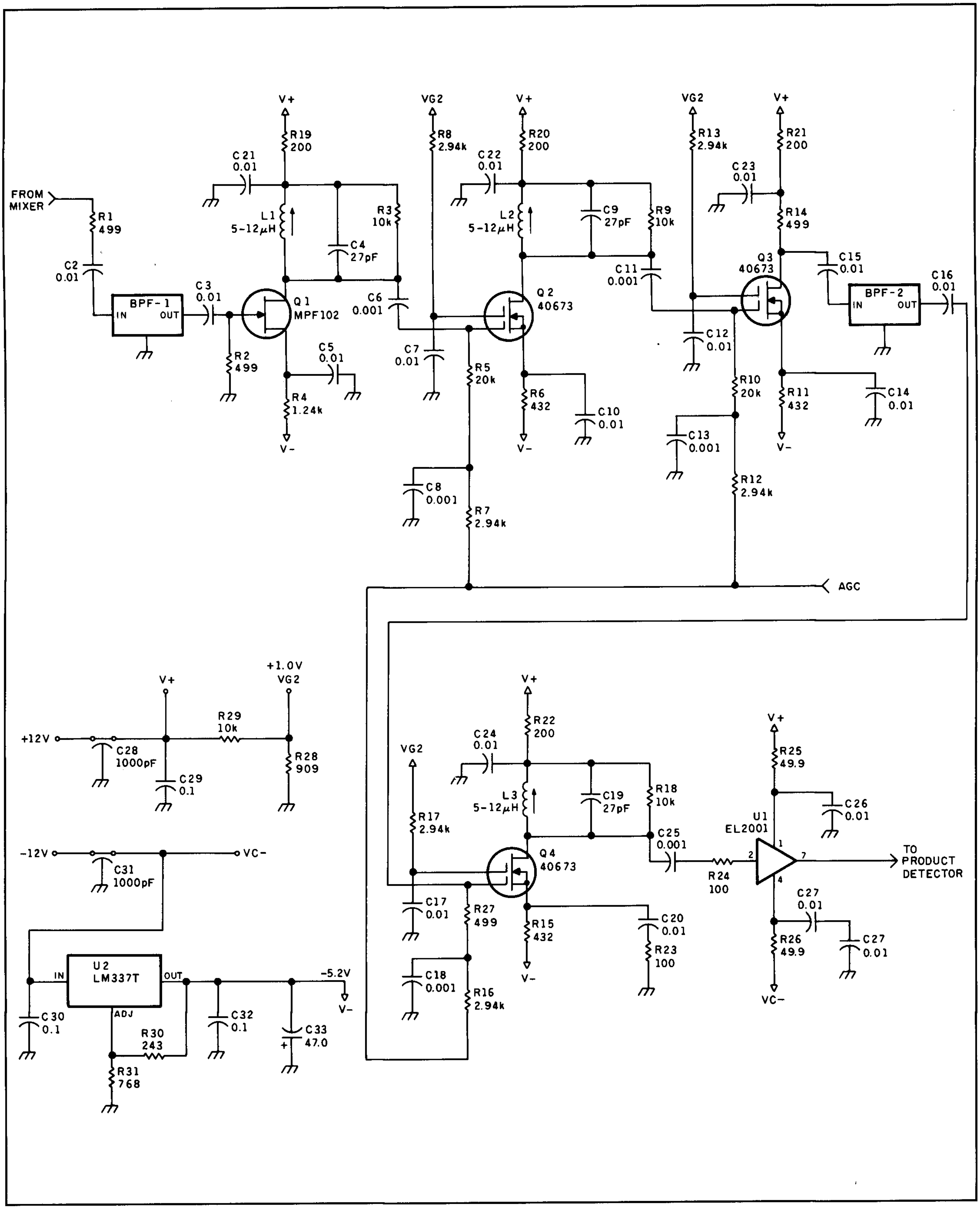


Figure 7. The IF amplifier uses discrete transistors. Note the post IF crystal filter to prevent broadband noise generated in the IF from reaching the product detector. This reduces fatiguing high frequency hiss.

dual power supply voltages V+ = 12 volts and V− = −5.2 volts to obtain the required AGC control range. Gain control is accomplished by varying gate 1's voltage while keeping gate 2's voltage constant and low (Vg2 = 1.0 volt). This departure from the traditional method of varying gate 2's voltage provides better gain versus control voltage linearity and less signal distortion.[3]

Product detector. Most of the requirements applicable to a mixer can be relaxed when considering a product detector circuit. However, IMD is of some concern because it's possible for signals within the IF passband to produce spurious products. Therefore, I decided to use a low distortion passive double-balanced mixer. The schematic diagram of the product detector is shown in **Figure 8**.

The TAK-3H is a Mini-Circuits very low distortion mixer requiring +17 dBm LO drive. Components L6, C14, and R10 serve as a low-pass filter for the audio signal and a 50-ohm termination for high frequencies. Transistor Q3 and associated components serve as a buffer between the RF port of the product detector and the AGC circuit.

BFO. The schematic diagram of the BFO is shown in **Figure 8**. The oscillator is crystal controlled. Trimmer capacitor C8 provides a couple of kilohertz of frequency pulling. Zener diode D1 regulates the power supply voltage of the first stage. The second stage, Q2, is an amplifier and provides +17 dBm of LO drive to the product detector. The low-pass filter (composed of L7, C16, L8, and C17) serves to suppress the harmonics and reduce the possibility of harmonic mixing in the product detector. The filter also ensures that the waveform at the LO port is symmetrical, as required for best balance. Variable capacitor C7 sets the level of the LO drive.

Audio amplifier. Figure 9 is a schematic diagram of the audio amplifier. The signal from the product detector is fed into amplifier U1A. A six-pole Chebyshev low-pass filter (LPF) with a corner frequency of 2.5 kHz reduces high frequency noise and enhances the receiver's overall signal-to-noise characteristic. The LPF is comprised of integrated circuits U1B, U2A, U2B, and associated components. **Reference 17** gives the LPF design data.

A three-pole Chebyshev high-pass filter (HPF) with a corner frequency of 250 Hz helps reduce the low frequency rumble and 60-Hz hum. The HPF is composed of integrated circuit U2A and associated components. The HPF design data is discussed in **Reference 18**.

U4 is a 10-watt audio power amplifier, chosen to reduce the harmonic distortion level. Because it has more audio power capability than is required, the amplifier can operate in its lowest distortion region. When used in a typical application circuit,[19] U4 provides 30 dB of voltage gain and can drive a 4 or 8-ohm speaker with a distortion level of 0.2 percent (typical). Potentiometer R13 serves as a volume control.

The frequency response of the audio amplifier is shown in **Photo F**.

Automatic gain control (AGC). I wanted the AGC loop to follow rapid variations in signal levels without "clicking" or "pumping." To achieve this goal, I used a modified full hang AGC system. The schematic diagram of the AGC circuit is shown in **Figure 10**. The AGC is IF derived.

After being buffered by Q3 (**Figure 8**), the IF signal is applied to RF amplifier U1 (**Figure 10**) which is configured as an inverting amplifier. Gain control R1 serves as an AGC voltage range control. Diode D1 serves as an AGC detector. In a departure from the traditional full hang AGC circuit, I replaced a single memory capacitor with two main memory capacitors (C5 and C6). Capacitor C6 of the first RC network (R4-C6) gets charged by rapid signal changes, while capacitor C5 of the second RC network (R3-C5) reacts only to relatively long signal changes. This dual capacitor system allows a fast attack time without an audio click. Diodes D2, D3, D4 and D5 decouple the two networks.

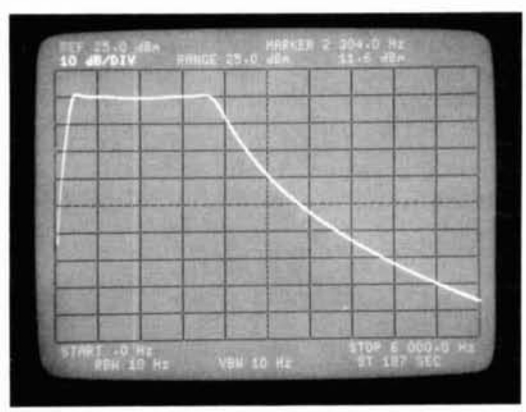


Photo F. Audio amplifier.

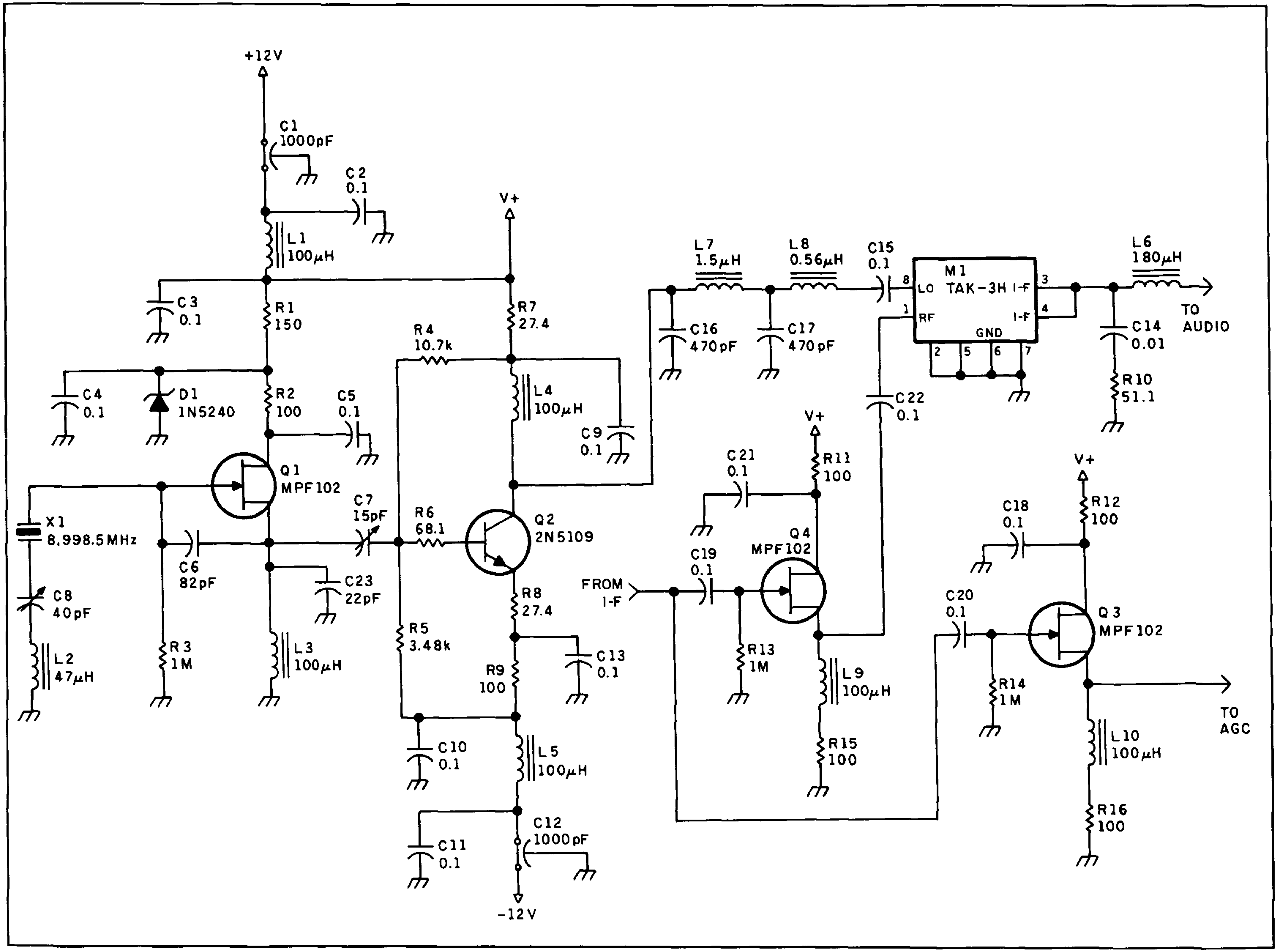


Figure 8. The BFO and product detector. A high level double-balanced diode mixer is used for the product detector, assuring excellent linearity.

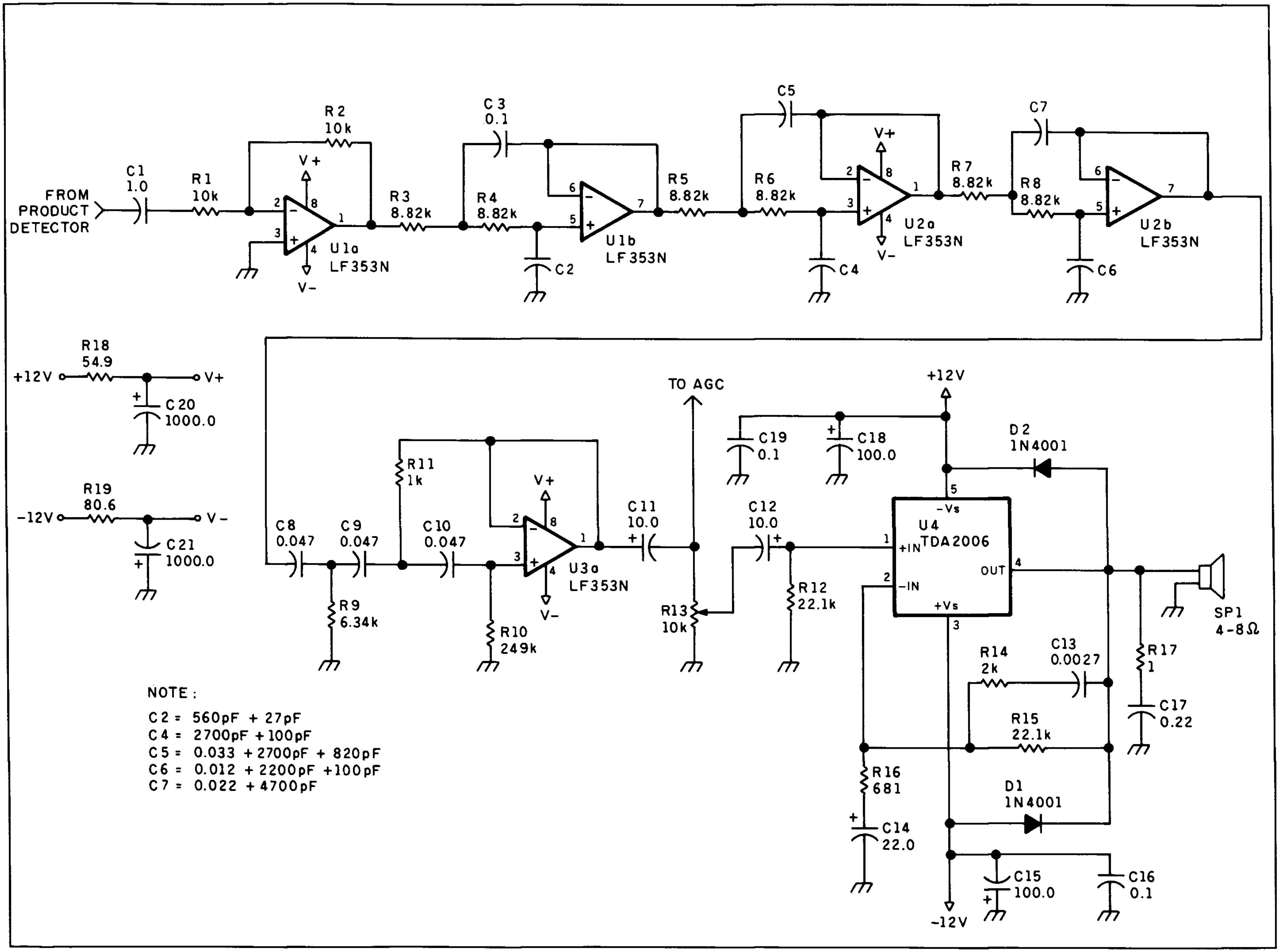


Figure 9. The audio amplifier. A six-pole lowpass filter tailors the audio to improve the signal-to-noise characteristics of the receiver and reduce high frequency noise. The 10-watt audio IC amplifier ensures excellent linearity at normal listening levels.

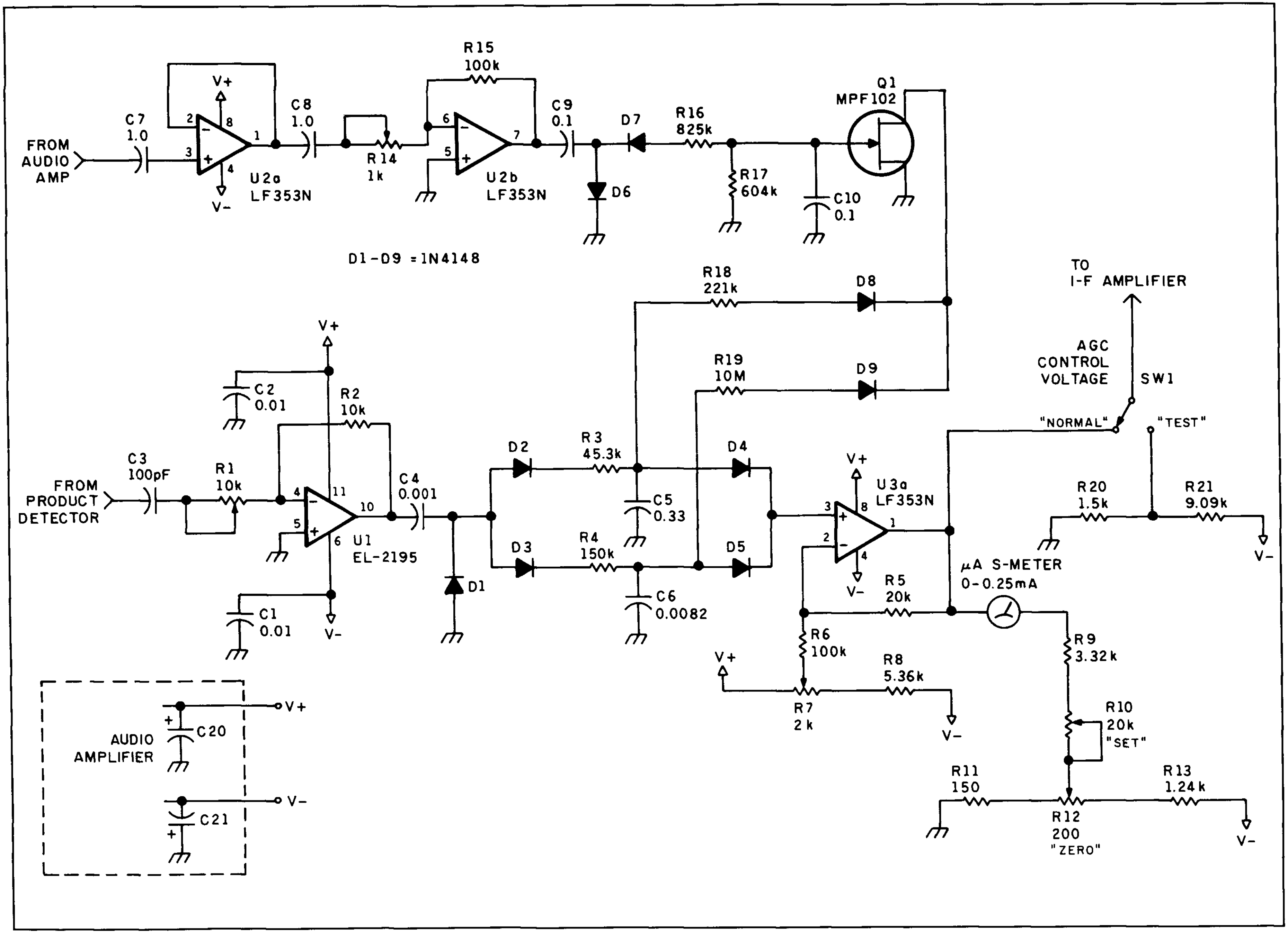


Figure 10. The AGC stages. AGC performance is often overlooked by many receiver designers, but its performance can make or break a receiver design. A unique dual hang time constant is used to reduce popping and clicks on the recovered audio. AGC is developed from the IF RF envelope, not the product detector audio, further enhancing AGC performance.

The audio signal reaches capacitor C10 after being buffered by U2A, amplified by U2B, and rectified by D6 and D7. Weak and rapidly changing signals have little effect on capacitor C10. As a result, the main memory capacitor (C6) will charge quickly, cause a momentary reduction of the IF gain, and discharge through R19 and the drain of Q1. On the other hand, sustained signals charge up capacitor C10. Potentiometer R14 sets the gain of U2B so the negative voltage developed on capacitor C10 pinches off JFET Q1. Main memory capacitors C5 and C6 remain charged as long as the gate of Q1 stays negative. When the signal disappears, C10 starts to discharge through R17, and when Q1 is no longer pinched off, C5 and C6 discharge quickly through Q1. Potentiometer R7 sets the AGC threshold, while potentiometers R10 and R12 set the S-meter range and "zero."

Power supply. I used triple power supply HBAA-40W from POWER-ONE Inc. in this project.

The receiver power supply requirements are:

+12 volts, 600 mA
−12 volts, 180 mA
+5 volts, 700 mA

Digital display. I chose a six-digit LED readout similar to the one described in **Reference 11** to display the frequency.*

Construction

The receiver is housed in a 17 by 13 by 5-inch aluminum chassis box from LMB. The digital readout window, tuning knob, volume control, ON-OFF switch, S-meter, and audio jack are mounted on the cabinet's front panel. The antenna connector, AC power socket, ground terminal, and fuse holder are mounted on the rear panel. I drilled a number of ventilation holes in the chassis cover to lower the temperature inside the cabinet and minimize VFO frequency drift. The modular construction helps improve shielding and allows for easy module replacement for repair or future experimentation.

Five modules (bandpass filter, mixer, VFO, IF amplifier, BFO, and product detector) are enclosed in individual aluminum project boxes and mounted on the bottom of the main chassis box. The two remaining boards, which contain the digital display and audio/AGC circuitry, aren't enclosed, but are fastened to the bottom of the chassis box with brackets.

The components within each of the five modules are mounted on pieces of Vector board (Vector part no. 8007). The ground plane is on the component side and all ground connections are made directly to the upper surface of the board. The components can be placed directly in the holes or may be mounted on Vector pins (Vector part no. T68).

Observe standard RF circuit building methods; i.e., short leads, proper placement of bypass capacitors, and point-to-point wiring. All signal inputs and outputs are terminated with BNC connectors. Connections between modules are made with RG-174 coaxial cable using BNC connectors. DC voltages are supplied to the modules via feedthrough capacitors mounted on the side panels of the module boxes. Four holes are drilled at the corners of each Vector board. Stand-offs are placed between the boards and the module bottoms, and four screws fasten the boards and the modules to the main chassis bottom. If painted project boxes are used for the modules, remove all paint from the bottom and the edges to ensure good electrical contact between sections of the shield.

The audio/AGC board uses a piece of a Vector board mounted vertically to save space. The digital display board is also mounted vertically, and is the only board where wire-wrapping is acceptable. The connections from the display board to the LED module are made via wire-wrap insulated wires bunched together with plastic tie wraps. The LED display module is an assembly of six seven-segment LED modules mounted on a piece of a Vector board. The display module is attached to the front panel with two screws and covered with a rectangular "window" made out of red anti-glare plastic material glued to the front panel.

Pay special attention when building the VFO module. Rigid construction assures frequency stability. The tank coil is covered with a low-loss polystyrene Q-dope and placed as far as is practically possible from the walls of the enclosure. Mount the frequency determining capacitors on Vector pins, as component selection will be required to compensate for the VFO frequency drift. Keeping this in mind, make sure the cover of the VFO module is easy to remove, providing access to the components. I used a surplus 25:1 reduction drive to couple the shaft of the variable capacitor to the turning knob.

*The primary focus of this article is optimum receiver dynamic range. This cursory display scheme only samples the VFO operating frequency. For utmost accuracy, it's necessary to sample and count the VFO, BFO, and (if present) heterodyne oscillators to produce a true composite additive count which represents the actual receiver carrier frequency. Ed.

Alignment

Input bandpass filter. If the inductances are within ±5 percent of their nominal value, no alignment is required. If not, I recommend value "tweaking" using a spectrum or impedance analyzer. You can perform a crude frequency response check with a scope and a signal generator.

Mixer. Diplexer trimmer capacitors C16 and C18 can be aligned by injecting a signal into the antenna connector (14.250 MHz), tuning the receiver to that frequency, and peaking both trimmer capacitors while observing the signal with a scope at the output of buffer U2. Potentiometer R18 is included to ensure that the mixer switches operate in a 50-percent duty cycle mode and provide the best suppression of third-order IMD products. The best alignment results can be achieved using a spectrum analyzer. A test setup for measuring the two-tone IMD[1,4,9,11,20] is required. With the spectrum analyzer at the output of buffer U2 (**Figure 5**), adjust R18 to bring the third-order IMD products to a minimum. If you don't have access to a spectrum analyzer, I recommend you use an alternative method. Using the same setup, connect a scope to the output of the IF amplifier (pin 8 of U1 in **Figure 7**). Disable the AGC circuit by switching SW1 to the TEST position. Calculate the frequency of a third-order product (2F1 – F2 or 2F2 – F1) and tune the receiver to display this new frequency on the scope. Adjust R18 to obtain the minimum value of this third-order IMD product.

VFO. Before applying the Q-dope to the oscillator coil, check the VFO tuning range. With C1 in midposition (**Figure 6**), measure the frequency at the output of the VFO module with a frequency counter while moving the plates of C8 from one extreme position to the other. For the SSB portion of the band, you should cover a range from 10.3 to 10.7 MHz. You can shift the frequency range by squeezing the turns of L1 together or spreading them apart. Trimmer capacitor C1 provides fine tuning.

A frequency meter, time, and patience are needed to optimize the short and long-term frequency drift of the VFO. Along with several components, capacitors C2, C3, C4, C5, and C6 have an effect on the amount and direction of the frequency drift. Select these components with an eye to obtaining an acceptable VFO stability. While NPO capacitors may provide a good starting point, one or more of the capacitors must have negative temperature drift to compensate for the positive temperature drift of the L1 core. Try silver mica and polystyrene capacitors. A good supply of small value capacitors is a must. Mica capacitors, even those from the same batch, may exhibit different temperature drift properties. Make your capacitor selection under normal operating conditions. The VFO module should be enclosed and the housing covered. Allow sufficient time for the solder joints to reach their normal temperature, and for the temperature within the housing to settle to its nominal value. You can make a reasonable judgement on the duration of the measurement time by observing the frequency counter display. You'll find that the frequency drift slows down considerably at some point.

IF amplifier. Peak coils L1, L2, and L3 while injecting a signal into the antenna connector (14.25 MHz). Tune the receiver to the frequency and observe the signal with a scope at the output of buffer U1 (**Figure 7**). The AGC circuit should be disabled with switch SW1 in TEST position.

The gain from the input to the output of the amplifier should be 92 ±3 db. You can change the value of resistor R23 to obtain the nominal gain. At 92 db of gain, the noise level at the output (with the signal generator off) should be approximately 30 mV p-p. You'll need a good quality calibrated signal attenuator to make the gain measurement.

BFO, product detector. A rule of thumb for setting the BFO frequency is to place it approximately 20 dB down the slope of the IF passband curve. An experienced circuit builder can set the BFO frequency quite accurately by listening to the recovered audio signal. Trimmer capacitor C8 sets the BFO frequency (**Figure 8**).

Trimmer capacitor C7 sets the +17 dBm LO level. With a scope connected to pin 8 of the mixer, adjust C7 to establish 4.5 volts p-p of LO drive.

AGC. The AGC board needs a number of adjustments, which should be done in proper sequence. First, connect a scope to pin 1 of U3A (**Figure 10**) – the output to IF.

Step 1. Disconnect the input (product detector) and set the output voltage to –1.4 volts with "threshold" potentiometer R6.

Step 2. Set the "AGC loop gain" potentiometer R1 equal to 390 ohms.

Step 3. Connect a signal generator via an attenuator to the antenna connector. Reconnect the product detector to the AGC board. Set the frequency to 14.25 MHz and the signal level to 50 mV rms. Set the attenuator to 60 dB. The signal is equal in strength to an S9 on the S-meter. Tune the receiver to this frequency. Potentiometer R14 adjusts the level of the audio signal which is applied to the gate of JFET Q1 after being rectified. If the level is too low, the gate voltage isn't sufficiently negative to pinch off Q1. Consequently, main memory capacitors C5 and C6 will keep discharging through Q1, even if strong signals are present at the antenna terminal. If the level is too high, the gate voltage is too negative, and Q1 is pinched off all the time. With the signal present at the input of the receiver, the AGC output voltage should move to a level more positive than the nominal −1.4 volts. Set potentiometer R14 to the maximum gain position. Turn off the signal generator while monitoring the output signal level on the scope. Reduce the gain of the amplifier by turning R14 back until you reach a point where the output voltage drops to a nominal −1.4 volts. Turn R14 an additional 5 degrees.

Step 4. Disconnect the signal generator. Connect the antenna to the input terminal. Perform this adjustment when you can receive strong local stations with signal levels no lower than S9 + 40 dB. Adjust potentiometer R1 so an audio signal from a strong local station isn't excessively loud (doesn't distort) on one hand, and doesn't produce audio clicks (too much loop gain) on the other. It may take you several tries to find the optimal gain setting.

Step 5. Reconnect the signal generator to the antenna connector. With the signal generator turned off, adjust ZERO potentiometer R12 so the S-meter needle is at the "0" position.

Step 6. Turn the signal generator on. With a 50-mV rms signal and the attenuator set to 60 dB, adjust "set" potentiometer R10 so the S-meter needle indicates S9. The attenuator is now a convenient tool for checking the S-meter scale. Reducing the attenuation by 20 dB should produce an S9 + 20 dB reading, and so on. If the reading differs considerably, consider replacing the existing S-meter scale with a new one.

Summary

One of my objectives in producing this design was to prove that a wide dynamic range receiver could be constructed using basic design principles. This is by no means a weekend project and should be undertaken only by builders willing to spend a great deal of time on construction and measurements. While the availability of laboratory-quality test equipment is desirable, it is possible to perform all necessary measurements and adjustments with equipment available to most construction-minded Amateurs. The design and modular construction leaves room for individual experimentation and simplification.

It's really gratifying when I can switch from my state-of-the-art commercially produced transceiver to this home-built receiver and continue a QSO when operating under adverse receiving conditions.

Acknowledgements

I'd like to acknowledge the contributions made by my professional colleagues. Technical assistance and encouragement was provided by Paul Rumbaugh, Terry Flach, WU6N and Ed Oxner, KB6OJ. ■

Appendix

	BPF	Mixer	Post Mixer Amp	BPF-1	Main IF Amp
Noise Figure (NF) (dB)		8.5	5	3	5
Noise Figure (NFt)		7.1	3.16	2.0	3.16
Gain (Av) (dB)		−8.5	+10.5	−3.0	-
Power Ratio (G)		7.1	11.22	2.0	-

The system noise factor (NFt)

NFt = N1 + (N2−1)/G1 + (N3−1)/G1×G2 + (N4−1)/G1×G2×G3

NFt = 7.1 + (3.16−1)/(1/7.1) + (2.0−1)/(1/7.1×11.22) + (3.16−1)/(1/7.1×11.22×1/2)

NFt = 31.5

The system noise figure (NF)

NF = 15 dB

The minimum discernable signal (MDS)

MDS = −174 dBm + NF + 10 log BWn, where BWn is the noise bandwidth of the receiver for BWn = 2200 Hz, 10 log BWn = 33 dB

MDS = −174 dBm + 15 dB + 33 dB = −126 dBm

Sensitivity at 10 dB S/N (Ps)

Ps = 10 dB + MDS = 10 − 126 = −116 dBm,

or 0.35 μV rms at the receiver input loaded with a 50-ohm antenna.

The upper limit of the dynamic range (Pu)

Pu = 1/3 (MDS + 2 IP3in) = 1/3 (−126 + 2 × 43.5) = −13 dBm **(a)**

The spurious free dynamic range (SFDR) **(b)**

$$SFDR = Pu - MDS = -13 - (-126) = 113 \text{ db}$$

The blocking dynamic range (BDR) **(c)**

$$BDR = Pd - MDS = 22 - (-126) = 148 \text{ dB} \quad \textbf{(d)}$$

Notes:

(a) IP3in: third-order input intercept point of the receiver.
IP3in = IP3in of the mixer + INPUT BPF loss = 43 + 0.5 = 43.5 dBm

(b) Sometimes referred to as: "IM dynamic range" or "two-tone dynamic range."

(c) Sometimes referred to as: "single-tone dynamic range." Both SFDR and BDR were measured at the ARRL Lab standard signal spacing of 20 kHz.

(d) Pd: dB desensitization point in dBm (see **Figure 2.**

REFERENCES

1. Cornell Drentea, WB3JZO, *Radio Communication Receivers*, Tab Books Inc., Blue Ridge Summit, 1982, pages 88-135.
2. A. Zverev, *Handbook of Filter Synthesis*, John Wiley, New York, 1967.
3. V. Drozdov, RA3AO, *Amateur HF Transceivers (Lyubitel skie kv Transiveri)*, in Russian, Moscow, USSR, 1988.
4. Doug DeMaw, W1FB, *Practical RF Design Manual*, Prentice-Hall Inc., Englewood Cliffs, New Jersey, 1982, pages 23-25.
5. Ed Oxner, KB6OJ, "Designing a Super-High Dynamic Range Double-Balanced Mixer," Application Note AN 85-2, Siliconix, Santa Clara, California, 1986.
6. Ed Oxner, KB6OJ, "Junction FETs in Active Double-Balanced Mixers," Application Note AN 73-4, Siliconix, Santa Clara, California, 1973.
7. Dr. Ulrich L. Rohde, DJ2LR, "Performance Capability of Active Mixers," *Ham Radio*, March 1982, pages 30-35, April 1982, pages 38-44.
8. *RF/IF Signal Processing Handbook* Volume 1, Mini-Circuits, Brooklyn, New York, 1985, pages 7-23.
9. *Wideband Signal Processing Component Catalog*, VARI-L Company, Denver, Colorado.
10. P.E. Drexler, "Effect of Termination Mismatches on Double-Balanced Mixers," *Microwave Journal*, January 1986, pages 187-190.
11. Wes Hayward, W7ZOI, and Doug DeMaw, W1FB, *Solid State Design for the Radio Amateur*, ARRL, Newington, Connecticut, 1977, Chapter 6.
12. John Grebenkemper, KI6WX, "Phase Noise and Its Effects on Amateur Communications," *QST*, March 1988, pages 14-20, April 1988, pages 22-25.
13. P. Chadwick, "Phase Noise Intermodulation and Dynamic Range," *Frequency Dividers and Synthesizers IC Handbook*, Plessey Semiconductors, 1988.
14. *The ARRL 1986 Handbook*, ARRL, Newington, Connecticut, 1985, Chapter 12.
15. "Iron-Powder and Ferrite Coil Forms" *Design Catalog*, Amidon Associates, Inc., North Hollywood, California, 1984, pages 1-8.
16. Doug DeMaw, W1FB, "The Principles and Building of SSB Gear – Part 4," *QST*, December 1985, pages 37-40.
17. F. Al-Nasser, "Tables Speed Design of Low-pass Active Filters," *EDN*, March 15, 1971, pages 23-32.
18. H.R. Wilbur, "Filter Gain is Adjustable Using a Simple Resistance Ratio," *Electronic Design*, November 8, 1979, page 96.
19. *Linear Integrated Circuits*, Databook, SGS-ATES, 1981.
20. R. Watson, "Guidelines for Receiver Analysis," *Microwaves and RF*, December 1986, pages 113-122.

Bibliography

1. N. Ayrandjian, "Simple Computation of Spurious-Free Dynamic Range," *RF Design*, January 1987, pages 65-66.
2. William Richardson, W3IMG, "IMD and Intercept points of Cascaded Stages," *Ham Radio*, November 1984, pages 28-34.
3. Joe Reisert, W1JR, "High Dynamic Range Receivers," *Ham Radio*, November 1984, pages 97-105.
4. James W. Healy, NJ2L, "ICOM IC-781," product review, *QST*, January 1990, pages 39-43.
5. Frank Noble, W3MT, "A Variable Frequency Crystal Oscillator," *QST*, March 1981, pages 34-35.

Parts Lists

Input Bandpass Filter

Reference no.	Component	Supplier
C1-C9	*dipped mica, polyester film, ceramic epoxy coated 5 percent or better tolerance*	*1,2,3*
L1	*8 turns*	
L2	*7 turns core: T80-6*	*5*
L3	*19 turns*	
L4	*4 turns wire: no. 20 enameled*	*1*
L5	*32 turns*	

Note: a) Capacitors can be picked from a ±10 percent tolerance stock with a capacitance meter.
b) If an impedance analyzer is available, the exact inductance values can be obtained by either squeezing the turns together or spreading them apart.

Mixer

Reference no.	Component	Supplier
U1	*74HC74 CMOS flip-flop*	*1,2,3*
U2	*EL2001CN buffer from Elantec*	*11,12,13,14*
U3	*Si8901A mixer from Siliconix*	*11,12,13,14*
U4	*LM317T adjustable voltage regulator*	*1,2,3,7,8*
Q1,Q2	*MRF586 RF transistor from Motorola*	*12,13,14*
T1,T3	*T4-1 wideband RF transformer from Mini-Circuits*	*6*
T2	*T9-1 wideband RF transformer from Mini-Circuits*	*6*
C15,C17	*330 pF ±5 percent dipped mica, polyester film, ceramic epoxy coated*	*1,2,3*
C16,C18	*7-40 pF trimmer capacitor*	*1,3*
C19,C22	*1000-pF feedthrough ceramic capacitor*	*1*
L1,L2,L5, L6	*47-µH RF choke*	*1,2,4,8*
L3,L4	*12 turns, core: T80-6, wire: no. 16 enameled*	*1,5*
R2	*97.6-ohm ±1 percent 0.25-watt resistor*	*1,2,3*
R3	*232-ohm ±1 percent 0.25-watt resistor*	*1,2,3*
R5,R6,R7 R8	*42.2-ohm ±1 percent 0.25-watt resistor*	*1,2,3*
R18	*2-k multi-turn potentiometer*	*1,2,3,4*
Q1,Q2	*press-on heatsink from Thermalloy 2228B series*	*1*

VFO

Reference no.	Component	Supplier
U1	*LM317T adjustable voltage regulator*	*1,2,3,7,8*
Q1	*2SK152-1 JFET transistor from Sony*	*11*
Q2	*MPF102 JFET transistor*	*1,7,8*
D1	*1N4148 or 1N914 small signal diode*	*1,3,4,7*
L1	*22 turns total, core: T80-6*	*5*
	6 turns from cold end, wire: no. 16 enameled	*1*
L2	*100-µH RF choke*	*1,2,8*
C1	*3-15 pF trimmer capacitor*	*1,3*
C2	*51 pF ±10 percent ceramic disc NPO*	*3,8*
C3	*22 pF ±10 percent ceramic disc NPO*	*3,8*
C4	*27 pF ±10 percent ceramic disc NPO*	*3,8*
C5	*33 pF ±10 percent dipped mica*	*1,3,8*
C6	*39 pF ±10 percent ceramic disc NPO*	*3,8*
C7	*5.0 pF ±10 percent dipped mica*	*1,3*
C8	*15-400 pF air variable capacitor, double bearing*	*surplus*
C17	*1000-pF feedthrough ceramic capacitor*	*1*

IF Amplifier

Reference no.	Component	Supplier
U1	*EL2001CN buffer from Elantec*	*11,12,13,14*
U2	*LM337T adjustable voltage regulator*	*1,2*
Q1	*MPF102 JFET transistor*	*1,7,8*
Q2,Q3,Q4	*40673 dual-gate MOSFET*	*1,8*
BPF-1	*crystal filter from Fox-Tango Corp., part no. 2309 BW = 2.1 kHz, Zin = Zout = 500 ohm*	*9*
BPF-2	*crystal filter from Fox-Tango Corp. part no. 231 BW = 2.4 kHz, Zin = Zout = 500 ohm*	*9*
L1,L2,L3	*5-12 μH variable inductance, Toko part no. 154, ANS-T1008Z*	*2*
C4,C9,C19	*27 pF ±10 percent dipped mica, polyester film, ceramic epoxy coated*	*1,2,3*
C28,C31	*1000-pF feedthrough ceramic capacitor*	*1*

BFO, Product Detector

Reference no.	Component	Supplier
M1	*TAK-3H frequency mixer from Mini-Circuits*	*6*
X1	*8998.5-kHz crystal part no. XF 901*	*1*
Q1,Q3	*MPF102 JFET transistor*	*1,7,8*
Q2	*2N5109 RF transistor*	*1,12,13,14*
D1	*1N5240 10-volt, 0.5-watt Zener diode*	*1,2,3,4*
Q2	*press-on heatsink from Thermalloy 2228B series*	*1*
C7	*3-15 pF trimmer capacitor*	*1,3*
C8	*7-40 pF trimmer capacitor*	*1,3*
C1,C12	*1000-pF feedthrough ceramic capacitor*	*1*
L1,L3,L4 L5,L9,	*100-μH RF choke*	*1,2,8*
L2	*47-μH fixed inductor*	*1,2*
L6	*180-μH RF choke*	*1,2,8*
L7	*1.5-μH fixed inductor*	*1,2*
L8	*0.56-μH fixed inductor*	*1,2*
C6	*82 pF ±10 percent, dipped mica*	*1,2,3*
C16,C17	*470 pF ±10 percent, ceramic epoxy coated*	*1,2,3*
C23	*22 pF ±10 percent, polyester film*	*1,2,3*
R4	*10.7-k ±1 percent, 0.25-watt resistor*	*1,2,3*
R5	*3.48-k ±1 percent 0.25-watt resistor*	*1,2,3*

Audio Amplifier

Reference no.	Component	Supplier
U1,U2,U3	*LF353 or TL072 dual BIFET op amp*	*1,2,3,4,8*
U4	*TDA 2006V 10-watt audio power amplifier*	*11,12*
D1,D2	*1N4001 rectifier diode*	*1,2,3,4,8*
C2-C10	*dipped mica, polyester film, ceramic epoxy coated ±5 percent*	*1,2,3*
R3-R8	*8.82 k ±1 percent 0.25-watt resistor*	*1,2,3*
R9	*6.34 k ±1 percent 0.25-watt resistor*	*1,2,3*
R10	*249 k ±1 percent 0.25-watt resistor*	*1,2,3*
R13	*10-k potentiometer audio taper*	*1,2,3,4*

AGC

Reference no.	Component	Supplier
U1	*EL2195CJ wideband op amp from Elantec*	*11,12,13,14*
U2,U3	*LF353 or TL072 dual BIFET op amp*	*1,2,3,4,8*
Q1	*MPF102 JFET transistor*	*1,7,8*
D1-D9	*1N4148 or 1N914 small signal diode*	*1,3,4,7*
R1,R6,R10,R12 R14	*single turn trimming potentiometer*	*1,2,3,4*
R16	*825-k ±1-percent 0.25-watt resistor*	*1,2,3*
R17	*604-k ±1-percent 0.25-watt resistor*	*1,2,3*
C5	*0.33 μF 10 percent polyester film*	*1,2,3*
C6	*0.0082 μF 10 percent ceramic epoxy coated*	*1,2,3*
C10	*0.1 μF 10 percent dipped mica*	*1,2,3*
SW1	*SPDT switch*	*1,2,3,4,7*

Notes:

1. 0.25-watt, 1-percent tolerance resistors have been used throughout this project. The use of 5 percent resistors is quite acceptable (unless indicated otherwise), as long as resistor ratios are maintained in voltage divider applications.

2. Unless indicated otherwise, capacitors are ceramic with 20 percent tolerance. The 2SK152 JFET can be replaced with a MPF102 JFET; however, the MPF102 JFET may require more time to attain an adequate frequency drift.

If you want to relax the requirements for the IMD level in the product detector, you can replace the TAK-3H mixer with a less expensive SBL-1 mixer requiring only +7 dBm LO drive. In this case, the 2N5109 transistor could be replaced with a 2N2222.

The TDA 2006V audio amplifier can be replaced with another low-distortion audio power amplifier with a power rating not lower than 8 watts.

The EL2195 op amp can be replaced with another wideband operational amplifier with a gain-bandwidth product no lower than 150 MHz.

Parts Suppliers

1. Newark Electronics
   4801 N. Rawenswood Avenue
   Chicago, Illinois 60640-4496
   (312)784-5100
   (213)516-6511

2. Digi-Key Corporation
   701 Brooks Avenue, South
   P.O. Box 677
   Thief River Falls, Minnesota 56701-0777
   1-800-344-4539

3. Mouser Electronics
P.O. Box 699
Mansfield, Texas 76063
1-800-346-6873-Sales
1-800-992-9943-Catalog

4. Allied Electronics
401 E. 8th Street
Fort Worth, Texas 76102
1-800-433-5700

5. Amidon Associates
12033 Otsego Street
N. Hollywood, California 91607
(213)760-4429

6. Mini-Circuits
P.O. Box 166
Brooklyn, New York 11235
(718)934-4500

7. Radio Shack
Check your local outlet.

8. Small Parts Center
6818 Meese Drive
Lansing, Michigan 48911
(517)882-6447

9. Fox-Tango Corporation
751 S. Macedo Boulevard
Port St. Lucie, Florida 34983
(407)879-6868

10. Spectrum International
P.O. Box 1084
Concord, Massachusetts 01742
(617)263-2145

11. Marshall Industries
(312)490-0155
(714)458-5395

12. Wyle Labs
(714)863-1611
(213)322-8100

13. Avnet Electronics
(714)754-6111
(213)558-2345

14. Hamilton
(714)641-4100
(312)860-7700

By Lewis F. McIntyre, *KB6IC*
3711 Gayle Avenue
Omaha, Nebraska 68123

ERROR CORRECTION IN DATA TRANSMISSION

Using Hamming codes to detect and correct errors in digital transmissions

HF communications techniques for hams have undergone a dramatic change over the past ten years. Electro-mechanical ASRs have given way to computer terminals, and Baudot has lost some of its popularity to ASCII, AMTOR, and packet. Data rates have increased from the venerable 45 baud to 300. But a "quick" packet exchange under less than ideal conditions proves that error detection and correction codes haven't kept up with these changes. In fact, packet and AMTOR use only error detection codes – requesting repeats when errors are found. AMTOR is effective, but slow. Under the best of conditions, its efficiency level is 50 percent. Packet, too, can quickly bog down on a noisy circuit because of its higher data rate.

For ordinary chitchat, straight RTTY is hard to beat. It's too bad the computer can't "fill in the blanks" on a few hits the way a human operator can, or can it?

Actually, your computer can fill in the blanks. Hamming codes, developed by Dr. Robert Hamming almost 40 years ago,[1] let computers detect and correct errors in digital transmissions. These codes are used in many applications. For instance, Hamming codes are used to ensure data integrity in the memory portion of the new VHSIC chips.

I'd like to introduce you to some Hamming code basics, and share some look-up tables and ideas for future development. I'll also analyze how these codes might be used to improve packet and AMTOR link performance.

How they work

Hamming codes can correct single-bit transmission errors. The mathematical process involved is quite complicated, so I'll skip the theory for now and go directly to an example.

Say you want to encode a four-bit data word into a seven-bit word called a "Hamming sequence." This is a 7/4 Hamming code (four data bits and three parity bits, for total of seven transmitted bits) which can correct single errors and detect two-bit errors in a received word. To see how this is done, pick a number between 0 and 15. Let's try 4. In **Table 1**, the encode table, find the number's Hex value (04H). Follow along the line to find the opposite value – 100 1100, or 4CH. This is the Hamming sequence you'll transmit. After you receive that sequence, decode it using **Table 2** – the decode table. The answer is 04H. Now simulate noise by changing any one of the seven received bits to its opposite value. Try making the least significant bit (LSB) a 1. This makes the received sequence 100 1101, or 4DH. Look up this Hamming sequence in **Table 2**, and read the decoded value. You'll find it's still 04H. Change any other bit, and you'll still obtain the correct value, 04H, from **Table 2**.

Why? The answer is obvious if you look at the decode table (**Table 2**). This table is eight times larger than the encode table (**Table 1**), because the decoded value of 04H (and all 15 other decoded answers) is decoded at eight entries. It's decoded once at the "no errors" entry of 100 1100 (marked with the asterisk), and seven times at 100 1101, 100 1110, 100 1000, 100 0100, 101 1100, 110 1100, and 000 1100 (all of the seven single-bit error positions possible).

Now change *two bits*. Make the sequence 100 1111. Your answer will be 07H, with errors detected. Two-bit errors will always give the wrong answer, but will never decode as a "no errors" entry marked with an asterisk. Depending on the word, three or four bits out of the seven sent would have to be changed for that to happen.

It's a bit harder to create a Hamming sequence[2] When doing so, you need to get into the mathematical part of the process. But because it doesn't really come into play once a look-up table sequence has been defined, I'll hold off on the theory once again. At this point, I'd rather pique your interest with some practical HF applications for this encode/decode scheme.

Data Word	Encoded Hamming Sequence Binary	HEX
00H	000 0000	(00H)
01H	110 1001	(69H)
02H	010 1010	(2AH)
03H	100 0011	(43H)
04H	100 1100	(4CH)
05H	010 0101	(25H)
06H	110 0110	(66H)
07H	000 1111	(0FH)
08H	111 0000	(70H)
09H	001 1001	(19H)
0AH	101 1010	(5AH)
0BH	011 0011	(33H)
0CH	011 1100	(3CH)
0DH	101 0101	(55H)
0EH	001 0110	(16H)
0FH	111 1111	(7FH)

Table 1. Encode table for 7/4 Hamming sequences.

Received Hamming Sequence	Data Word	Received Hamming Sequence	Data Word	Received Hamming Sequence	Data Word	Received Hamming Sequence	Data Word
000 0000	00H*	010 0000	00H	100 0000	00H	110 0000	08H
000 0001	00H	010 0001	05H	100 0001	03H	110 0001	01H
000 0010	00H	010 0010	02H	100 0010	03H	110 0010	06H
000 0011	03H	010 0011	0BH	100 0011	03H*	110 0011	03H
000 0100	00H	010 0100	05H	100 0100	04H	110 0100	06H
000 0101	05H	010 0101	05H*	100 0101	0DH	110 0101	05H
000 0110	0EH	010 0110	06H	100 0110	06H	110 0110	06H*
000 0111	07H	010 0111	05H	100 0111	03H	110 0111	06H
000 1000	00H	010 1000	02H	100 1000	04H	110 1000	01H
000 1001	09H	010 1001	01H	100 1001	01H	110 1001	01H*
000 1010	02H	010 1010	02H*	100 1010	0AH	110 1010	02H
000 1011	07H	010 1011	02H	100 1011	03H	110 1011	01H
000 1100	04H	010 1100	0CH	100 1100	04H*	110 1100	04H
000 1101	07H	010 1101	05H	100 1101	04H	110 1101	01H
000 1110	07H	010 1110	02H	100 1110	04H	110 1110	06H
000 1111	07H*	010 1111	07H	100 1111	07H	110 1111	0FH

Received Hamming Sequence	Data Word	Received Hamming Sequence	Data Word	Received Hamming Sequence	Data Word	Received Hamming Sequence	Data Word
001 0000	00H	011 0000	08H	101 0000	08H	111 0000	08H*
001 0001	09H	011 0001	0BH	101 0001	0DH	111 0001	08H
001 0010	0EH	011 0010	0BH	101 0010	0AH	111 0010	08H
001 0011	0BH	011 0011	0BH*	101 0011	03H	111 0011	0BH
001 0100	0EH	011 0100	0CH	101 0100	0DH	111 0100	08H
001 0101	0DH	011 0101	05H	101 0101	0DH*	111 0101	0DH
001 0110	0EH*	011 0110	0EH	101 0110	0EH	111 0110	06H
001 0111	0EH	011 0111	0BH	101 0111	0DH	111 0111	0FH
001 1000	09H	011 1000	0BH	101 1000	0AH	111 1000	08H
001 1001	09H*	011 1001	09H	101 1001	09H	111 1001	01H
001 1010	0AH	011 1010	02H	101 1010	0AH*	111 1010	0AH
001 1011	09H	011 1011	0BH	101 1011	0AH	111 1011	0FH
001 1100	0CH	011 1100	0CH*	101 1100	04H	111 1100	0CH
001 1101	09H	011 1101	0CH	101 1101	0DH	111 1101	0FH
001 1110	0EH	011 1110	0CH	101 1110	0AH	111 1110	0FH
001 1111	07H	011 1111	0FH	101 1111	0FH	111 1111	0FH*

Table 2. Decode table for 7/4 Hamming sequences. Asterisks indicate "no errors detected." All other entries have single errors detected.

Implementing the Hamming codes

Why, if they're so easy to implement, aren't Hamming codes more popular? First, there's no agreed-upon protocol. Second, the string of four data bits isn't long enough. The Baudot code uses five bits, with extra characters (FIGS/LTRS) to shift back and forth between two 32-character alphabets. ASCII uses seven bits, and eight are preferred to allow full data transfer. Can't you just break an ASCII word into two four-bit "nibbles," encode and send each, then decode, correct, and add them back up at the receiver?

In theory, you can. But you'll encounter another HF error – fading. Fading causes the entire loss, or "erasure," of one or several words. And, if the receiver was expecting a LSB "nibble" when the fade started, but picked up a most significant bit (MSB) nibble when it ended, it would assemble an incorrect word. Because the receiver's definition of MSB and LSB is now out of sync, all subsequent words will be reassembled incorrectly.

In manual systems like straight RTTY, you could treat this problem the same way you'd treat a FIGS/LTRS garble. You'd use a key to direct the computer to shift the order in which it's reassembling the nibbles.

You could also devote one of the four bits as a flag, indicating whether it's an MSB or LSB nibble. This would leave six bits – enough to encode a 64-character alphabet like Baudot.

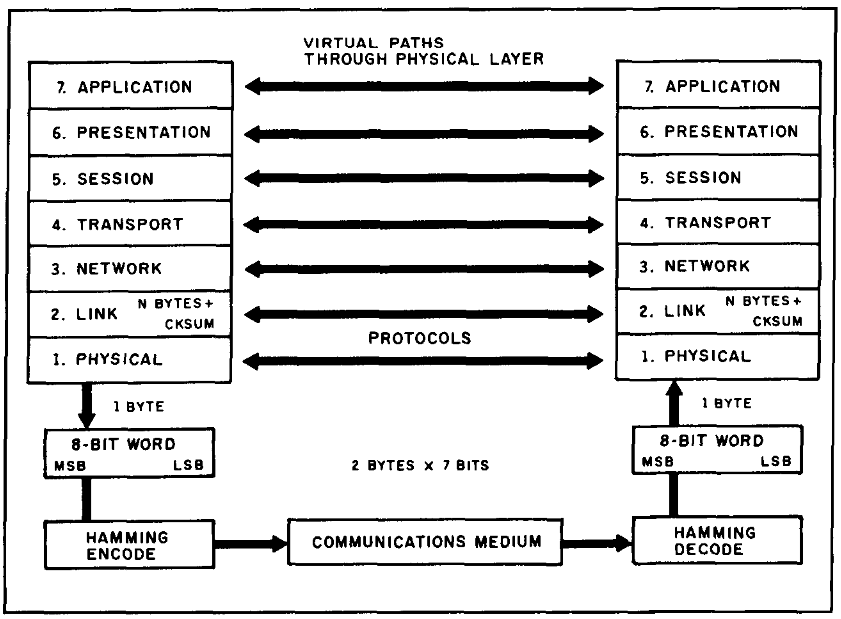


Figure 1. Hamming application to AX.25.

Packet applications

A third possibility would be to place the Hamming codes inside an error-detecting block code. You could send a block of a fixed number of characters (say 127), compute a checksum of each character, and then send the checksum. The receiver would perform a similar process, acknowledging the text if it agrees with the checksum, or asking for a repeat if it's incorrect. This common algorithm for data transfer is used by XMODEM for landline and in the link protocol in AX.25 packet.

The AX.25 packet protocol is organized in "layers." The link layer organizes a block, computes checksums, determines if a block is received correctly, and handles repeat requests. The bottom-most layer is the physical layer. This layer is normally concerned only with modulation and demodulation. It's at the physical layer that you'd intercept a single eight-bit word on its way to the modulator, encode it, and reverse the process at the receiver. **Figure 1** shows how you could include Hamming encoding and decoding at this level without disturbing any of the other layers, except, perhaps, to allow more time for the longer Hamming codes. What do you gain by doing this?

The AX.25 protocol already accounts for packets of incorrect length. Thus, erasures, and the framing problems they may generate, can be detected and handled by requests for retransmission generated by the existing link protocol. And, because the checksum is also encoded as a Hamming sequence, errors which can't be corrected will probably be detected as well, resulting in a retransmission request. You can do all of this without touching the higher-order packet protocols. In all probability, this scheme will correct *all* single-bit errors per Hamming sequence, and detect all erasures (incorrect packet length) and uncorrectable errors (bad checksum). How high is this probability? Let's see.

Number crunching

The Bit Error Rate (BER) is the probability that any bit will be changed. The basic packet word is eight bits long, and each bit must be correct. The probability of receiving an entirely correct eight-bit word is:

$$P(8 \text{ correct bits}) = (1 - BER)^8 \qquad \textbf{(1)}$$

Packets come in various lengths; 128 words is representative. The last word is a checksum, allowing errors in the block to be detected. All 128 words must be received correctly. The probability that 128 correct eight-bit words, or one entire packet of representative length, will be received is:

$$P(\text{correct packet}) = P(8 \text{ correct bits})^{128} \quad \textbf{(2)}$$

The probability that a correct packet will be received is defined as the number of successes divided by the number of attempts. The inverse of this is the average number of attempts that must be made to get one good packet through:

$$N(\text{attempts}) = 1/P(\text{correct packets}) \quad \textbf{(3)}$$

The results are plotted on the graph in **Figure 2**. They show that, without error correction, the number of attempts remains at essentially one transmission per packet up to about 0.0001 BER (1 bit per 10,000 altered). The number of attempts then rises quickly to an average of two transmissions at a BER of 0.0005 and three at 0.001. Beyond that level, the number of attempts required to successfully get a packet through become astronomical.

By comparison, a Hamming sequence will be accepted if it has either no errors, or a single-bit error. Because a Hamming sequence is seven bits long, the no-error probability is:

$$P(7 \text{ correct bits}) = (1 - \text{BER})^7 \quad \textbf{(4)}$$

and the probability of exactly one error is:

$$P(\text{exactly 1 bit error in 7}) = 7*\text{BER}*(1 - \text{BER})^6 \quad \textbf{(5)}$$

Thus, the probability of no errors, or exactly one error, is the sum of **Equations 4** and **5**:

$$P(0 \text{ or } 1 \text{ error in 7}) = P(7 \text{ correct bits}) + P(\text{exactly 1 bit error in 7}) \quad \textbf{(6)}$$

Because you can only encode four bits onto a seven-bit Hamming sequence, you need 256 Hamming words with one or zero errors each, to convey 128 words with no errors. The probability of this occurring is:

$$P(256 \text{ correctable sequences}) = P(1 \text{ or } 0 \text{ errors in 7})^{256} \quad \textbf{(7)}$$

And, like the eight-bit word, the number of attempts required is:

$$N(\textit{Hamming attempts}) = 1/P(256 \text{ correctable sequences}) \quad \textbf{(8)}$$

This is also plotted in **Figure 2**. You can see that Hamming sequences require no retransmittal until there's a BER of 0.005 – nearly twice the BER that will bring an uncorrected link to its knees! A Hamming encoded link can maintain a useful throughput with less than two attempts per packet until it reaches a BER of 0.02 – nearly 20 times higher than the link without error correction.

Of course, this doesn't come without cost. You may have noticed that Hamming sequences are 7/4 longer; that is, they are nearly twice as long as the unprotected packet. Does this overhead pay for itself?

Yes. The number of bits per packet is the basic number of bits per individual packet times the average number of attempts:

$$\text{TXBITS(Hamming)} = 7*256*N(\text{Hamming attempts}) \quad \textbf{(9)}$$

and

$$\text{TXBITS(Normal)} = 8*128*N(\text{attempts}) \quad \textbf{(10)}$$

These are plotted in **Figure 3**. Note that for low error rates, the uncorrected link without Hamming codes outperforms the Hamming link by almost 2:1, requiring only 1024 bits compared with 1792 Hamming bits. The link errors are too few to justify the high overhead of the Hamming bits. At about 0.0005 BER, the two are equal in performance. On the average, the Hamming link will require less than two transmissions for BERs up to 0.01.

This analysis doesn't include the possibility of using the more robust, but slower, Hamming codes to support HF data rates which could go as high as 1200 baud – unthinkably fast for conventional HF packet.

Application to AMTOR

What about AMTOR? AMTOR uses a seven-bit alphabet, with just four 1s and three 0s. A total of 35 characters can be

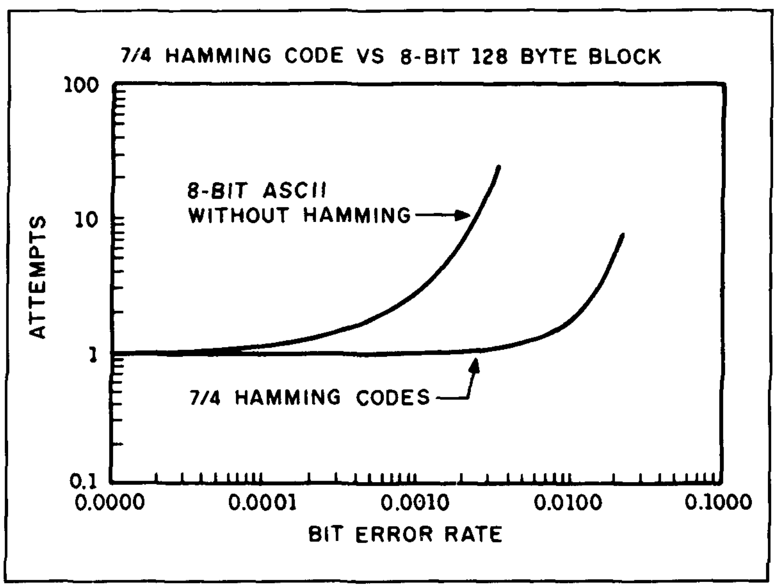


Figure 2. Attempts versus BER, 128-byte packet.

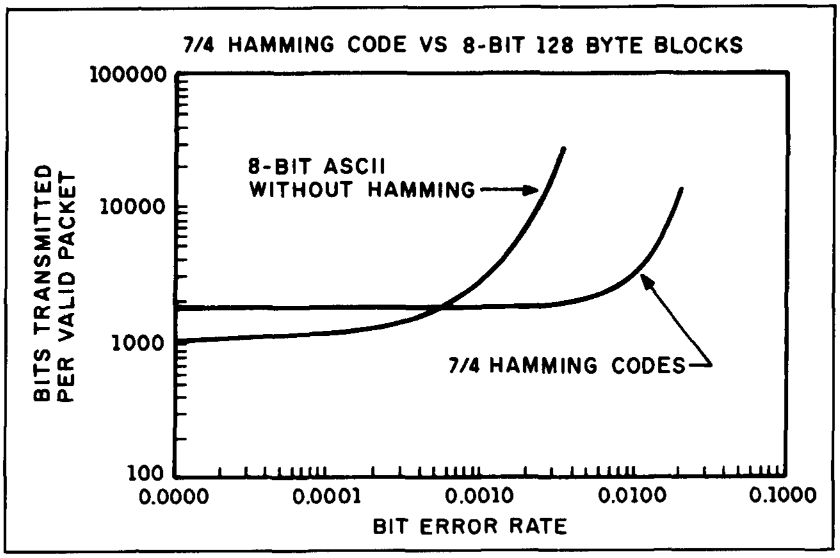


Figure 3. Average transmitted bits per 128-byte packet.

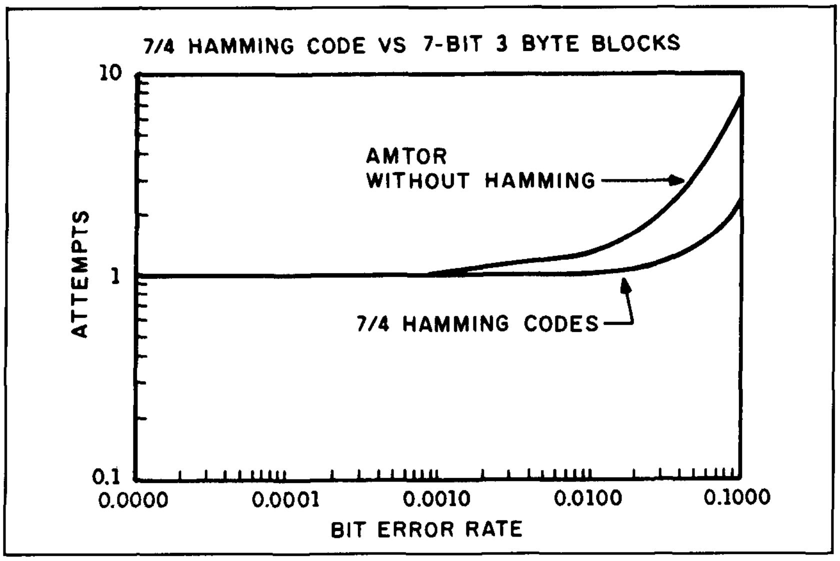


Figure 4. Attempts versus BER, AMTOR 3 bytes × 7 bit group.

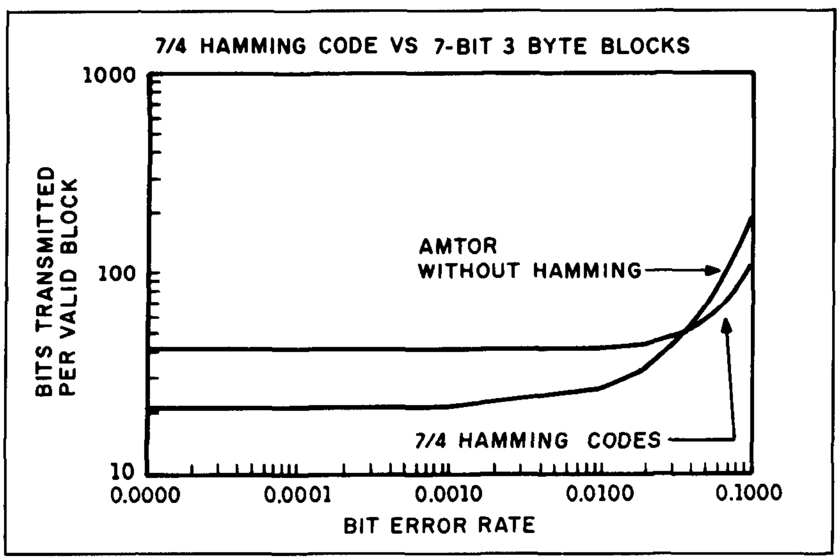


Figure 5. Average transmitted bits per AMTOR 3 bytes × 7 bit group.

encoded. Characters are sent in groups of three. Any character which doesn't confirm to this 4/3 sequence is detected as an error, and generates an RQ (Repeat Request).[3]

The analysis is basically the same as it is for packet. The exception is that the basic block is three seven-byte words, which can be encoded onto six seven-byte Hamming sequences. **Figures 4** and **5** show the results. Surprisingly, Hamming sequences have little advantage over short AMTOR sequences. Uncorrected AMTOR more than holds its own until enormously high BERs of 0.05 are reached. Only in the presence of absolutely incredible noise levels of 0.1, does AMTOR Hamming gain a 2:1 advantage over straight AMTOR. These results shouldn't come as a great surprise to AMTOR enthusiasts. The key lies in the very short AMTOR block lengths. However, I haven't taken the effects of false acknowledgement into consideration here — specifically the misinterpretation of an "ACK" (Acknowledgement) as an "RQ" (Repeat Request).[3]

Frankly, AMTOR doesn't appear to be much improved by Hamming sequences. Using a higher data rate wouldn't change this situation much because AMTOR spends a significant amount of time waiting for transmitters to change over. Shortening the data transmission time wouldn't appear to reduce the overall time significantly. AMTOR efficiency might improve if a longer sequence with Hamming codes is used, but that would involve a major change to the AMTOR protocol.

Technical details of Hamming codes

How do Hamming codes work? To understand how they work, you must first understand the concept of Hamming distance. Hamming distance is the number of bits that would have to be changed to transform one binary word into another. For example: 0011, 0101, 1001, and 0000 are all within one Hamming distance of 0001. By contrast, 1110 is separated by four bits distance from 0001, and all four bits would have to be altered to change 1 (01H) to 14 (0EH). Hamming distance can be computed by "exclusive OR'ing" the two binary words together and counting the 1s.

Hamming sequences which can correct single-bit errors use an "alphabet" in which all the legal sequences that can be transmitted have a Hamming distance of three from each other. For example, **Table 1**, the encode table used in the text, has sixteen seven-bit sequences out of a possible 128, all of which differ from each other by three or more bits. The other 112 possibilities represent erroneous sequences caused by noise that creates a one-bit change in one of the sixteen legal sequences. Because the legal sequences are separated from each other by three bits, these erroneous sequences will be separated by two or more bits from all other legal sequences — except for the one which was actually sent. Using the example in the text, if the sequence encoding 04H is altered at the LSB to 100 1101, this sequence is still at least two bits removed from any other **Table 1** sequence, except 100 1100. Thus, you can assume that an erroneous sequence should actually be the legal sequence "closest" to it.

Two-bit errors still won't produce a legal sequence. They won't decode correctly, either. Hamming codes of distance three can't distinguish between a correctable single-bit error and an uncorrectable two-bit error.

Generating the Hamming sequence

Hamming sequences use parity bits based on the message word to be encoded. The parity bits are defined so they will actually point to zero (the error-free condition), or a binary number representing the bit position in error. Because of this, bits in a Hamming sequence are numbered from 1 to N, rather than the binary starting point of zero. Three parity bits are needed to handle a seven-bit sequence, leaving four bits available as message bits.

Table 3 shows how parity bits are defined for 7/4, 15/11 (four parity bits), and 31/26 (five parity bits) sequences. Parity bits are assigned to locations corresponding to integer powers of two within the sequence (bits P1, P2, P4, P8, and P16), and message bits to all others. A "1" at the intersection of a message bit row and parity bit column, means that the message bit should be included when determining the corresponding parity bit. Following the example in the text, "1s" appear opposite M3, M5, and M7, under parity bit P1. These bits are XOR'd together to form the parity bit 1. P2 is the XOR of message bits M3, M6, and M7, and P4 is the XOR of M5, M6, and M7.

Hamming sequences were originally intended for hardware generation and detection.[4] **Figure 6A** shows how the 7/4 code above can be created in hardware for the transmitter. The parity bits are woven into the sequence as shown. This figure illustrates the generation of the 1001100 sequence for 04H, given in the text.

The receiver in **Figure 6B** generates the same parity bits — P1, P2, and P4 — and XOR's each with its corresponding received parity. The resulting three-bit word is called the "syndrome," and points to the binary position of the bit in error. As shown, the receiver copies 1001101, generating a syndrome of 7. The syndrome is applied to a 1-of-8 decoder. Zero is the "no errors detected" condition; 1, 2, and 4 indicate that the parity bits themselves were in error and aren't needed to correct the message bits. Finally, 3, 5, 6, and 7 are XOR'd with their corresponding message bits. Bit 7 in the example is XOR'd by the decoded syndrome, back to its correct value of 0.

The Hamming codes were developed when such hardware solutions were essential for implementing the technique. Solutions of this type are still necessary for longer sequences where the decode tables can become prohibitively large. However, as noted in the text, look-up table schemes in software are now a more efficient implementation for short sequences like the 7/4 code.[2]

	P16	P8	P4	P2	P1
M3	0	0	0	1	1
M5	0	0	1	0	1
M6	0	0	1	1	0
M7	0	0	1	1	1
			(7/4 sequences)		
M9	0	1	0	0	1
M10	0	1	0	1	0
M11	0	1	0	1	1
M12	0	1	1	0	0
M13	0	1	1	0	1
M14	0	1	1	1	0
M15	0	1	1	1	1
		(15/11 sequences)			
M17	1	0	0	0	1
M18	1	0	0	1	0
M19	1	0	0	1	1
M20	1	0	1	0	0
M21	1	0	1	0	1
M22	1	0	1	1	0
M23	1	0	1	1	1
M24	1	1	0	0	0
M25	1	1	0	0	1
M26	1	1	0	1	0
M27	1	1	0	1	1
M28	1	1	1	0	0
M29	1	1	1	0	1
M30	1	1	1	1	0
M31	1	1	1	1	1
	(31/26 sequences)				

Table 3. General scheme for determining parity for 7/4, 15/11, and 31/26 Hamming sequences. A "1" indicates that the corresponding message bit should be included in the XOR tree for that particular parity bit.

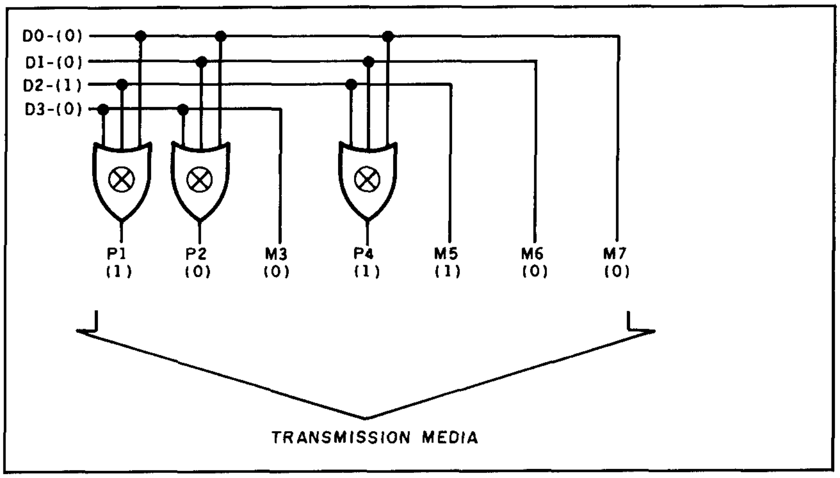


Figure 6A. Hardware encoding of data 0100 onto 7/4 sequence 1001100.

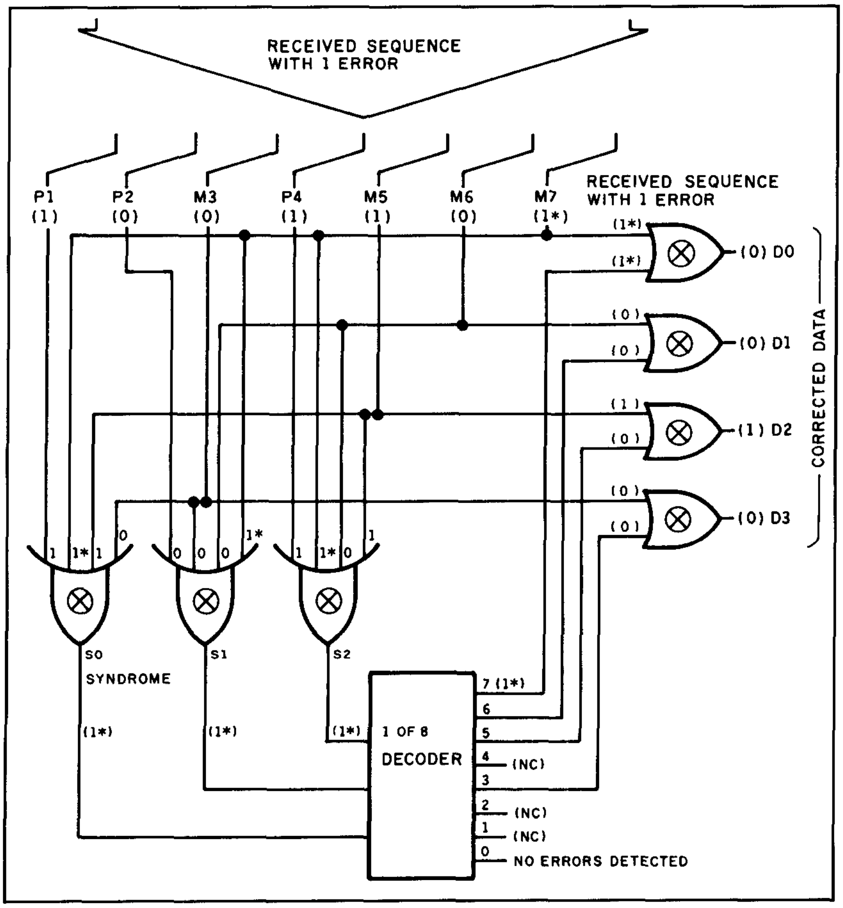


Figure 6B. Hardware decoding of a 7/4 receiver. "*" indicates bit in error and corrections.

Hamming codes assume that two-bit errors within a word are much less likely to appear than single-bit errors. This isn't strictly true. Many errors occur in bursts, causing multiple errors and even complete loss, or erasure, of entire the word. Nevertheless, Hamming codes can easily correct most errors in communications.

Summary

The main limitation to the more widespread implementation of this simple scheme is the lack of an accepted protocol. The techniques I've described here are easily implemented on any computer/TU system capable of straight ASCII operation. The integration of this technique into the AX.25 packet protocol may be a bit more complicated, but it would improve HF packet throughput significantly. In fact, Hamming code applications could improve this form of packet to such a dramatic extent, that some serious research may be in order.

I invite all who wish to experiment with the development of a new protocol based on the Hamming technique to contact me, either by mail or at my packet address, KB6IC @ KØBOY. ■

References

1. R.W. Hamming, "Error Decting and Error Correcting Codes," *Bell System Technical Journal*, Volume 29, April 1950, pages 147-160.
2. Ben White, "Hamming-Code Decoding," *Dr. Dobbs Journal*, October 1989, pages 52-56.
3. *The ARRL 1986 Handbook*, 63rd Edition, The American Radio Relay League, Newington, Connecticut, 1986, page 19-10.
4. John D. Lenk, *Handbook of Logic Circuits*, Reston Publishing Company, Reston, Virginia, 1972, pages 90-93.

By Marv Gonsior, *W6FR*
418 El Adobe Place
Fullerton, California 92635

TELEPHONE SUSCEPTIBILITY TO RFI

Some solutions to a difficult problem

Telephone interference from RF can be a big problem. Even if you aren't dealing with it now, if you live in an urban area, you probably will at some point. This is a near certainty with the proliferation of electronic telephones, their associated gadgetry, and modems. Since the breakup of the AT&T system, and with the consequent widespread ownership of our telephones, Amateurs are now faced with the obligation, if not the responsibility, to solve this complex problem. Almost every case will vary to some degree. Gone are the days of harmonic TVI. This new source of misery has reared its ugly head instead! Here are some solutions to an often serious problem.

Background

The auto-rectification which takes place in the multitude of new electronic telephones of questionable design, is becoming a problem for Amateurs. The responsibility of the new privately owned telephone companies stops at the junction box where their lines terminate upon entering a structure, instead of at the telephone unit – unless the phone is leased. The telephone company will attempt to solve the problem up to that point by placing a paper capacitor across the line or an inductive filter in series with it. Neither of these solutions is generally effective, because you then have a relatively large antenna made up of all the internal wiring within the structure.

Before the advent of electronic telephones, the cure was relatively simple. You connected a 0.01 disc capacitor across the microphone, or earphone, or both. This is still worth a try in most minor cases of telephone interference, particularly if you're dealing with a conventional, nonelectronic unit.

Recently, I purchased a quality AT&T Trimline™ type 220 electronic telephone. I anticipated the interference problem, but was confident I could solve it easily. I wound up facing a great challenge, because the level of interference I experienced from my 14-MHz SSB signal was severe and unacceptably high. I soon learned just what knowledgeable friends meant by their wishes of "good luck!"

Trying to solve the problem

My first attempt at solving the problem, provided no detectable improvement. I tried using the AT&T Z100A radio-interference filter. It's nicely designed, from a solely mechanical standpoint, because of its modular construction. However, I found it was worthless as an RF filter on 14 MHz at my location. It was also rather expensive. The warranty covers only defects; everything else is excluded. AT&T states that it was designed "to *eliminate* extraneous noises from local AM radios that your phone might pick up. It will also *minimize* stations, citizen band, and ham radios." From my experience, that's just about what it will do – "eliminate" the former and *maybe* "minimize" the latter.

The filter consists of two side-by-side inductors in a common ferrite cup-core configuration, one in series with each side of the line. Two different inductors from two separate units measured 6.9 and 7.2 mH, respectively, with a Q of 5.5 – which explains why they're so ineffective. They simply do not offer sufficient impedance at HF because they are, in effect, essentially capacitors at 14 MHz. Perhaps at other frequencies, like 160 meters and much lower (the frequencies for which they were apparently designed), they might be more effective. However, when I performed a simple bench test, I found no evidence to support this line of thought. I'm told that in one case, a telephone company service representative placed three Z100As in series without achieving any measurable attenuation at 14 MHz. I tried two in series with the same result. The test setup I used (shown in **Figure 1**) attempts to simulate the telephone line, at the telephone unit, with RF coupling.

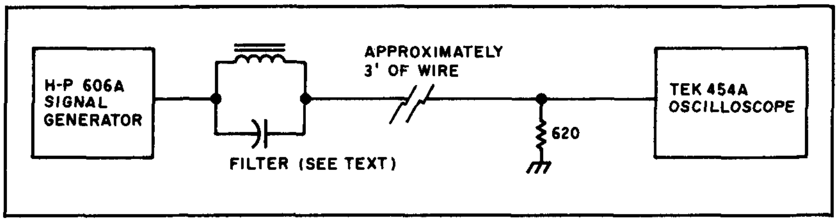


Figure 1. Test equipment used to simulate the telephone line, at the telephone unit, with RF coupling.

Other attempts

I tested another telephone "noise filter," no. 52-322, manufactured by Genmark Products of Concord, California against my station equipment. The company recommends this filter for "eliminating radio frequency interference." Its design is strikingly similar to that of the Z100A; it has the same electrical configuration. However, the two inductors measured 11.9 mH each, and the filter's performance at 14 MHz was equally poor! I had hopes that it would be a dark horse winner. It would have been quite cost effective, because it was one fourth the price of the Z100A.

I tried to cure my Trimline telephone using a number of approaches — all aimed directly at the unit. Here are some of them:

- small 200-μH ferrite toroids in series with each line,
- telephone-to-wall extension line cord wrapped through a 2.5-inch ferrite toroid with 10 turns,
- 34 turns of telephone-to-wall extension line cord wrapped around a 3/8-inch by 8-inch ferrite rod, and
- two small, 1.2-mH chokes on ferrite cores in series with the lines.

None of these ideas netted me any degree of success, except the ferrite rod filter, which did make a dent in the problem. But, besides being messy, it was an unacceptable solution. Unfortunately, none of the ferrites had identifiable mixes. In desperation, I tried another configuration which solved the problem.

Eureka!

Using a gutted Z100A as a mechanical test bed, solely for the convenience of being able to use the connectors and the empty circuit board for mounting, I parallel-resonated two toroidal cores at 14 MHz. This offered high impedance. One core was inserted in series with each side of the line. I used Amidon part no. 37-6 (yellow color code). Amidon part no. 37-2 (red color code) also works quite well. These inexpensive cores are both 3/8-inch OD and made of powdered iron for high Q. For frequencies above 14 MHz, Amidon's 37-6 core would probably perform a bit better. It took 17 inches of no. 28 enameled wire wound tightly through the cores to yield 3.2 μH of inductance, with about 36 pF of mini-disc capacitors (made up from a kit of Radio Shack's part no. 272-806) to resonate at 14 MHz. I used a grid-dip oscillator (GDO) to trim them[1] but none of the values are too critcal as long as you obtain resonance at mid-band. Using the test setup in **Figure 1**, I measured approximately 30 dB of attentuation in the 14-MHz band. These cores will just fit side-by-side inside the Z100A case. The result, after hours of frustration, was most rewarding. I can now run a kW with a beam pointed directly at the house, without a sound in the telephone.

Here's a second broadband solution. It's not as effective as the resonant filter, but has a large impedance advantage. Amidon recommends inserting ferrite chokes consisting of 20 turns of no. 26E wire on their FT50A-75 cores (initial permeability of 5000) in both lines at the telephone unit[2] This resulted in a measured 120 μH of inductance. Alternatively, they recommend running about 6 to 9 turns of the telephone-to-wall cable through one of their FT-140J ferrite cores (initial permeability of 5000), which measure 1.5 inches OD.

I tried both of these solutions with almost complete success while pointing a kW on 14 MHz at a spot near the house. However, when the beam was pointed directly at the house, these solutions minimized, but didn't totally eliminate, the interference. All of this points out the level sensitivity of this problem. Refer to **Figure 2** for the impedance-versus-frequency measurements for Amidon's type FT50A-43 and FT50A-75 filters. They were taken on a HP 4191A RF impedance analyzer. **Figure 3** shows the results using Amidon's higher-Q type-61 material (found in their FT-50A-61 core) for higher frequency response. You may place different filters in series; they will be additive.

Using these data, you can optimize a broadband filter for the frequency(ies) involved. Unfortunately, my test data indicates that only a resonant filter may provide significant attentuation to be effective for severe cases at 28 MHz.

Other considerations

It appears that, in addition to the various active semiconductors inside "modern" electronic telephones (I doubt that they would meet the old FCC criteria of modern design), there's usually a varistor or two. These are real culprits, that make a fine rectifier. I've been told that some have dealt successfully with this problem by going inside the telephone base and bypassing the varistors with a disc capacitor. Your neighbors may not take kindly to this approach, and it could be a bit difficult to solder the capacitor onto the board, but it might be worth a try.

Hardly a day goes by that I don't hear of someone who's struggling with telephone interference. Some have had partial success in reducing it to an acceptable level by winding a number of turns through one of Radio Shack's part no. 273-104, MFJ's part no. 107, or Amidon's part no. 2X-43-XXX (which requires an enclosure). All of these are snap-together ferrite, square toroids. They are, unfortunately, rather bulky and somewhat unsightly. However, from my experience with a severe case, only the resonant method offered sufficient impedance to provide a total solution.

In lieu of the Z100A as an enclosure, and because of its cost and size, I'm using K-Mart's Modular Jack, part no. TA61 (Gemini Industries, Clifton, New Jersey 07014), as an adapter. It's available for one tenth of the cost of the Z100A. However, *it requires a short telephone* standard modular cable held in place with silicone rubber (RTV), 5-minute epoxy, or Krazy Glue™ along with a standard telephone modular plug (see **Figure 4**). **Figure 4** also shows some very good and conveniently available mountings for the toroids. This enclosure will make assembly easier, as it provides more working space. It's white and offers a reasonably unobtrusive appearance by your telephone. If you have the room, and the interference level is sufficiently low, you may connect the series filters inside the base of the telephone unit at the point where the lines enter.

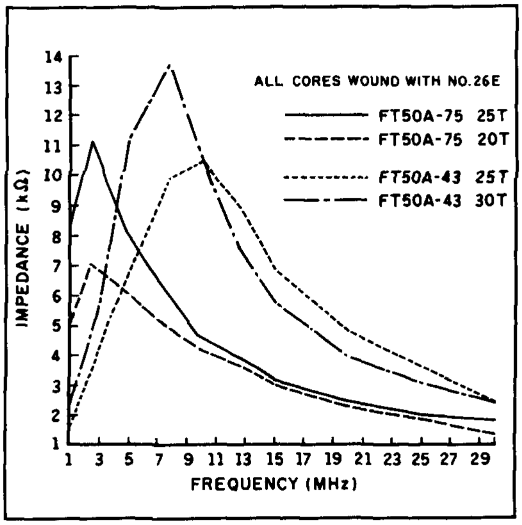


Figure 2. Impedance versus frequency measurements for Amidon FT50A-75 and FT50-43 filters.

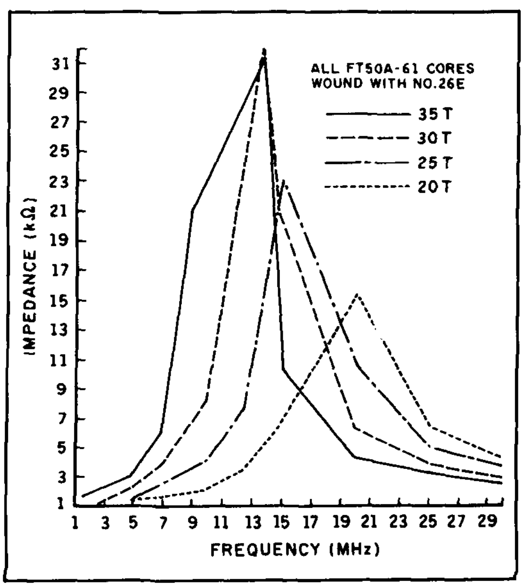


Figure 3. Impedance versus frequency measurements for Amidon FT50A-61 filters.

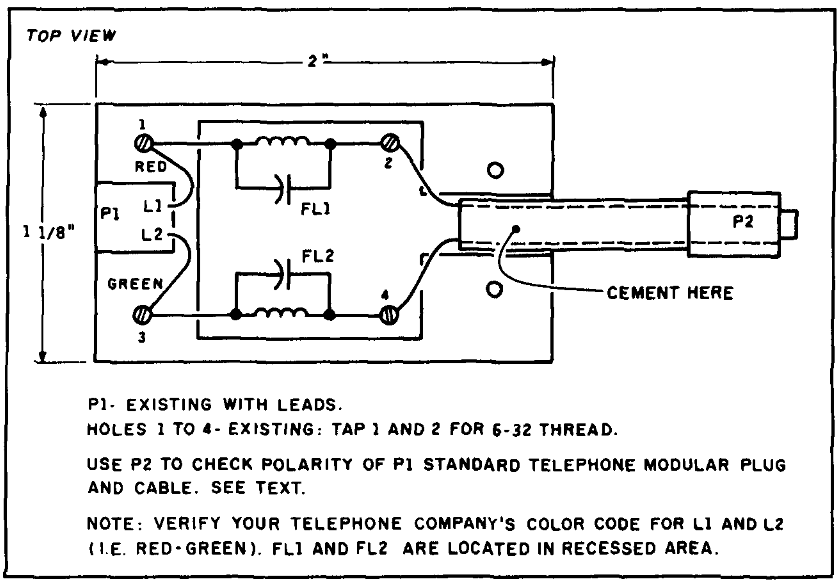


Figure 4. Assembly of the telephone filter using an enclosure with conveniently available mountings for the toroids.

My monoband solution may not be effective for those with multi-band operations. In serious situations involving multi-band operations, you should probably series resonate filters for each band you work. Alternatively, you could use two broadband filters in series. Amidon's catalog solution may provide a broadband answer to this problem, but I'm unable to test these components at high power at other operating frequencies and conditions.

From the impedance measurements shown in **Figure 2**, and from my near total success at 14 MHz, I feel certain that Amidon's filter configuration will be quite effective in a broadband sense — but only in those cases where the interference level is sufficiently low. It's apparent that the core losses at high frequency are a limiting factor in most of the materials. A low-pass filter would be a desirable alternative. However, because there really isn't a good ground available in a telephone installation, the design of that type of filter in this application would probably be impractical.

You'll find reference reading on the subject of telephone interference in various sources.[3,4] Recently, articles have appeared in Los Angeles and St. Louis newspapers, describing the serious, widespread problems not only of telephone interference, but interference with other equipment as well, resulting from FM broadcast stations. I'm certain that these situations aren't at all unique.

Conclusions

From my limited work, it appears that there may not be a single broadband, external solution to this complex problem. Success in eliminating telephone interference depends upon the relative signal level, type of telephone equipment, installation, and frequency(ies) involved.

One possible solution is to use a combination of the methods I tried, as your circumstances dictate. For example, in a severe case, it appears that only a resonant filter will be adequate. It could be used in conjunction with a broadband, nonresonant filter to cover other operating frequencies. This has the added benefit of the broadband filter supplementing the resonant one.

In minor cases, the broadband series chokes will probably offer sufficient impedance to solve the problem. Two modular filters, one resonant and one nonresonant, make a nice kit to investigate problems. I use the kit to determine what course of action is necessary relative to the signal level encountered. Now, I can truly say "good luck" to my friends! I hope others will share their knowledge about this complex area of RF susceptibility.

Acknowledgment

My sincere appreciation is extended to Amidon Associates, Inc. for their assistance in making measurements and providing guidance during the course of my work. ■

REFERENCES

1. *The ARRL Handbook*, 1989 edition, page 2-35, Figure 69.
2. *Amidon Associates Catalog*, Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607, September, 1989, page 54.
3. *Interference Handbook*, Federal Communications Commission, 1986 edition, page 56.
4. *The ARRL Handbook*, 1989 edition, page 40-18.

By Max Carter
46 14th Street
Wheatland, Wyoming 82201

SUPER NARROW-BAND TECHNIQUES EQUALIZE POWER INEQUITY ON 1750 METERS

Join an intrepid group of radio experimenters — the LowFers.

Most of you are probably aware that Part 15 (15.112) of the FCC rules allows unlicensed operation in a small band of frequencies in the long wave (low-frequency) portion of the radio spectrum. This band (1750 meters) extends from 160 to 190 kHz.* Operation is on a totally unprotected basis. Transmitters are limited to 1-watt input and antennas to 50 feet (including feed line). The transmission mode can be anything except spark; all out-of-band emission must be kept at least 20 dB down. Ham calls can't be used for ID.

The power and height restrictions are especially severe given the fact that 50 feet represents roughly 1/100 wavelength at these frequencies. With a radiation resistance typically measured in milliohms, a 1-watt transmitter feeding a 50-foot antenna is lucky to radiate a milliwatt or so at 175 kHz. Almost all (99.9 percent) of the input power is dissipated as heat in the large inductor needed to resonate the antenna, and in the ground.

Lowdown on the LowFERS

Conditions on 1750 meters are brutal. Not only is the band filled with signals from powerful government and broadcast transmitters, but you must contend with "power-line carriers" (signals used by electric utilities) inadvertently radiated from long power transmission lines. If that weren't enough, *noise* from natural and manmade sources (SCR-operated light dimmers and power-line corona discharge are two) can make the band unusable at some locations and times of the day or year.

Despite its limitations, an intrepid group of radio experimenters has occupied the 1750-meter band for a number of years.

Despite its limitations, an intrepid group of radio experimenters has occupied the 1750-meter band for a number of years. Perhaps they're fascinated by the band's unique propagation characteristics. Maybe they're simply attracted to the challenge of perfecting hardware and listening techniques to overcome its difficulties. These experimenters call themselves LowFERS (Low Frequency Experimental Radio Stations).

LowFERS have achieved a great deal. They copy ground wave signals (*identify* may be a better word) routinely at several hundred miles. Sky wave copy, while somewhat unpredictable, is somewhat routine to 1000 miles for some operators. I believe the

*That's 375 kHz below the broadcast band. Ed.

copy of several West Coast LowFERS in Hawaii is the DX record!

The 1750-meter band

The 1750-meter band is characterized by predictability and stability. Its main strength is long distance ground-wave propagation. But there's a paradox here: 1000-mile LowFER DX — usually accomplished using CW, conventional long-wave hardware, and aural copy — has relied on relatively infrequent skip conditions, most often at night. Daytime DX, when stable ground-wave propagation predominates, is usually limited to a few hundred miles (see **Appendix A**), despite sharp filters and sharp ears at the receive end. Why? Because a typical LowFER can't radiate enough raw *power* to overcome background noise at the receiver. Higher powered stations are easy to copy at 1000 miles during the day.

I'd like to discuss the details of some interesting baseband signal processing techniques ideally suited to correcting this power inequity. The techniques, which are by no means new or revolutionary, find wide application in such diverse fields as deep space communications and statistical analysis. You may notice some similarity to coherent CW (CCW).[1,2]

While this isn't a construction project, I'll present some actual working circuits. Frequencies are low enough so common, garden-variety CMOS ICs and perfboard construction techniques are adequate (no GaAsFETS, MMICs, or striplines are needed on LW).

In recent years, the LowFERS' goal has been to demonstrate that reliable (or at least predictable) ground-wave *communication* is possible at 1000 miles within the constraints of FCC rule Part 15.

Super-slow signaling

Super-slow signaling, combined with super-narrow receiving equipment, is the key to achieving power parity on 1750 meters. The idea is to take advantage of the fact that as bandwidth becomes narrower, signal-to-noise ratio (S/N) increases in direct proportion. For example, reducing bandwidth from 100 to 10 Hz results in a ten-fold (10-dB) improvement in S/N. Most of you are familiar with a narrow audio filter's ability to pull an otherwise "nonexistent" signal out of noise. To derive the maximum benefit from this principle, the channel bandwidth (expressed in Hz) must be matched to the signaling rate (expressed in bauds) of the data being transmitted; that is, 1-Hz bandwidth for each baud of signaling rate. (The signaling rate, *bauds*, shouldn't be confused with the data-transfer rate, *bits per second*, which can be faster.) The channel bandwidths used in the super-slow system seem to border on the preposterous. They are: 10, 1, 1/10, and 1/60 Hz. Filter bandwidths of 10 Hz are about the narrowest you can use for aural/CW LowFER DX work.

Narrowband problems

There are problems associated with extremely narrow communications channels. The first lies in tuning the transmit and receive ends to the same frequency. Anyone who has used a very narrow CW audio filter knows how hard this can be at times. So, imagine the difficulty you might experience at 1/60 Hz!

A second problem involves the lack of damping that some narrow filters exhibit. In an effort to achieve narrowness of bandwidth, the filter builder may resort to a high-Q design using a lot of positive feedback. Such a filter can be susceptible to impulse noise. It can ring like a bell when "struck" with a high-amplitude noise spike. Some filter designers seek to overcome this by cascading a number of low-Q sections. This can be effective but, because bandwidth decreases in proportion to the square root of the number of cascaded sections, you may need hundreds of sections to achieve a nonringing 1/10 or 1/60-Hz filter.

Three techniques are used to obtain nonringing, supernarrow bandwidths. They are: fully synchronous modulation-demodulation (and coding-decoding), Binary Phase-Shift Keying (BPSK), and the integrate-sample-reset (integrate and dump) filter.

If the full potential of these techniques can be realized, the payoff in equivalent transmitter power is truly awesome. Assuming your baseline starting point is a 10-baud, 10-Hz channel bandwidth aural-CW receiving system, the super-narrow system operating at 1/60 baud could give you as much as 36.8 dB improvement in S/N. This breaks down to 3 dB from synchronous detection, 6 dB from BPSK, and 27.8 dB from bandwidth reduction, which is equivalent to increasing transmitter power to 4800 watts.

As you might expect, manual and aural reception methods are out at these slow rates. The transmission and reception process must be "automated." As you'll see, the system isn't tuned; it's programmed.

Synchronous detection

You could call the synchronous detector a universal detector. Depending on its configuration and the carrier source, it can be used to detect AM, SSB, FM, and PM signals, or exotic combinations of all four. That's because it's sensitive to both the amplitude of the incoming signal, as well as its phase. A synchronous detector consists of a carrier source and a multiplier.

The voltage appearing at the multiplier output is the instantaneous *product* of the voltages appearing at the inputs. Voltage inputs can be either positive or negative, analog or digital, AC or DC, or a combination of these. It all depends on the application and the specific component(s) used in the multiplier. The voltages used for explanation purposes here, regardless of the actual voltages involved, are scaled to values between +1 volt and −1 volt. Thus, if input A has +1 volt present and input B has +1 volt present, the output voltage will be +1 volt, or +1 × +1 = +1. If input A is +1 volt and input B is −1 volt, the output will be −1 volt, or +1 × −1 = −1. You could also have the combination of 0 volts at input A and +1 volt at input B, resulting in an output of 0 volts; that is, 0 × +1 = 0.

Figure 1 shows one way to implement a synchronous detector. Quite simply, it's an *inverting amplifier and a switch*. The switch-select input is driven with the hard-limited synchronous carrier. The square wave carrier reverses the input signal polarity, alternately multiplying by +1 and −1.

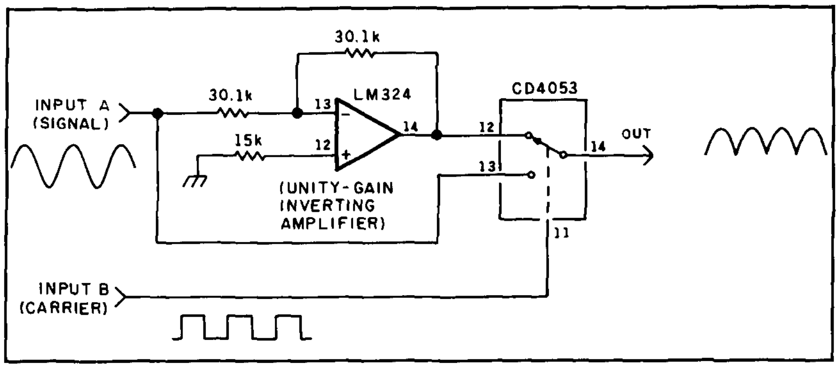


Figure 1. One way to implement a synchronous detector. This circuit works at audio frequencies.

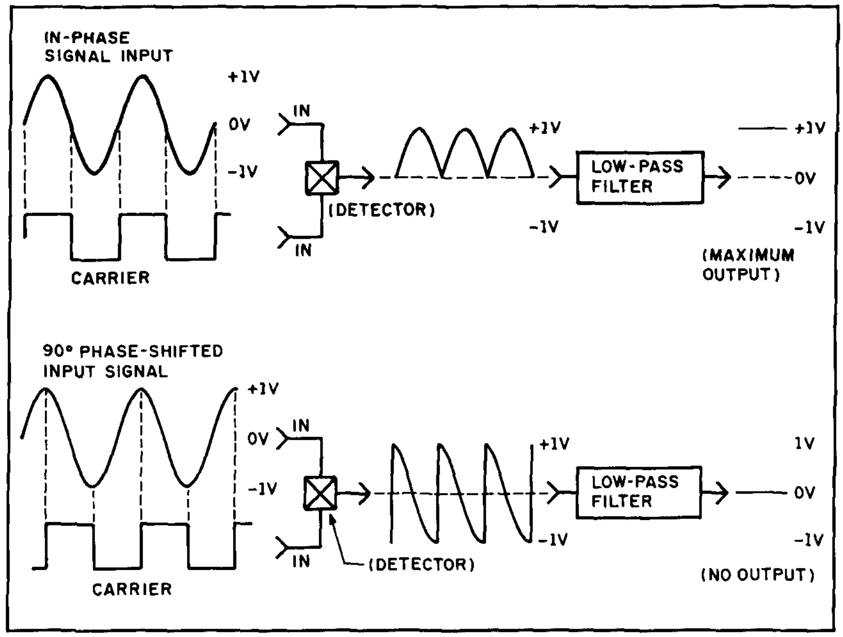


Figure 2. A synchronous detector produces maximum output when the input signal is perfectly in phase with the carrier. It rejects signals which are 90 degrees in phase from the carrier.

Phase shift

Looking at **Figure 2**, you'll note that maximum detector output occurs when the incoming signal is in perfect phase with the carrier. There's no output when the signal is shifted 90 degrees in phase either way from the carrier. Synchronous detection, sometimes called "cross-correlation," offers a 3-dB S/N advantage over envelope (diode) detection. This is because noise tends to be random in phase, spending half as much time (on average) at ±90 degrees phase to the carrier, and producing no detector output. The signal, on the other hand, presumably spends all of its time at 0 degrees phase, producing maximum detector output. The detector in the super-slow system is an exclusive-OR logic gate.

Carrier recovery

The other half of a synchronous detector, the carrier source, is a bit harder to provide. In fact, carrier quality determines the success of the overall detector. The carrier in a receiver must somehow be "recovered" — usually from the incoming signal. Carrier recovery schemes range from simple filtering of the carrier from the incoming signal with a high-Q filter (using filter ringing as an advantage), to complex phase-locked loops. The super-slow hardware I'll describe "cheats" somewhat in that the carriers at both the transmitting and receiving ends of the communication channel are derived from a third, master source.

BPSK

BPSK takes maximum advantage of synchronous detection at the transmit and the receive end. To understand why, consider how a multiplier works. Think of an amplitude shift keyed (AM) signal as a voltage varying from a maximum of ±1 volt (key down), to a minimum of 0 volts (key up).

The output resulting from multiplication with a perfectly synchronous carrier (±1 volt) is a voltage which varies from +1 volt to 0 volts, or −1 volt to 0 volts (if the carrier happens to be 180 degrees out of phase with the incoming signal). This is a total

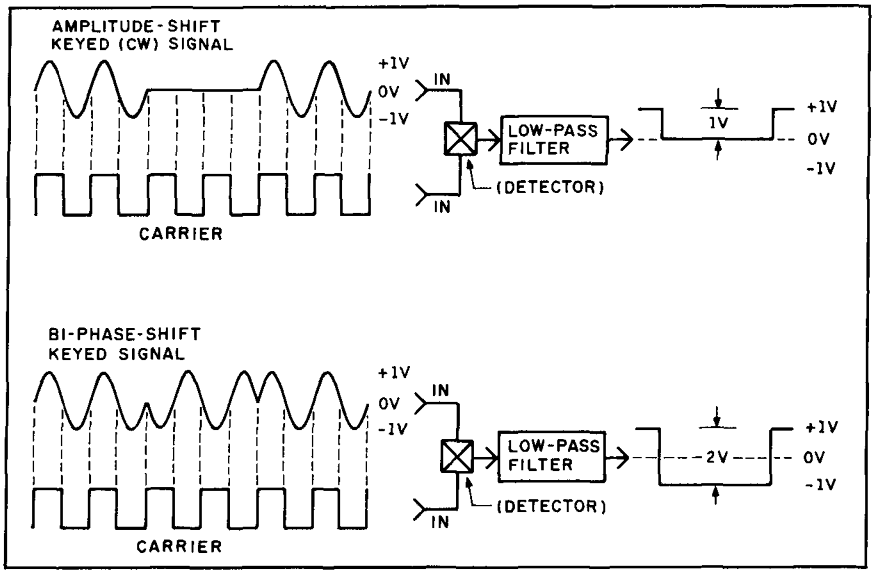


Figure 3. When detected synchronously, BPSK produces twice the detector output voltage as amplitude-shift keying.

output swing of 1 volt. On the other hand, a biphase shift keyed signal doesn't change in amplitude, it only changes phase – from a point in phase with the carrier to one 180 degrees out of phase. After multiplication with the carrier, the resulting output varies from +1 volt to −1 volt. This is a total output swing of 2 volts and a gain of 6 dB. **Figure 3** illustrates this point.

Where does the 6 db come from? It comes from the transmitter. A BPSK transmitter also uses a multiplier. In this case, it's used to generate the phase-shifted signal. An analysis of the multiplication process at the transmitter reveals that transmitter power, instead of being sent as a carrier plus two sidebands (as in amplitude shift keying), is transmitted completely in sidebands. No carrier is sent (see **Figure 4**). Transmitter power is concentrated in the sidebands and fully recovered by the multiplication process at the receiver.

Eliminating the voltage reference

A second, very important reason for using synchronous detection and BPSK, is to relieve the receiver of the need to develop a voltage reference for the mark-space decision after the detector and filter integrator. The output of the detector is *bipolar.* Once synchronous operation is established (the function of the carrier recovery circuitry), the slice point for the mark-space decision is simply *zero volts.* The hardware actually performs this operation with numbers, but the principle is the same. A detector-filter output voltage above zero is called a mark (or logical 1); an output voltage below zero is called a space (or logical 0), or vice versa. You'll find this capability very useful in detecting the infinitesimally small, slowly changing signals you're after. It's a significant improvement over the AM systems that have been tried (namely CCW).

An experimental super-slow system

There's an experimental detecting and filtering circuit which tests the validity of the super-slow approach. However, this circuit needs some fairly elaborate support. So, before I explain the detector-filter, I'll look at the overall system. Because space is limited, I won't show the circuit schematics and associated descriptions for everything. They are available from the sources listed in **Appendix B.**

The "system" consists of a transmitter and a receiver. Block diagrams are shown in **Figures 5** (transmitter) and **6** (receiver). The transmitter consists of a WWVB receiver and master reference generator (MRG), carrier and clock generators, encoder, modulator, power amplifier and antenna. The receiver consists of an antenna, preprocessor (RF and IF amplifiers, mixers, filters, first and second injection generators, and analog-to-digital, or A/D, converter), WWVB receiver and MRG, carrier and bit clock generation circuitry, phase detector-integrator-sampler and decoder.

You should take note of two things. First, both the carrier and signaling clock at the transmit and receive ends are generated from identical 100-Hz master reference sources. Second, both master reference sources are synchronized to WWVB – the time and frequency service provided by the National Institute of Standards and Technology (NIST) on 60 kHz. The service originates from the same transmitter site as WWV, near Fort Collins, Colorado. NIST assures us that WWVB is receivable (local noise conditions permitting) anywhere in the continental United States and Hawaii. The absolute synchronization provided by WWVB is the key to the success of the super-slow system.

The WWVB receiver-MRG is basically a phase-locked loop. A phase comparison is made of a locally synthesized 60-kHz square wave with the amplified, filtered, and limited signal from WWVB. A filtered DC correction voltage is applied back to the MRG to keep it dead on frequency.

Synchronous BPSK transmitter

The transmitter's operating frequency is selected by programing the carrier generator (see **Figure 5**). Dip switches set the division ratio of a programmable divider which, in turn, causes a phase detector-VCO to output the desired multiple of the reference frequency.

The transmit bit clock (or signaling interval) is generated by dividing the 100-Hz reference from the MRG by the selected ratio (10, 100, 1000 or 6000) to produce the desired clock rate. The bit clock dividers can be synchronized either manually with a push button, or from an external sync source like a WWVB time decoder. Normally, the bit clock counter is synchronized to the UTC 10-minute mark; that is, UTC xx:00:00, xx:10:00, xx:20:00, and so on. Clock synchronization is maintained indefinitely after initialization unless WWVB lock is lost for a very long time (weeks), or there's a power failure.

Character clock is derived by dividing the bit clock by 10. Bit and character clocks are supplied to the encoder.

The transmit encoder consists of an eight-bit parallel-to-serial shift register and a J-K flip-flop. The seven-bit ASCII character to be sent is presented to the shift register inputs and clocked out in normal, serial non-return to zero (NRZ), least-significant bit first, using asynchronous character format with a start bit at the beginning of each character frame and two stop bits at the end. The J-K flip-flop then converts the bit coding to non-return to zero-M (NRZ-M) (differential). For details, see **Figure 7**.

The ASCII characters (words) to be sent can be produced in several ways, depending on the way the system is used. For operation in beacon ID-only mode (the mode used 99 percent of the time on the LowFER band), a simple sequentially addressed diode matrix is chosen. For a longer message, or beacon ID plus message, a 2716 EPROM can be used. The 2716 holds eight messages of up to 255 characters each, or one message of up to 2047 characters.

A parallel-computer interface is another character source. (It can be useful in making a contact with another station — an undertaking that could take days at 1/60 baud!) The computer is programmed with the message. It then outputs the appropriate ASCII code. ASCII NULs (all data bits 0) are sent when no characters are available from the computer.

The BPSK modulator is the simplest component in the entire system. It consists only of an exclusive-OR gate. The carrier

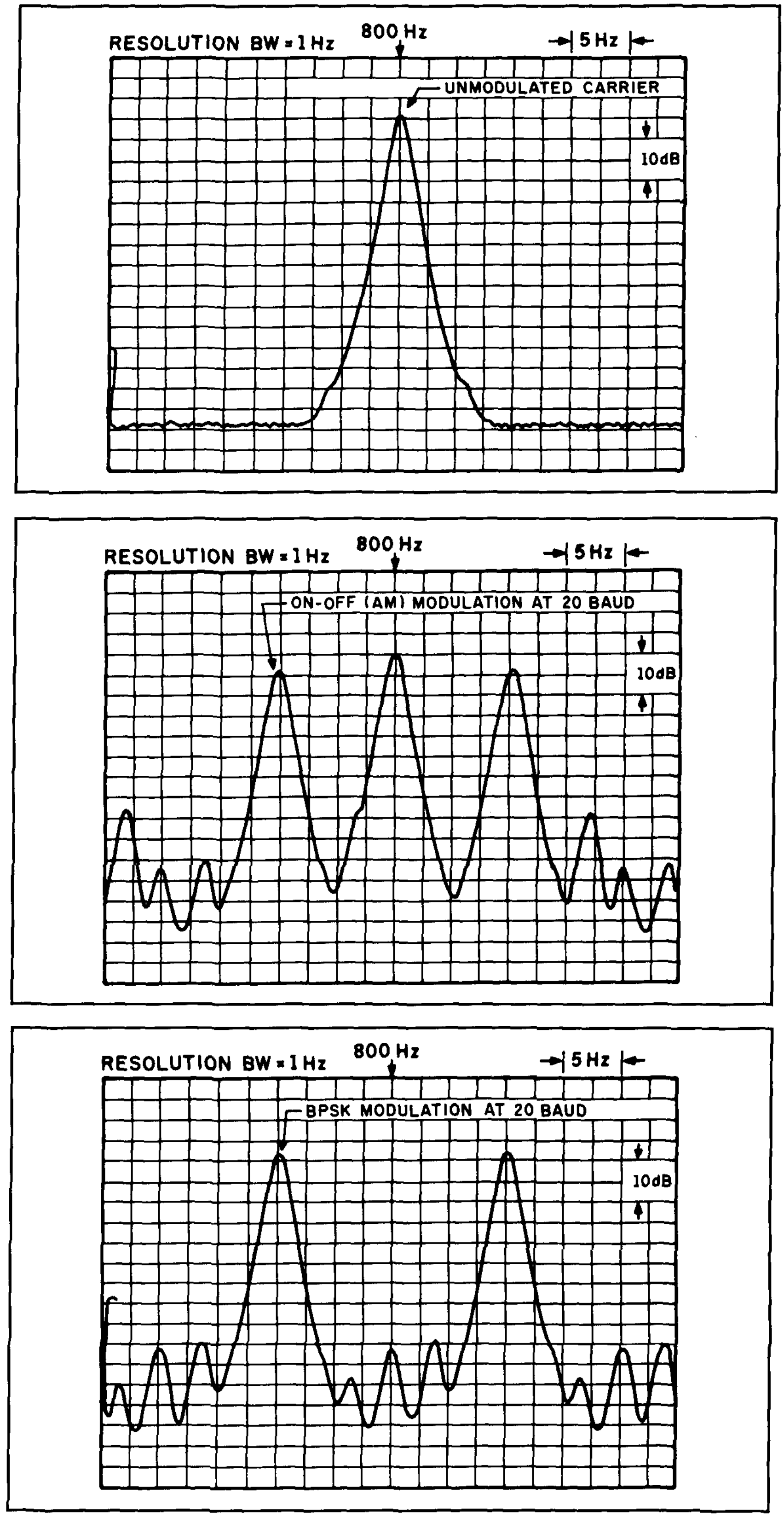


Figure 4. Spectrum analyzer output recordings comparing a) an unmodulated carrier at 800 Hz, b) a carrier modulated with 20-baud symmetrical data (equivalent to a continuous string of dots at about 24 wpm) using amplitude-shift (on/off or CW) keying, and c) BPSK modulation at the same rate. Notice that sideband power is 6 dB higher for BPSK modulation, and that virtually no power is transmitted in the carrier. Peak envelope power (PEP), as viewed on an oscilloscope, is identical in all three cases.

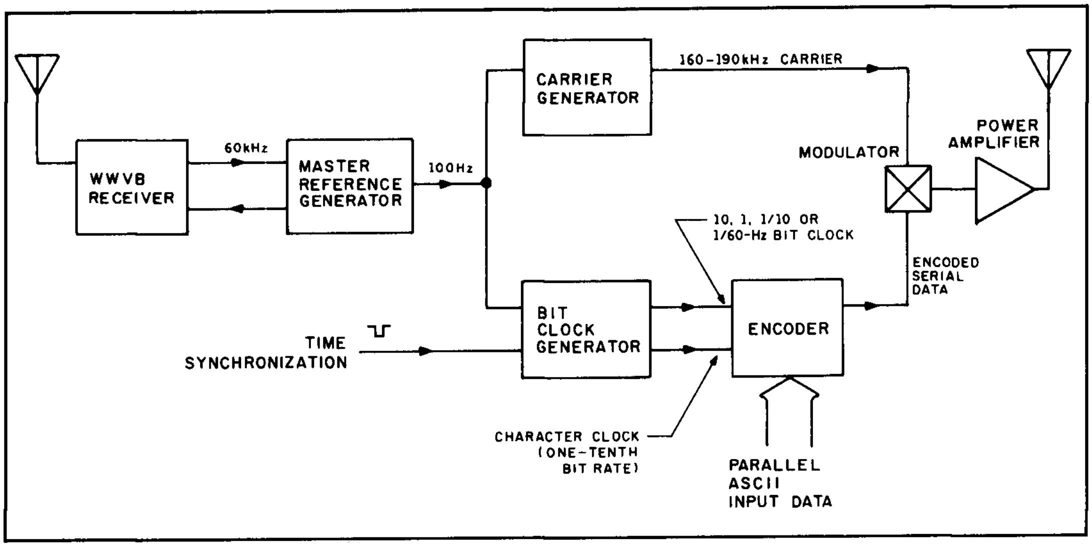


Figure 5. Transmitter block diagram.

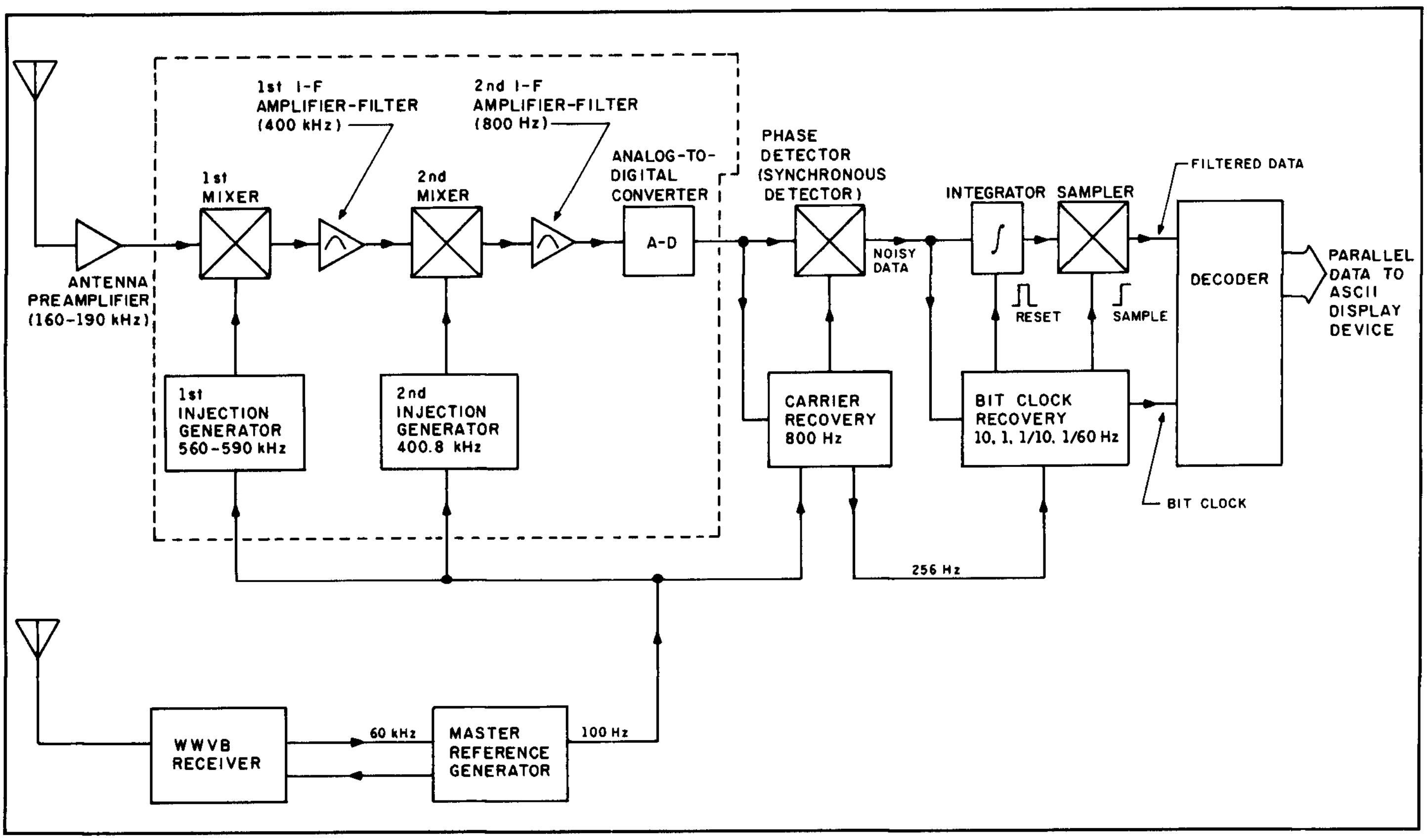


Figure 6. Receiver block diagram. All injection frequencies are synthesized from a reference locked to NIST station WWVB. Circuits shown within the dotted line are lumped under the function "preprocessing" in the accompanying text. As you can see, the preprocessor bears a strong resemblance to an ordinary superhet receiver.

phase is reversed from time to time in step with the serial data from the encoder, as illustrated in **Figure 8**. The truth table for an exclusive-OR gate gives more information about how this is accomplished.

Receiver preprocessing

A certain amount of preconditioning is necessary before super-narrow signal detection, filtering, and decoding can take place at the receive end. This includes interception of the radiated signal, amplification and leveling, conversion to a frequency usable by the detector/filter/decoder, anti-alias (Nyquist) filtering, and A/D conversion. The circuit shown within the dashed lines of **Figure 6** performs these tasks. It appears to be a rather ordinary superhet receiver and, in most respects, it is.

Receiver features

The circuit is more or less optimized for operation in the LowFER band. It uses a 96-inch active whip antenna, miniature IF transformers as RF filters, doubly balanced mixers, ceramic resonator filters in the first IF amp, and a four-section multiple feedback active bandpass filter as the second IF amplifier. The resulting overall antenna-to-detector bandwidth is about 60 Hz. The receiver also features a noise blanker and automatic gain control (AGC).

However, the injection-frequency generators (usually called LO and BFO) are what make this receiver-preprocessor somewhat unusual. The generators use circuitry similar to that of the carrier generator at the transmitter. In each case, a VCO maintains a precise frequency output; that is, a specific frequency in the 560 to 590-kHz range in the first injection generator, and 400.8 kHz in the second (see **Figure 6**). As a result, a *precise* 800 Hz (accurate to 7 or 8 decimal places!) is delivered to the A/D converter. This fact is of great importance for the detection and filtering processes which follow.

A/D converter

The receiver uses an unusual multi-bit A/D conversion process called Delta-Sigma modulation. The D-S modulator **(Figure 9)** is quite simple. It consists of an analog integrator followed by a voltage comparator, which is followed by a D flip-flop. The 800-Hz IF input signal is applied through a coupling capacitor to the integrator and, in turn, to the comparator. The high and low outputs from the comparator are latched into a D flip-flop clocked at 204.8 kHz (800 Hz × 256). The output pulses from the D

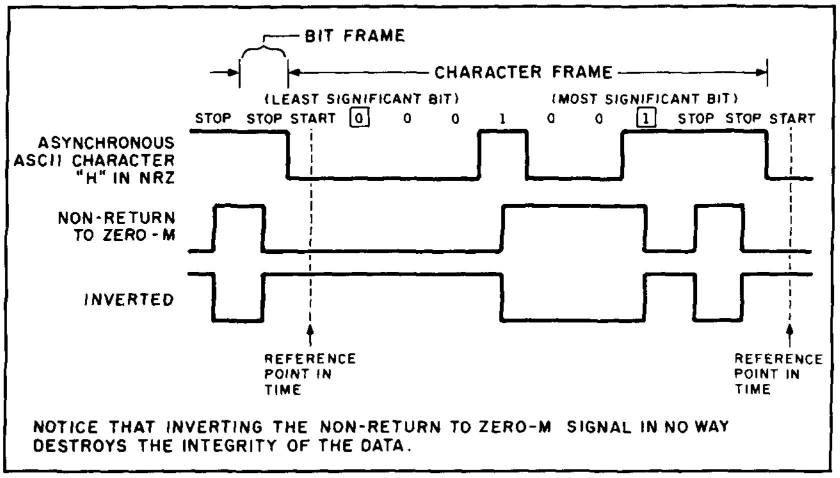


Figure 7. Example of an ASCII character encoded in asynchronous format non-return to zero and non-return to zero-M (differential) bit-coding schemes.

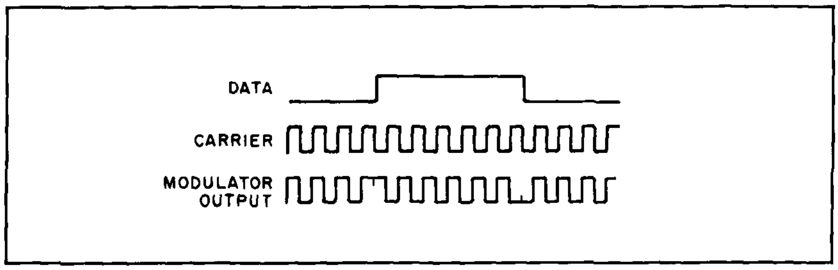


Figure 8. BiPhase modulation.

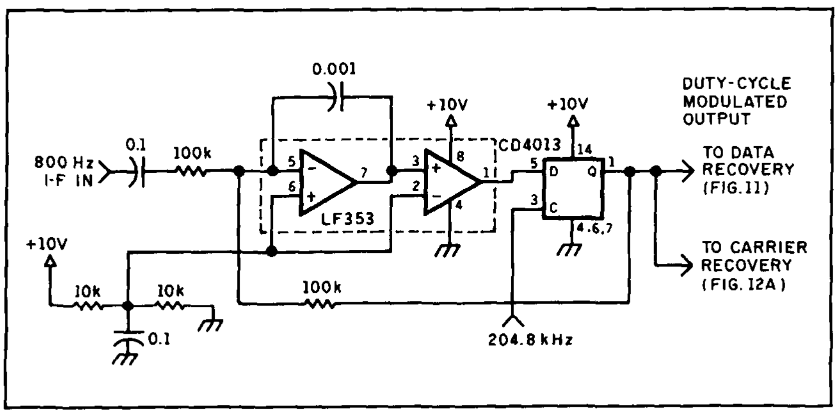


Figure 9. The Delta-Sigma modulator (A/D converter).

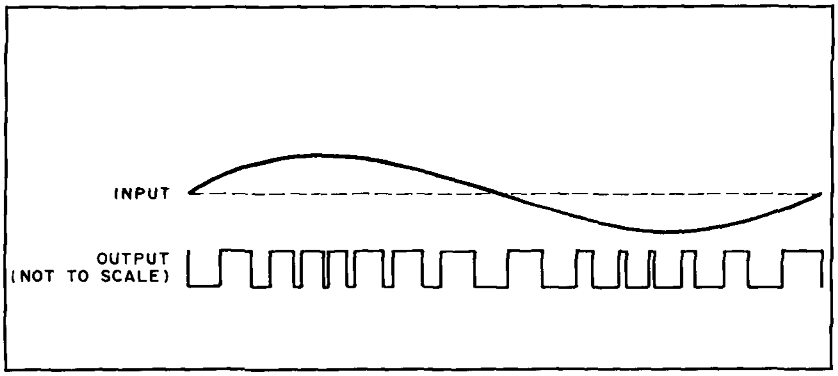


Figure 10. Duty-cycle modulation.

flip-flop are then fed back to the input integrator and added *destructively* to the input signal.

The output pulses, or "charge packets," from the latch will be of a width and polarity sufficient to cancel the input signal exactly. The resulting output pulse train conveys all the phase and amplitude information of the analog input signal in a form of encoding called duty-cycle modulation (see **Figure 10**). With this type of encoding, several signal processing steps can be performed serially using common CMOS digital logic chips.

The detection and filtering processes are divided into three separate but interconnected and interdependent circuits. These are: the synchronous detector and integrate-sample-reset filter, the carrier recovery circuit, and the signaling interval recovery circuit.

Synchronous detector and integrate-sample-reset filter

"Sum of products," the simplest variation of a digital signal processing (DSP) operation, is used to extract the data imbedded in the (presumably) very noisy signal. The duty-cycle encoded digital pulse train from the A/D converter is first applied to the detector (**Figure 11**), an exclusive-OR gate, where it's "mixed" (multiplied) with the locally generated 800-Hz carrier (more about the carrier later). This process "rectifies" the signal to the duty-cycle modulated equivalent of "DC" (baseband). The baseband signal is then applied to the digital integrate-sample-reset filter.

Detector and integrator

The filtering process has three steps: averaging (integrating) the output of the detector, sampling the output of the integrator, and resetting the integrator. Seven cascaded CD4516 four-bit binary up-down counters (in effect, a single 28-bit counter) form the integrator. The output of the detector is connected to the up-down command inputs of the counters. If the output of the detector happens to be high, the counter counts upward (0001, 0010, 0011, and so on). If the output of the detector happens to be low, the counter will count downward (1111, 1110, 1101, and so on). The clocking of the counters takes place at the same 204.8-kHz rate as the sample flip-flop in the A/D converter.

Digital filtering

Noise tends to be random in nature, so,

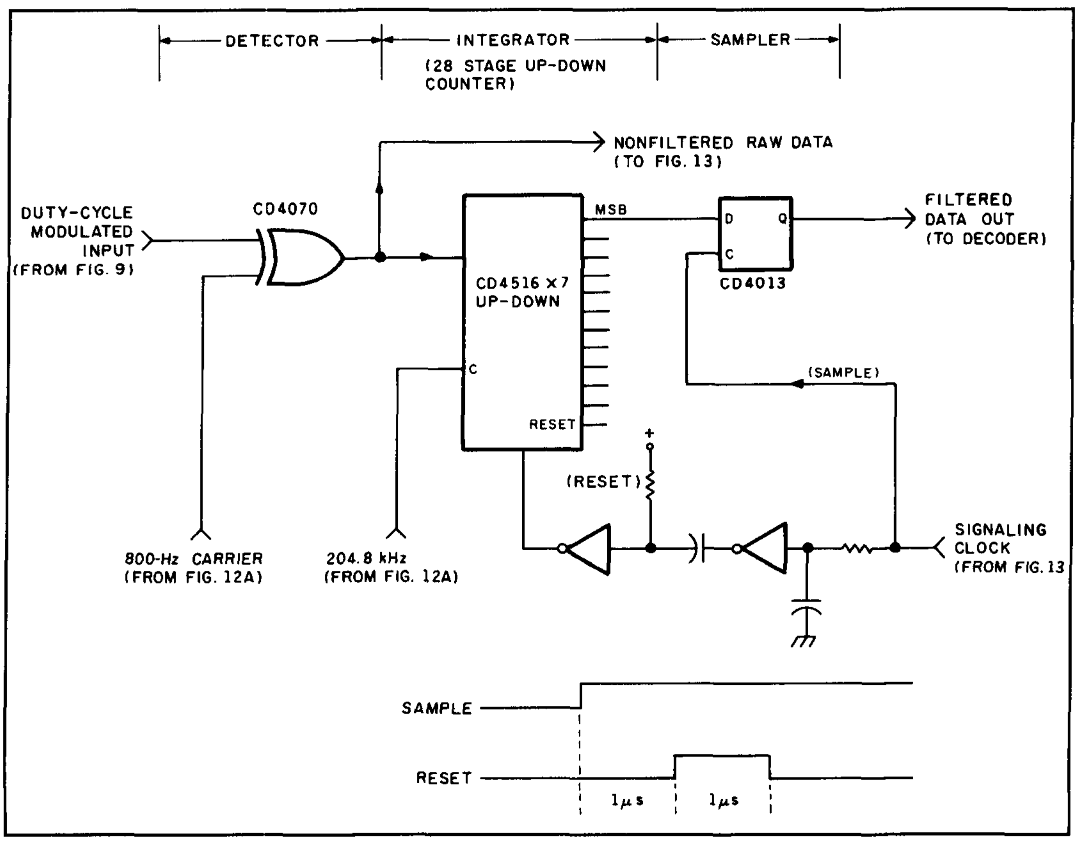


Figure 11. Simplified schematic of the data recovery detector-integrator-sampler.

during a given integration time, it's presumed noise will produce an equal number of up and down counts — just as the flipping of a coin will produce an equal number of heads and tails. The net count will average out to zero. The longer the integration time, the closer to zero the net count will be. On the other hand, a coherent signal will bias the count in either the up or down direction — depending on whether it's in or out of phase with the carrier. The 28-bit counter-integrator can accommodate a maximum count of 2 raised to the 28th power, 134217728 counts up and 134217728 counts down (268435456 counts total), before beginning to repeat. This means the integrator can accommodate a data rate as slow as 134217728 divided by the 204800-Hz clock rate, or 1/655 Hz.

The state of the most significant bit (MSB) from the counter indicates whether the count is in up or down territory. At the end of the integration interval, a microsecond before the counter is reset to 0, the last and most significant stage of the counter chain is sampled by a D flip-flop. The output of the sample flip-flop is the filtered serial data.

Filter system advantages

One of the advantages of this digital filtering process is the total absence of ringing in the output. The filter operates in the time domain, and has what is called an "ideal step response;" it can't ring. Its response in the frequency domain (frequency response) closely matches that of the (sin X/X shaped) frequency spectrum emitted by a BPSK transmitter. It's a "matched" filter and offers maximum S/N benefit.

Flexibility is another advantage. The center frequency depends only on the frequency of the carrier clock. Filter bandwidth depends only on the sample rate. Theoretically, there's no limit to the narrowness of bandwidth. As you'll see, the same circuit easily accommodates all the rates of the superslow system.

Recovering the carrier

A reference carrier source is the major requirement for proper operation of the synchronous detector in **Figure 11**. For carrier recovery, you must reconstruct, in the receiver, the original timing used for the modulation process at the transmitter. This timing information is contained in the *transitions* in the incoming signal. These occur at *twice* the fundamental rate of the carrier, or 1600 Hz. The timing recovery circuit is shown in **Figure 12A**.

Squaring circuit

The duty-cycle encoded pulse train from the A/D converter is first sent through a 64-bit shift register clocked at 204.8 kHz, which delays it for 312.5 μs — the equivalent of one-fourth of a cycle at 800 Hz. The delayed signal is compared (multiplied) continuously with the nondelayed input signal using an exclusive-OR gate. The resulting output is a (duty-cycle encoded equivalent) frequency exactly twice that of the input, or 1600 Hz. This process is called "squaring."

Carrier generator

The 1600-Hz output from the squaring circuit will probably be buried in noise. To extract the carrier from the noise, another sum-of-products operation is performed similar to that of **Figure 11**. The signal is first multiplied by a locally generated 1600-Hz carrier in the phase detector (the second ex-OR gate). This results in rectification of the duty-cycle modulated equivalent of DC. The output of the second multiplier drives the up-down command inputs of another string of cascaded CD4516 counters clocked at 409.6 kHz (1600 Hz × 256).

Up-down counters

The counters integrate the output of the phase detector. Noise, being random in phase, will presumably produce equal numbers of up and down counts which, when averaged over a long enough period of time, will result in a net count of 0. (Because there's no reset function involved in the operation of this integrator, the net count could average to any number within the range of the counters. The point is, the count will average to some equilibrium number.) Any coherent incoming signal will tend to bias the count to some *specific* number. That number represents the phase angle difference between the incoming signal and the locally generated 2X (1600 Hz) carrier.

The integrator output controls the carrier-generator phase, and because its output is fed back to the carrier-phase detector, the counter will tend to drift to a count which produces equilibrium. When in equilibrium, the carrier generator will produce a frequency of exactly 1600 Hz — offset exactly 90 degrees from the incoming signal. The two signals are then said to be "in quadrature" (see **Figure 2**).

Time to reach quadrature

The length of time it takes to reach quadrature depends on the length of the counter, and the signal-to-noise ratio. The greater the noise, the longer it will take. Because the

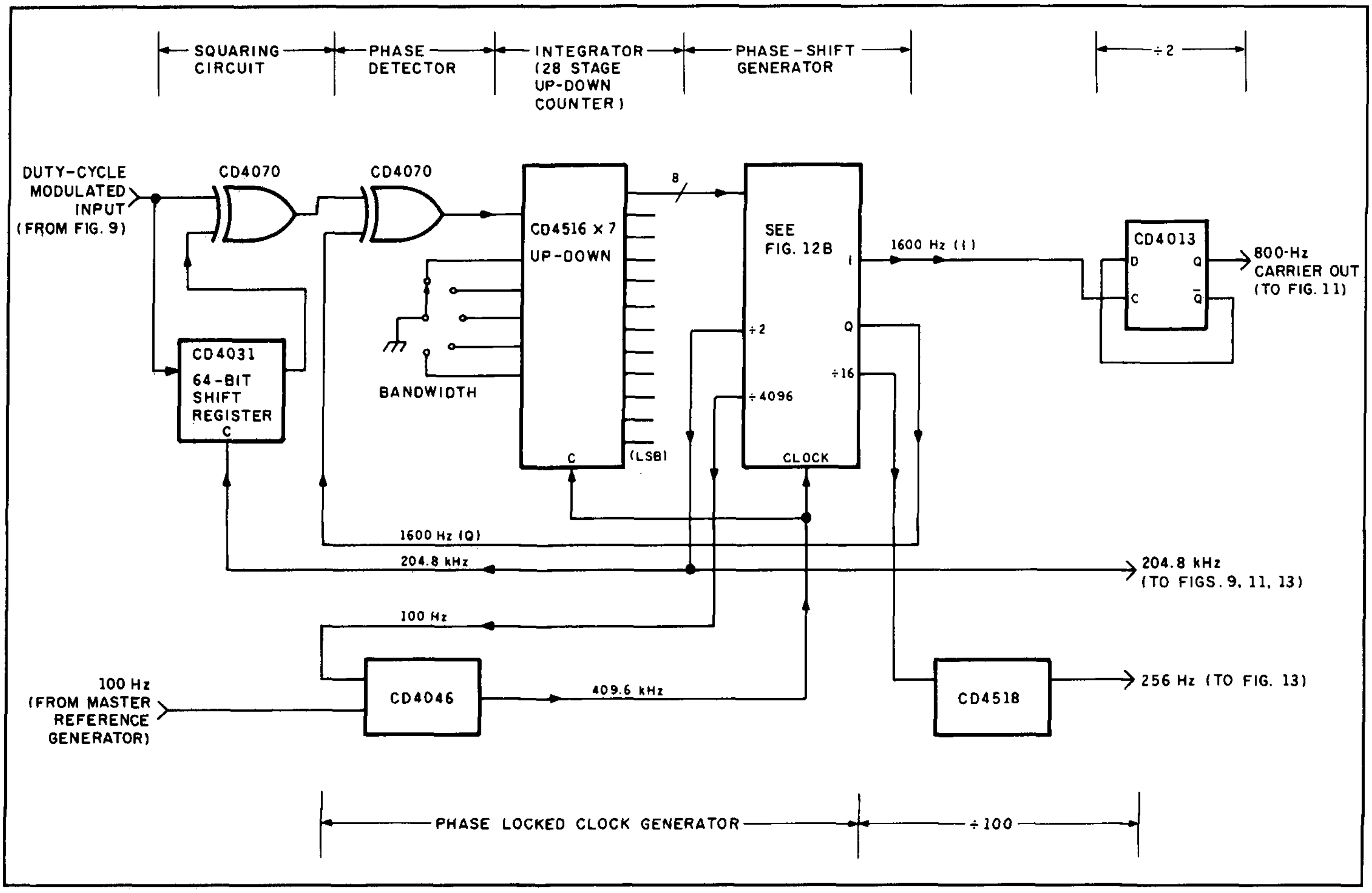


Figure 12A. Simplified schematic of the carrier recovery circuit.

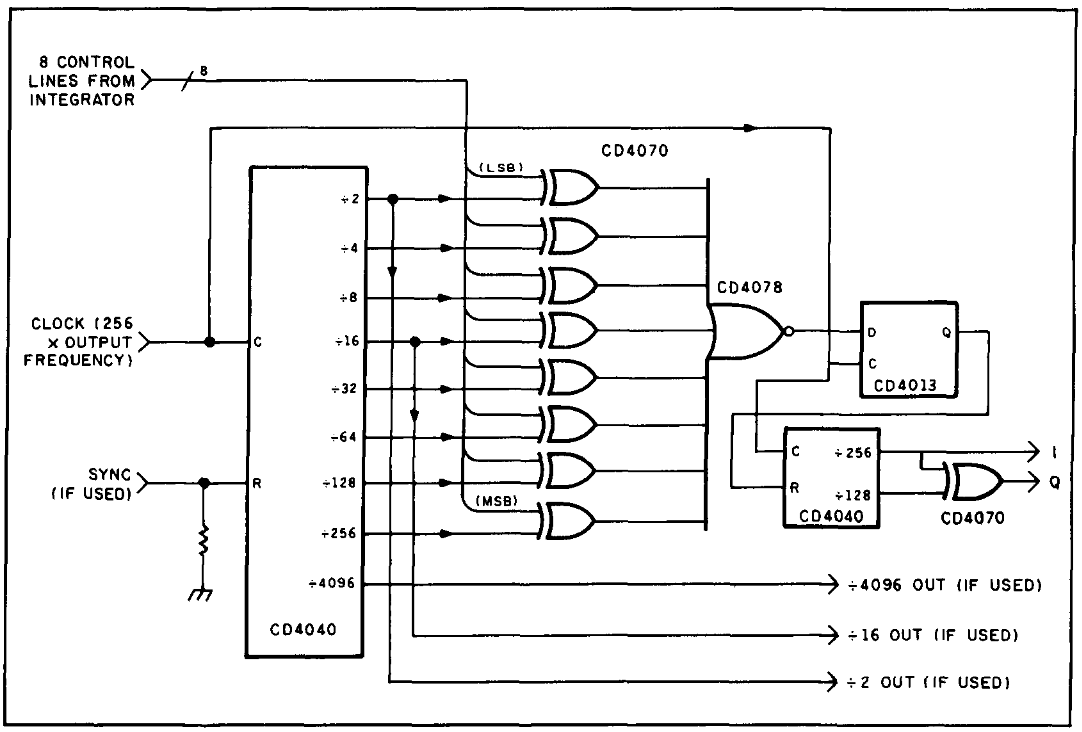


Figure 12B. The digital phase-shift generator. (Used in *Figures 12A* and *13*).

seven cascaded counters have a maximum count of about 268 million (2exp28), reaching quadrature could take as long as 5-1/2 minutes at 409600 counts per second if the input signal is noise free. It will take even longer if it's not (the usual case). The count can be shortened as needed, using the rotary bandwidth switch. This switch disables the counter chips in succession to widen the circuit bandwidth.

Only the eight most significant output lines from the counter are used. This results in a smoothing, or filtering, of the noise-induced jitter in the count. Though the early counter stages may be counting up and down furiously, the *average* count, as represented by the eight most-significant bits, can be quite stable. The output lines of the counter are directed to the digital phase shifter, which is part of the 1600-Hz carrier-frequency synthesizer.

Synthesizer

The synthesizer consists of a CD4046 clock oscillator (phase-locked to the receiver's MRG) operating at 256 times the 1600-Hz carrier frequency (409.6 kHz), a CD4040 ripple counter **(Figure 12B)** which generates 256 discrete phases from the clock, a bank of eight CD4070 ex-OR gates to select one of the generated phases, and two CD4013 D flip-flops to generate the 1600-Hz comparison signal.

The counter steps the synthesizer until phase quadrature of the 1600-Hz output is achieved. Because the synthesizer can only generate discrete phase steps, it can't be in absolute quadrature with the incoming signal at every instant. But, because the carrier output of the circuit is a bit jittery, the *average* phase of the generated carrier will be obtained. The jitter in the carrier is simply

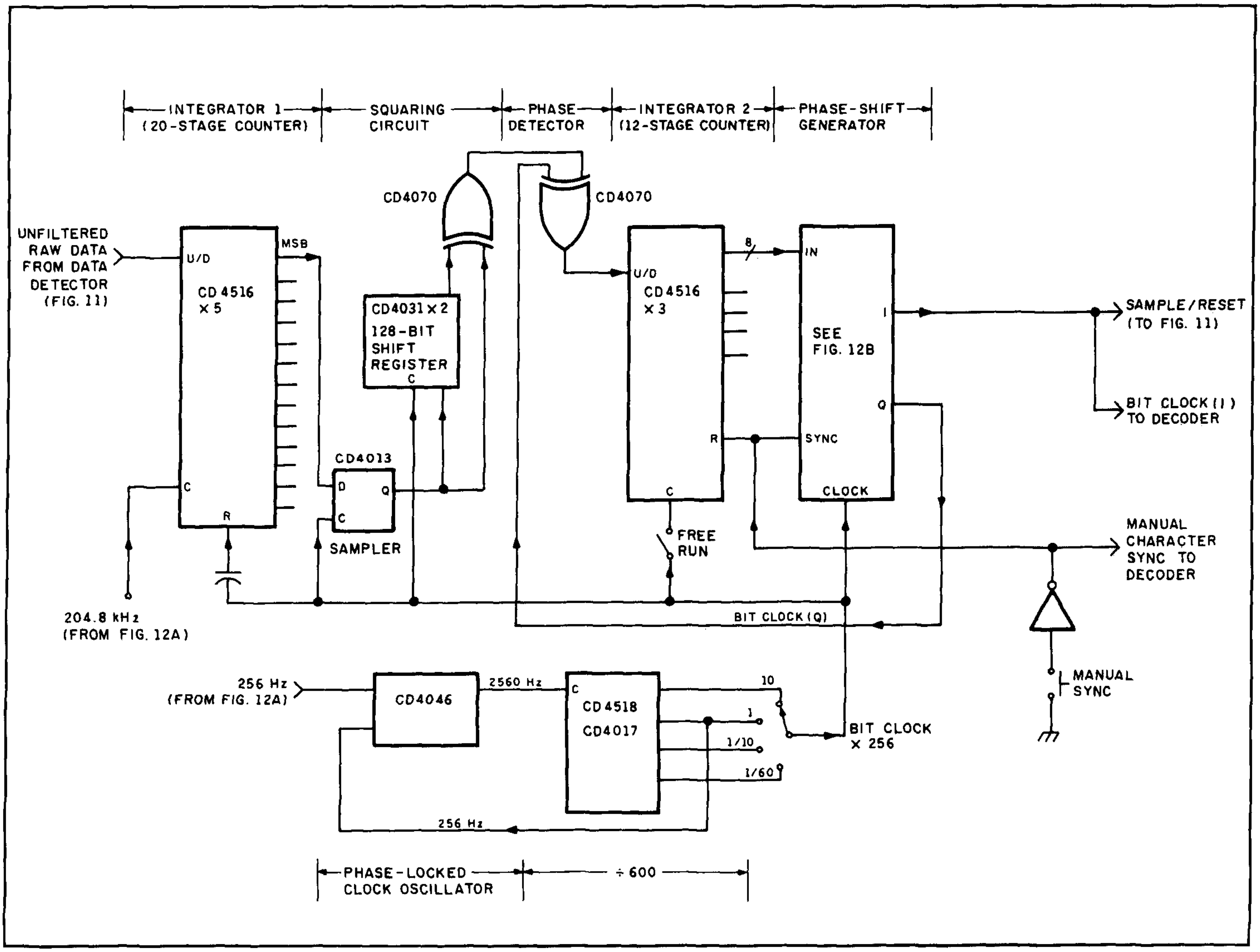


Figure 13. Simplified schematic of the signaling interval timing (bit clock) recovery circuit.

filtered out, as is any other noise, by the integrate-sample-reset filter following the data detector shown in **Figure 11**. Finally, the phase-adjusted 1600 Hz is divided by 2, and delivered to the data detector.

Recovering signaling-interval timing

There's one other signal the super-slow receiver must recover. It must retrieve signaling interval timing, or bit clock, at a frequency equal to the signaling rate. Because the super-slow transmission standard requires that the bit clock be coherent with the carrier (see **Appendix C**), providing the bit clock at the receiver is simply a matter of dividing the 100-Hz clock from the MRG by the appropriate ratio. For example, 10 baud would necessitate dividing 100 Hz by 10.

Bit-clock transitions

Having the bit clock at the correct frequency isn't enough. Bit-clock transitions must occur at the correct *time* as well. Two methods for synchronizing bit-clock generation are used in the receiver — internal and external-manual. Internal synchronization uses the receiver's automatic synchronizing circuitry to extract the timing directly from the signal. In external-manual synchronization, the synchronizing pulse is input manually through a push button from a source of external timing like WWV, or an accurate clock.

Automatic synchronization can be used at all rates in the super-slow system, but it's more appropriate for the faster rates (10 and 1 baud). Manual synchronization might be a better choice for the slowest rates (1/10 and 1/60 baud). The manual sync pulse is ordinarily supplied at one of the UTC 10-minute marks (see standard no. 7, **Appendix C**). You can opt to disable synchronization, allowing the bit-clock generator to free run, isolated from noise hits and signal dropouts. The receiver can go for days in the free-run mode and still be in perfect synchronization.

Signaling interval timing-recovery system

The bit-clock generation circuit is shown in **Figure 13**. It uses two separate integrators. Integrator 1 receives raw, duty-cycle modulated data from the data detector of **Figure 11**, integrates and synchronizes it to the slower clocking rate of the following delay register, and strips it of some noise and clocking components. The output of the first integrator sampler is delayed and squared, multiplied with the locally generated bit clock, and applied to the second integrator. This is where final noise filtering takes place. It works in exactly the same way as the carrier-recovery integrator in **Figure 12A**, except it's slower.

Noise drives the counter up and down randomly, while a coherent clock signal tends to drive it toward a specific count. The counter's eight most-significant output lines drive a digital phase-shift generator whose output is fed back to the input comparator. The phase shifter's output is the recovered bit clock. The bit clock generates properly timed sample and reset pulses for the data integrator in **Figure 11**, and is supplied, along with the recovered data from **Figure 11**, to the decoder.

Versatile bit-clock generator

A CD4046 PLL chip and some dividers supply clocking for the bit-clock generator at 256 times the bit-clock rate. The bit rate is selected by setting the four-position switch to the appropriate position. With a little imagination, other rates can be accommodated easily. As a matter of fact, the bit-clock recovery circuit will work with *any* rate from about 100 baud (or faster if faster ICs are used) down to virtually DC simply by supplying the X256 clock frequency.

Decoding

The decoding process first converts the NRZ-M (differential) bit-coded serial data stream from the data-integrator output sampler (**Figure 11**) into serial NRZ, and then to seven-bit parallel format. One of the reasons for using the ASCII code is because there are many devices available which directly convert parallel-format ASCII to a readable form. The most common is a parallel-interfaced computer printer, or a computer with a parallel-interfaced I/O port. There are also several small LCD and LED devices which accept a direct parallel ASCII input.

Two versions of the decoder have been built: microcomputer and CMOS. The microcomputer version uses a programed D8748HD microcomputer chip (the program was written by LowFER Mark Mallory) which drives a 32-character LCD, parallel printer, and RS-232 serial output.

The program has the ability to find the character-framing information in the data stream even under noisy conditions. In effect, a special auto-correlation filter algorithm stores the data and continually looks for the characteristic "110" stop-start pattern — making its decision to correct character framing on the basis of a long-term average. This is possible because the transmitter keys the characters *coherently* or *synchronously*; that is, one character immediately follows another, with no intermediate "dead air." (See transmission standard no. 8, **Appendix C**.)

The CMOS decoder uses a handful of CMOS parts in a circuit reminiscent of the circuits presented thus far. The circuit drives a four-character LED display and parallel-printer output. The CMOS version has only some rudimentary character-framing logic to detect the stop-start bits, but does feature external-manual character synchronization capability for 1/10 and 1/60 baud. The synchronizing pulse comes from the circuit shown in **Figure 13**.

Both decoders have free-run capability because, once the position-in-time of the characters has been determined, there's no reason to keep looking for start and stop bits. As with the bit-clock generation circuits, the decoder(s) can free run indefinitely and still remain in perfect synchronization. Both decoders suppress any inadvertently received ASCII control characters, which can interfere with a computer printer when the data is noisy.

My version of the super-slow receiver (shown in **Photos A** and **B**) uses both decoders running in parallel.

Photo A. The super-slow receiver. Two separate decoder/displays are used. At the upper center of the panel is a 32-character LCD. To the right is a four-character LED display. The two decoder-displays operate in parallel. The meter monitors AGC voltage and outputs of the various phase detectors, as well as crystal-heater and battery voltages. This particular receiver can be set to four different channels (out of 3000 possible).

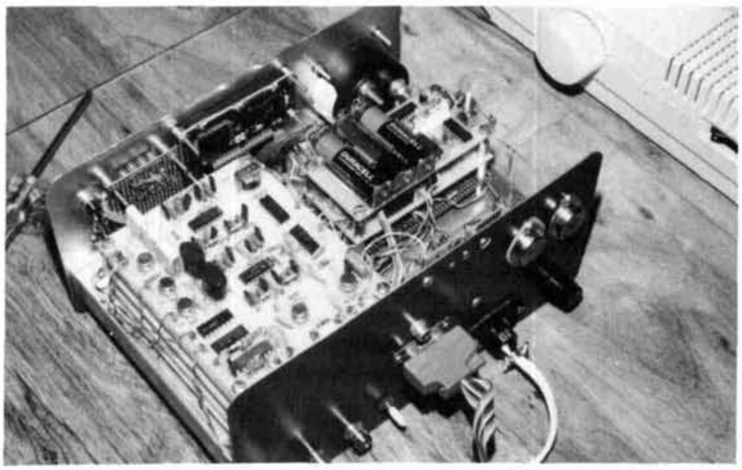


Photo B. Interior of super-slow receiver. No MMICs, striplines or GaAsfets here!

Conclusion

Routine Part 15 ground-wave communication to 1000 miles on 1750 meters has yet to be accomplished. The only two compatible receivers known to exist are in Salt Lake City, Utah and Wheatland, Wyoming. They are separated by a distance of 375 miles. Virtual 24-hour ground-wave reception at 10 baud between these two points seems to be the norm from about October through March. This tapers off gradually to about 6 hours per day (6 a.m. to 12 noon) during the summer months. Reception has been tested at 550 miles at 1/10 baud for a number of hours. But unless more receivers are built, attainment of the 1000-mile goal will have to wait until one of the present receivers can be carried out to that range.

In the meantime, there's a test signal on the air in the 1750-meter band for anyone who wishes to experiment with these techniques. The beacon, which originates from Wheatland, Wyoming, conforms to all the parameters set forth in **Appendix C**. Carrier frequency is 175.250 kHz and signaling rate is 1/60 baud. The beacon sends the word MAX<SP>, continuously.

The transmitting antenna is a 50-foot base-loaded vertical with a combination cage and umbrella top hat over a 50 by 50-foot chicken-wire ground screen (see **Photo C**). The loading coil is a "basketweave" of 28 by 10 litz wire, 12 inches in diameter and 5 inches high. Final DC input power is 1 *watt* (8 volts at 120 mA). The antenna base current is about 175 mA, and radiated power is something on the order of 1 mW. The receiving antenna system is shown in **Photo D**.

If estimates of the 1000-mile range for the system are correct, the tiny signal from this beacon should cover over 3 million square miles – the western two-thirds of the United States, plus a good chunk of Canada.

If you're interested in 1750-meter experimentation, you might like to pursue three possible routes of evolution for the super-slow system. As always, the goal is to make easier to build equipment more widely available to the average LowFER. We need experimenters to produce printed circuit boards for the present circuitry, to work on reducing the complexity of the present circuitry with dedicated microcomputers (like the one now used for decoding), or to go for broke and build a total "receiver in a computer" using DSP techniques.

Those accustomed to routine transmission of data at high speeds over phone lines or VHF/UHF radio – both relatively noise-free media – may not appreciate the special challenge of doing it *slowly* via noisy LowFER radio. However, before you get too involved in building circuitry or writing programs, I suggest you read about conventional long wave transmitting and receiving methods. It's a different world down there! ■

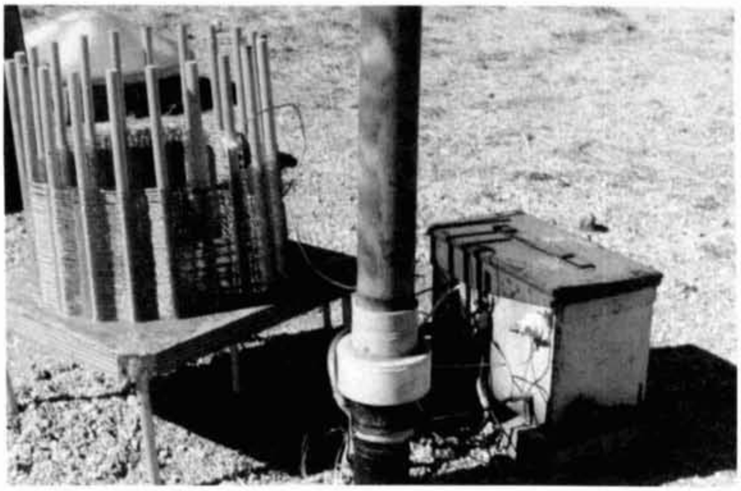


Photo C. Power amplifier, base insulator, and loading coil at the base of the transmitting tower. The coil is about 100 turns of 28/10 litz wire. About 75 turns are being used at present. A 50″ × 50″ mat of chicken wire lies just below the gravel. The tower is 50′high.

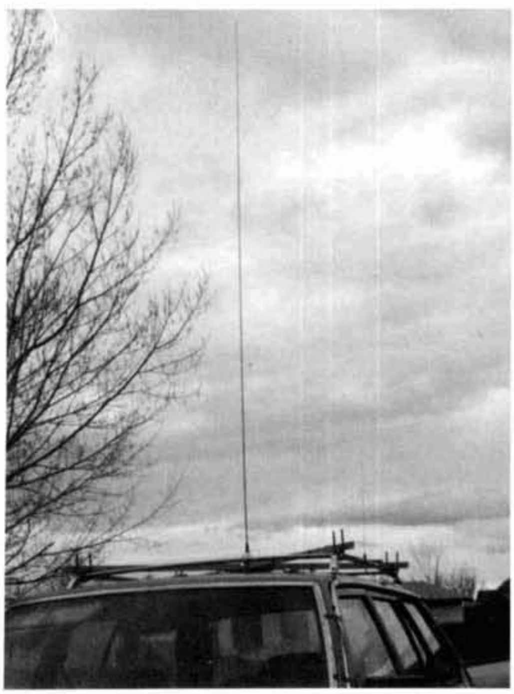


Photo D. The receive antennas used for DX testing. The large whip is used for the main receive frequency (160 to 190 kHz). The short antenna receives WWVB (60 kHz). Each antenna has a high impedance FET preamp at its base. You *should not* park this under a power line if any DX is expected. The system will thrive, however, parked on the floor of a deep mountain canyon, receiving its BPSK modulated long-wave signals from 250+ miles in the daytime. At such locations, even the AM broadcast band is dead! Daytime reception has been demonstrated at 550 miles.

REFERENCES

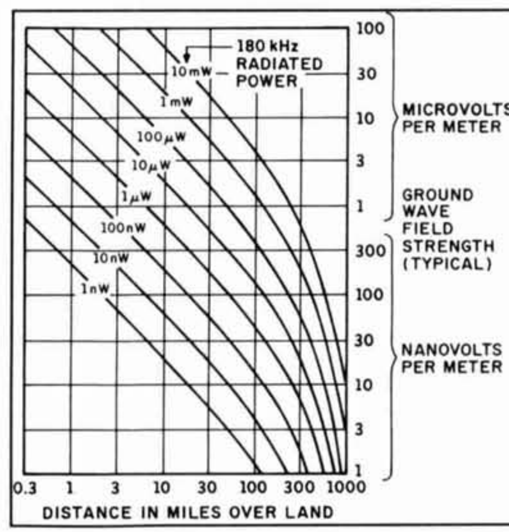
1. Raymond C. Petit, W7GHM, "Coherent CW–Amateur Radio's New State-of-the-Art?," *QST*, September 1975.
2. Adrian Weiss, W8EEG, "Coherent CW," *CQ*, June and July, 1977.

Appendix A—How far does a LowFER signal go?

By Dave Johnson "DJ," Los Banos, California.

How far down a LowFER signal goes depends on the amount of power transmitted, how much the signal is attenuated as it propagates, the noise conditions at the receiver, and the character of the receiver itself.

The graph in this appendix shows the typical ground wave field strength for propagation over land. Daytime long wave propagation is normally ground wave. At night, sky wave often provides much stronger signals at distances beyond about 150 miles. At distances over 100 miles, propagation over sea water is much better than it is over land.



At distances of less than 100 miles, doubling the range of a transmitter requires quadrupling the radiated power – assuming that the conditions at the receiver are the same. But to get from 500 miles to 1,000 miles, you'll need an approximate 300:1 increase in power. This explains why so many LowFER beacons "get out" 50 to 150 miles, yet so few get past 200 miles during daytime.

The graph also illustrates that major advances in receiver setups will make a big improvement in reception distance for weak beacons, but won't have much effect on the distance at which the stronger beacons can be received. This applies to daytime ground wave reception. Nighttime propagation follows a different set of much more complicated rules.

Judging by the daytime ground wave reception distances being reported for beacons whose radiated power is known to some degree, it appears that the fancier receiving setups are able to pull in CW signals down to around 200 nanovolts per meter.

Coherent receiving techniques combined with integration to achieve ultra-narrow bandwidths could probably push the "floor" down to 10 nanovolts per meter or less.

Appendix B—Interested in 1750?

This article can't begin to cover all aspects of LowFERing. If you want to learn more, there are several publications available which should answer any questions you might have:

- *Transmitting Antennas and Ground Systems for 1750 Meters*, Michael Mideke, Editor. Send $5 to: Michael Mideke, PO Box 123, San Simeon, California 93452.
- *The Low and Medium-Frequency Radio Scrapbook*, Ken Cornell. Send $17 to: Ken Cornell, 225 Baltimore Avenue, Point Pleasant Beach, New Jersey 08742.
- Ralph Burhans has reprints of several of his articles on long wave receiving antennas. Write to: Ralph Burhans, 161 Grosvenor Street, Athens, Ohio 45701.
- Schematics and descriptions for all of the circuits of the super-slow system are available (in draft form) for a self-addressed 10 by 13-inch manila envelope with $2 postage and $4 to cover copy costs from: Max Carter, 46 14th Street, Wheatland, Wyoming 82201.

There are also several LowFER newsletters. Reading these should eventually answer every question you could possibly have:

- "1750 Meters: Western Update," available for 12 SASEs and $10 from: Jim Ericson, 226 Charles Street, Sunnyvale, California 94086.
- "The Northern Observer," $10 (US checks OK) and 10 unstamped SASEs, from: Herb Balfour, 91 Elgin Mills Road, West, Richmond Hill, Ontario, L4C 4M1, Canada.
- "The Lowdown" (covers all activity below 530 kHz including beacon DXing, QSLing, and so on), $12 for 12 monthly issues from: Longwave Club of America, 45 Wildflower Road, Levittown, Pennsylvania 19057.

There's also a 24-hour on-line bulletin board: LongWave DataBase System, phone: 703-528-7753; speed: 300/1200/2400 BPS (MNP4); protocol: 8 data, 1 stop, no parity.

Appendix C—Transmission Standards

In keeping with the necessity for absolute agreement between the transmit and receive ends of the super-slow LowFER system, the following transmission standards have been established:

1. Operating (carrier) frequency is to be a multiple of 10 Hz between 160 kHz and 190 kHz, or 160.010, 160.020, 160.030, and so on.
2. Transmitter frequency *and* signaling clock accuracy are to be maintained at all times (synchronized) to an internationally recognized standard for time and frequency such as WWVB, LORAN-C, Omega, or another primary standard, like a cesium-beam clock (that last one'll cost you about $25K). Meeting this requirement will automatically assure that the carrier and bit clock will be coherent relative to one another.)
3. Signaling rate is to be one of the following: 10, 1, 1/10, 1/60 baud, or anything else agreed upon by the transmit and receive ends.
4. Modulation is to be 180-degree phase-shift keying (BPSK).
5. Bit encoding is to be NRZ-M (differential). Mark (logical 1) is to be represented by a 180-degree carrier phase shift; space (logical 0) is represented by no shift.
6. Character coding should be seven-bit ASCII, least-significant bit first, preceded by a space (start bit) and followed by two marks (stop bits), for a total character length of ten bits. (In all likelihood, ASCII is the code your computer uses to talk to your printer.)
7. When transmitting at 1/10 or 1/60 baud, the character-start bit is to occur as close as possible to each UTC 10-minute mark (point in time). This automatically requires that subsequent bits in each character be sent at the UTC 10-second marks for 1/10 baud and the UTC minute marks for 1/60 baud.
8. Character transmission is to be fully synchronous. This requires that you send *ASCII NULs, plus the start and stop bits* (0000000011), when no data bits are available for transmission.

Figures 7 and **8** should clarify biphase modulation and the bit and character formats of the LowFER super-slow system.

By Thomas V. Cefalo, Jr., *WA1SPI*
29 Oak Street
Winchester, Massachusetts 01890

BASIC CONCEPTS OF SCATTERING PARAMETERS

S-parameters simplify network analysis

Network parameters describe the behavior of a network by characterizing the input and output ports. One network description, scattering parameters (often called S-parameters), was, until recently, known and used only in the microwave engineering community. Knowledge of the usefulness of scattering parameters has become so well known that they are used not only in microwaves, but throughout the entire RF field. S-parameters can be directly applied to a Smith chart. These parameters contain useful information on impedance, VSWR, stability, gain, and isolation. S-parameters have reached such a level of importance that many manufacturers supply this data with their products.

Traditional network parameters

The H, Y, and Z are other, more commonly known, network parameters. Like S-parameters, they are used to characterize networks. What are the advantages of using S-parameters over H, Y, and Z-parameters? For one thing, the voltage and current of each port is required to characterize a network when using H, Y, or Z-parameters. And while it may appear simple to determine this voltage and current, there are many problems associated with this task. **Equations 1-3** in **Figure 1** show the linear equations representing the H, Y, and Z-parameters for a two-port network.

H-Parameters: $$V_1 = h_{11}I_1 + h_{12}V_2 \quad \text{(1a)}$$

$$I_2 = h_{21}I_1 + h_{22}V_2 \quad \text{(1b)}$$

Y-Parameters: $$I_1 = y_{11}V_1 + y_{12}V_2 \quad \text{(2a)}$$

$$I_2 = y_{21}V_1 + y_{22}V_2 \quad \text{(2b)}$$

Z-Parameters: $$V_1 + z_{11}I_1 + z_{12}I_2 \quad \text{(3a)}$$

$$V_2 = z_{21}I_1 + z_{22}I_2 \quad \text{(3b)}$$

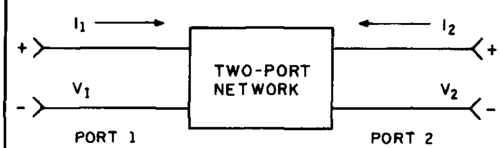
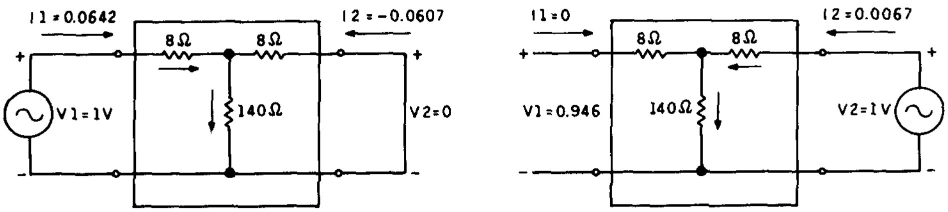


Figure 1. Linear equations representing H, Y, and Z-parameters for a two-port network.

Looking at these equations, you'll see that each of the parameters is a function of voltage or current. To solve for any one parameter, either a voltage or current is set to zero. For example, V_2 is set to zero when measuring y_{11} or in the case of Z-parameters, I_2 is set to zero to obtain z_{21}. You set the variables to zero by shorting (short forces $V=0$) or opening (open stops current flow $I=0$) a network's port. You can see this more clearly if you calculate the H-parameters of the T-network shown in **Figure 2**, where h_{11} = input impedance and h_{21} = forward current gain (with the output port shorted), while h_{22} = output admittance and h_{12} = reverse voltage gain (with the output port open-circuited).

This method of providing open and short circuits may be appropriate at low frequencies, but as the frequency is increased, you must deal with a new set of problems mainly related to stray inductance and capacitance. Because inductance and capacitance are functions of frequency, the idea of affirm-



$$h_{11} = \frac{V1 - h_{12}V2}{I1} = \frac{1 - h_{12}(0)}{0.064} = 15.57\Omega \qquad h_{22} = \frac{I2 - h_{21}I1}{V2} = \frac{0.0067 - h_{21}(0)}{1} = 0.0067$$

$$h_{21} = \frac{I2 - h_{22}V2}{I1} = \frac{-0.0607 - h_{22}(0)}{0.0642} = -0.945 \qquad h_{12} = \frac{V1 - h_{11}I1}{V2} = \frac{0.946 - h_{11}(0)}{1} = 0.946$$

Figure 2. Solving the H-parameters for a T network. Problems arise as frequency is increased. Stray inductances and capacitances make ideal open or short circuits difficult to obtain.

ing a short or open circuit becomes impractical – especially over a wide frequency range.

Another problem occurs when the device under test doesn't operate properly when terminated into a short or open circuit. Active devices may become unstable and oscillate. In some cases, the device can be destroyed by the short circuit termination. Some passive networks, like matching networks or filters, are designed to operate into a specified impedance, and will exhibit a completely different response. Therefore, take care before measuring a device to ensure it can safely and properly operate under these conditions.

Advantages of S-parameters

You can use S-parameters to characterize a network and avoid the problems associated with traditional parameters. S-parameters use traveling waves rather than total voltage and current. While the use of traveling waves may not be obvious at this point, they are the basis of S-parameters and provide many advantages.

When using S-parameters, a network is terminated with a resistive load which eliminates the need for short and open circuits. It's much easier to maintain a resistive impedance over a wide range of frequencies than a short or open circuit impedance. The useful range of a resistive termination extends well into the GHz range.

The impedance of most measuring systems is usually 50 or 75 ohms. Therefore, a network being measured is terminated into the same impedance. Under these conditions, the probability of an active device becoming unstable is greatly reduced. Because the network isn't terminated into a short circuit, the chance the device will be destroyed also declines.

The magnitude and shape of a wave propagating along a low-loss transmission line remains constant (that is, there's no loss or distortion). Consequently, when using S-parameters on a device which is connected to a low-loss transmission line, you can measure the device at a distance remote from the measuring system.

Transmission line theory

Because S-parameters are built around traveling waves, it's important to understand some basic principles of transmission theory. A transmission line may be defined as a medium for carrying electromagnetic waves from point A to point B. The equivalent circuit of a uniform transmission line is shown in **Figure 3**.

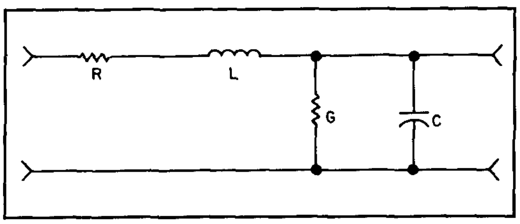


Figure 3. S-parameters are built around traveling waves. The equivalent circuit of a uniform transmission line — a medium for carrying electromagnetic waves from point A to point B — is shown.

The characteristic impedance of this transmission line is given in **Equation 4**.

$$Z_o = \sqrt{\frac{R + jwL}{G + jwC}} \qquad (4)$$

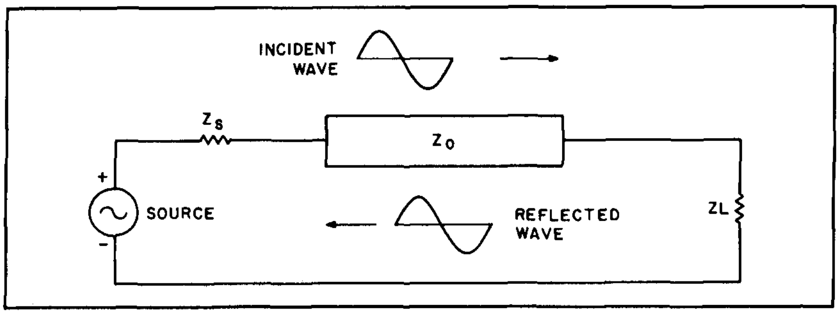


Figure 4. Uniform section of transmission line with characteristic impedance of Z_0 connected between source and load. Incident and reflected waves travel in opposing directions.

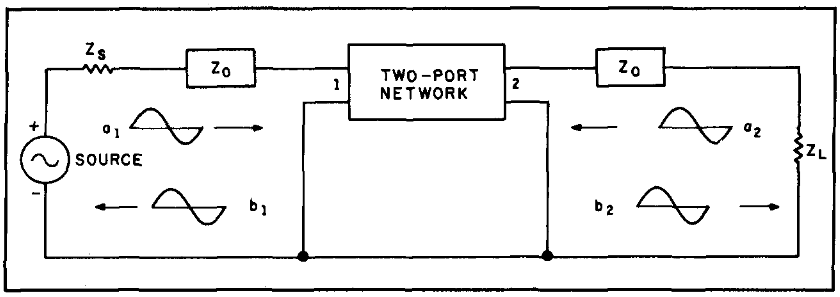


Figure 5. Two-port network inserted between two sections of transmission line. Total of four traveling waves now appear because the network appears as a terminating junction for each line.

For a lossless transmission line $R=G=0$, and **Equation 4** reduces to **Equation 5**, indicating that the impedance is independent of frequency.

$$Z_o = \sqrt{\frac{L}{C}} \tag{5}$$

Figure 4 shows a uniform section of transmission line with a characteristic impedance of Z_O, connected between a source and load. The reflected wave is a wave component of power, voltage, or current transmitted from the source to the load. These wave components along the transmission line are actually the traveling waves and, as shown in **Figure 4**, travel in directions opposite from each other.

Examine the behavior of the circuit in **Figure 4** more closely. When $Z_L = Z_O$, the incident wave is completely dissipated in the load; therefore, the reflected wave is eliminated. However, if these impedances aren't equal, a reflection exists. The reflected wave which is part of the incident wave propagates along the transmission line back towards the source, and is dissipated in Z_S if $Z_S = Z_O$. If these impedances aren't equal, part of the reflected wave will again reflect from the source impedance and re-propagate along the transmission line back towards the load. In the case of a lossless transmission line, this phenomenon would continue endlessly.

The reflection coefficient is another important expression in transmission line theory. This coefficient indicates how well the terminating impedance is matched to the transmission line impedance. The reflection coefficient shown in **Equation 6** is a complex quantity containing both magnitude and phase.

$$\Gamma = \frac{V_{reflected}}{V_{incident}} = \frac{Z_L - Z_o}{Z_L + Z_o} \tag{6}$$

For a perfect match to exist, the reflection coefficient must equal zero, indicating there are no reflections. If the reflection coefficient is 1, then a complete mismatch exists (that is, a short or open circuit), and the entire incident wave is reflected. To maintain circuit stability, the reflection coefficient must be between zero and 1.

Scattering parameters

Now that I've reviewed some of the basic concepts of transmission line theory, I'd like to begin exploring the properties of scattering parameters. Let's start by connecting a two-port network between two sections of transmission line and terminating the end of each line into an impedance of Z_s and Z_L as shown in **Figure 5**. You now have a total of four traveling waves, because the two-port network appears as a terminating junction for each transmission line. These additional traveling waves can be explained by applying transmission line principles to the circuit in **Figure 5**.

Starting at the source, a wave is generated and travels toward port 1. The wave becomes incident upon arrival at port 1, and is called a1. Part of incident wave a1 reflects from port 1, becoming the reflected wave b1, which travels back towards the source. The remaining part of incident wave a1, propagates through the network and becomes the b2 wave component, which is dissipated in the load. This is viewed as a reflected wave because it's leaving port 2. If there's a mismatch, part of the b2 will reflect from the load and travel back towards port 2. This wave component is now incident upon port 2 and is called a2.

This explains why four traveling waves exist where a1,a2 are incident upon a port and b1,b2 are reflected from a port. If you break these traveling wave components down further, you'll find they are actually voltages normalized to the square root of the transmission line's impedance as shown in **Equations 7a-d**.

$$a_1 = \frac{V_1 + I_1 Z_0}{\sqrt{2 Z_0}} = \frac{V_{i1}}{\sqrt{Z_0}} \quad \textbf{(7a)}$$

$$a_2 = \frac{V_2 + I_2 Z_0}{\sqrt{2 Z_0}} = \frac{V_{i2}}{\sqrt{Z_0}} \quad \textbf{(7b)}$$

$$b_1 = \frac{V_1 - I_1 Z_0}{\sqrt{2 Z_0}} = \frac{V_{r1}}{\sqrt{Z_0}} \quad \textbf{(7c)}$$

$$b_2 = \frac{V_2 - I_2 Z_0}{\sqrt{2 Z_0}} = \frac{V_{r2}}{\sqrt{Z_0}} \quad \textbf{(7d)}$$

The linear equations that describe S-parameters, based on the four traveling waves in **Figure 5**, are given in **Equations 8a** and **b**.

$$b_1 = S_{11}a_1 + S_{12}a_2 \quad \textbf{(8a)}$$
$$b_2 = S_{21}a_1 + S_{22}a_2 \quad \textbf{(8b)}$$

These equations look similar to the H, Y, and Z linear equations. Like them, one of the variables is set to zero when solving for a particular parameter. With S-parameters, a1 or a2 are the variables that are set to zero – depending on the parameter being measured. Unlike traditional parameters, these variables are set to zero by terminating the network with a resistive impedance equal to the measuring systems impedance Z_o. This is also easily proven by applying transmission line principles. Any incident wave upon the termination is totally dissipated because the termination impedance is equal to the system's impedance. Because these impedances are equal, any reflections associated with the termination are now completely eliminated. A reflected wave in the measuring system now represents the mismatch of the network's port.

The four S-parameters – S_{11}, S_{21}, S_{12}, and S_{22} – for a two-port network are derived from the linear equations given in **Equations 9a-d**. When measuring S_{11} or S_{21}, port 1 is driven from a source and port 2 is terminated. When measuring S_{22} or S_{12}, port 2 is driven from a source and port 1 is terminated.

Input Reflection Coefficient

$$S_{11} = \left. \frac{b_1}{a_1} \right| a_2 = 0 \ (Z_L = Z_0) \quad \textbf{(9a)}$$

Forward Transmission Gain

$$S_{21} = \left. \frac{b_2}{a_1} \right| a_2 = 0 \ (Z_L = Z_0) \quad \textbf{(9b)}$$

Reverse Transmission Gain

$$S_{12} = \left. \frac{b_1}{a_2} \right| a_1 = 0 \ (Z_s = Z_0) \quad \textbf{(9c)}$$

Output Reflection Coefficient

$$S_{22} = \left. \frac{b_2}{a_2} \right| a_1 = 0 \ (Z_s = Z_0) \quad \textbf{(9d)}$$

Equations 9a-9d indicate that S-parameters are ratios of normalized reflected-to-incident waves from the ports of a network. S-parameters fall into two categories – a reflection coefficient and a gain. S_{21} is the forward gain and S_{12} is the reverse gain. S_{11} and S_{22} are the input and output reflection coefficients due to impedance mismatch losses. S-parameters are complex quantities containing both magnitude and phase, and are a function of frequency. The later bit of information is important to know because you must realize that S-parameters are valid only at the frequency at which they were measured. For example, you couldn't design a circuit at, say, 500 MHz using S-parameter data for a device taken at 1 MHz, and expect the circuit to operate according to the design.

Because S-parameters are ratios of traveling waves, it would be helpful to go back and review **Equations 7a-d**. Here you actually have nothing more than a voltage divided by a resistance. Squaring the magnitude of these equations yields the more familiar term E^2/R, known as power. Now you can think of a1,a2 as incident power upon a port, and b1,b2 as reflected power from a port. On the basis of this information, you can arrange the S-parameter equations as ratios of power as shown in **Equations 10a-d**.

$$|S_{11}|^2 = \frac{\text{Power Reflected From Port 1}}{\text{Power Available From Port 1}} \quad \textbf{(10a)}$$

$$|S_{21}|^2 = \frac{\text{Power Delivered to Load at Port 2}}{\text{Power Available From Source at Port 1}} \quad \textbf{(10b)}$$

$$|S_{12}|^2 = \frac{\text{Power Delivered to Load at Port 1}}{\text{Power Available From Source at Port 2}} \quad \textbf{(10c)}$$

$$|S_{22}|^2 = \frac{\text{Power Reflected From Port 2}}{\text{Power Available From Port 2}} \quad \textbf{(10d)}$$

Although the concept of terms of power is easier to understand, it's more useful to define S-parameters in their decibel form. This is probably the most common form, because most network analyzers and S-parameter data from manufacturers is often given in this way. Also, most people are comfortable working with decibels. **Equations 11a-11d** give the decibel form of S-parameters.

$$|S_{11}|_{dB} = 20*log\,|S_{11}| \quad \textbf{(11a)}$$

$$|S_{12}|_{dB} = 20*log\,|S_{12}| \quad \textbf{(11b)}$$

$$|S_{21}|_{dB} = 20*log\,|S_{21}| \quad \textbf{(11c)}$$

$$|S_{22}|_{dB} = 20*log\,|S_{22}| \quad \textbf{(11d)}$$

Applying S-parameters

Now that you have an understanding of S-parameters, let's put them to practical use. I obtained the common emitter data for a 2N6604 high frequency transistor shown in **Table 1** from a manufacturer's data sheet. The transistor was biased for an Ic=10 mA and Vce=5 volts. Data was measured at a frequency of 1 GHz with a system impedance of 50 ohms. I would like to stress that the results obtained in the following calculations are valid only at 1 GHz.

At first glance, you can see that the transistor has an unmatched gain (terminated directly into 50 ohms) of 11.53 dB and the isolation from output to input is −23.22 dB.

As stated earlier, the stability of a network can be determined from S-parameters. When a network is unstable, the magnitude of the reflection coefficient is greater than 1. This is equivalent to terminating the network with a negative resistance which is needed to start an oscillation. On an E-I curve, negative resistance has a negative slope; that is, for an increase in voltage, there is a decrease in current, and visa versa. A negative resistance seems to generate power opposite to that of a positive resistance, which dissipates power.

A network is unconditionally stable if it can be terminated with any impedance having a positive real part (that is, R±jX, R = the positive resistor) and conditionally stable for some impedances with a positive real part. The stability is determined by the Rollet Stability Factor (K) given in **Equation 12**.

$$K = \frac{1 + |S_{11}*S_{22} - S_{12}*S_{21}|^2 - |S_{11}|^2 - |S_{22}|^2}{2 * |S_{21}| * |S_{12}|} \quad \textbf{(12)}$$

Unconditional stability is achieved when K>1. You'll have a potentially unstable state when K<1. A potentially unstable device isn't necessarily unusable, but you must take a different design approach to gain stability. For example, stability circles can be plotted

	S_{11}	S_{21}	S_{12}	S_{22}
Magnitude	0.61	3.77	0.069	0.29
Angle degree	−178	76	40	−62
Decibels dB	−9.88	11.53	−23.22	−10.75

Table 1. Common Emitter Data for 2N6604 High-Frequency Transistor.

on a Smith chart to determine what impedances, if any, will ensure stability. The calculations for the 2N6604 transistor, when plugged into **Equation 12**, indicate that it is unconditionally stable.

$$K = \frac{1 + (0.61*0.29 - 0.069*3.77)^2 - (0.61)^2 - (0.29)^2}{2*3.77*0.069}$$

$$= 1.058$$

Now that you've determined that the transistor has unconditional stability, you can calculate the Maximum Available Gain (MAG). The MAG indicates what the maximum gain of the transistor will be under stable conditions with the input and output conjugately matched. First, find the B_1 factor using **Equation 13**. This determines which sign (±) to use in **Equation 14** — the equation for calculating MAG. If B_1 is positive, use the negative sign in **Equation 14**.

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |S_{11}*S_{22} - S_{12}*S_{21}|^2 \quad \textbf{(13)}$$

$$MAG = 10*log\,\frac{|S_{21}|}{|S_{12}|} * |K \pm \sqrt{K^2 - 1}| \quad \textbf{(14)}$$

$$B1 = 1 + (0.61)^2 - (0.29)^2 - (0.61*0.29 - 0.069*3.77)^2$$

$$= + \textit{sign}$$

$$MAG = 10*log\,\frac{(3.77)}{(0.069)}\,(1.058 - \sqrt{(1.058)^2 - 1}$$

$$= 15.9\ dB$$

The input and output impedances are determined using **Equation 15**.

$$Zin = \frac{Z_0\,(1 + S_{11})}{1 - S_{11}} \quad \textbf{(15)}$$

$$Zout = \frac{Z_o\,(1 + S_{22})}{1 - S_{22}}$$

By way of example, you can calculate the output impedance of the transistor. Enter the S-parameters as complex quantities (magnitude and phase) as indicated in **Equation 15.**

$$Zout = \frac{50*(1 + 0.29 \angle{-62°})}{1 - 0.29 \angle{-62°}} =$$

$$\frac{50*(1 + 0.136 - j0.256)}{1 - 0.136 + j0.256} =$$

$$\frac{56.8 - j12.8}{0.864 + j0.256} = 56.39 - j31.52$$

Zout is 56.39-ohm resistor in series with a capacitor which has a reactance of 31.52 ohms. Using this information, you can design a matching network to match the output of the transistor to the impedance of a load Z_0. Also, because S_{11} and S_{22} are reflection coefficients, you could have plotted them on a Smith chart and found the impedance directly from the chart.

You can calculate the input VSWR from **Equation 16**. To calculate the output VSWR, replace S_{11} with S_{22}. As an example, the input VSWR is determined below.

$$VSWR = \frac{1 + |S_{11}|}{1 - |S_{11}|} \qquad \textbf{(16)}$$

$$VSWR = \frac{1 + 0.61}{1 - 0.61} = 4.12{:}1$$

As a general rule of thumb, impedances are considered well matched if the reflection coefficient is greater than or equal to 18 dB (that is, VSWR $\leq$ =1.28:1).

The S-parameters for a 0.1-dB Chebyshev lowpass filter are plotted in **Figure 6,** providing you with another example. Because this filter is a bilateral device ($S_{11} = S_{22}$ and $S_{21} = S_{12}$), only S_{11} and S_{21} are shown. This plot gives a picture of the filter's actual transfer characteristic and impedance over frequency. By looking at this plot, you can see the complete behavior of the filter in terms of the type of response, rolloff rate, cutoff frequency, passband and stopband attenuation, and the impedance.

The preceding examples express the power behind S-parameters. Although I've only given two, there are many other applications for S-parameters too numerous to explain here. To gain a more in-depth understanding of S-parameters and their use in circuit design, you might like to read some of the references listed in the bibliography at the end of this article.

Conclusion

S-parameters provide a gateway for characterizing a network at higher frequencies in places where other parameters fail. They are accurate, dependable, easy to understand, and practical to apply. S-parameters provide a greater and more meaningful insight into the performance of a network than do their corresponding parameters. By looking at the examples given, you'll note that a wealth of useful information can be obtained directly from S-parameters, without the need of higher level mathematics. As a matter of fact, you can pretty much characterize a network simply by observing its S-parameters. S-parameters play an important role in the RF field, and with all of their advantages, are worthwhile parameters to understand. ■

Bibliography

1. Hewlett Packard, Application Note 95-1, "S-Parameter Techniques for Faster, More Accurate Network Design."
2. Hewlett Parkard, Application Note 154, "S-Parameter Design."
3. Stephen F. Adam, Hewlett Parkard, *Microwave Theory and Applications*, Prentice Hall, 1969.
4. Chris Bowick, *RF Circuit Design*, Howard W. Sams & Co., Inc.
5. Jackie L. Huges, "Broadband Microwave Transistor Amplifier Design Using S-Parameters," *RF Design*, November 1987.
6. Thomas H. Stanford, "S-Parameters Aid Design of RF Amplifiers," *Microwaves & RF*, June 1988.
7. Stanley J. Domanski, "Designing a Microwave Amplifier," *Ham Radio*, September, 1987.

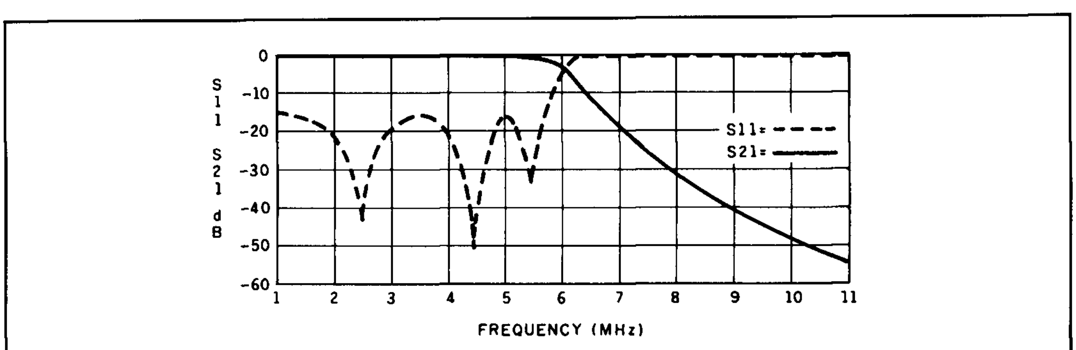


Figure 6. S-parameters plotted for a 0.1-dB Chebyshev low-pass filter. Because the filter is bilateral, only S_{11} and S_{21} parameters are shown.

ADVERTISER'S INDEX

We'd like to see your company listed here too. Contact Arnie Sposato, N2IQO, at (516) 681-2922 or FAX at (516) 681-2926 to work out an advertising program to suit your needs.


Yaesu Announces
One Year
Limited Warranty
on all Amateur Radio Products
INDUSTRY FIRST!
FT-411E
(ENHANCED)
INDUSTRY FIRST! 10-memory auto dialer
INDUSTRY FIRST! One-touch instant recall of favorite channel
INDUSTRY FIRST! APO (Automatic Power Off)
INDUSTRY FIRST! Built-in PL encode/decode
INDUSTRY FIRST! Backlit keypad & display
INDUSTRY FIRST! 10 battery saving sampling rates
INDUSTRY FIRST! PTT/keypad lock
INDUSTRY FIRST! 2 VFO's
INDUSTRY FIRST! Programmable channel steps (5, 10, 12.5, 20, 25)
INDUSTRY FIRST! Built-in VOX at no charge
INDUSTRY FIRST! User friendly for those with impaired vision. Keys have their own distinctive tone.
INDUSTRY FIRST! Built-in CTCSS paging
YAESU
2m FM TRANSCEIVER
FT-411E
(FT-411E shown actual size.)
FT-811
Small Size and Most Features
FT-470
Most Popular Dual Band
New
FT-911
Reliable Novice Radio
For more information on these and other Yaesu products, contact your nearest Yaesu dealer now.
YAESU
Performance without compromise.SM
© 1990 Yaesu USA, 17210 Edwards Road Cerritos, CA 90701

KENWOOD

The Ultimate Signal.

TS-950SD

"DX-clusive" HF Transceiver

The new TS-950SD is the first Amateur Radio transceiver to utilize Digital Signal Processing (DSP), a high voltage final amplifier, dual fluorescent tube digital display and digital meter with a peak-hold function.

• **Dual Frequency Receive Function.** The TS-950SD can receive two frequencies simultaneously.

• **New! Digital AF filter.** Synchronized with SSB IF slope tuning, the digital AF filter provides sharp characteristics for optimum filter response.

• **New high voltage final amplifier.** 50 V power transistors in the 150-watt final section, resulting in minimum distortion and higher efficiency. Full-power key-down time exceeds one hour.

• **New! Built-in microprocessor controlled automatic antenna tuner.**

• **Outstanding general coverage receiver performance and sensitivity.** Kenwood's Dyna-Mix™ high sensitivity direct mixing system provides incredible performance from 100 kHz to 30 MHz. The Intermodulation dynamic range is 105 dB.

• **Famous Kenwood interference reduction circuits.** SSB Slope Tuning, CW VBT (Variable Bandwidth Tuning), CW AF tune, IF notch filter, dual-mode noise blanker with level control, 4-step RF attenuator (10, 20, or 30 dB), switchable AGC circuit, and all-mode squelch.

Complete service manuals are available for all Kenwood transceivers and most accessories.
Specifications, features and prices subject to change without notice or obligation.

DSP

Digital Signal Processing

• **Digital Signal Processor.** DSP is a state-of-the-art technique that maximizes your transmitted RF energy.

• **High performance IF filters built-in**† Select various filter combinations from the front panel. For CW, 250 and 500 Hz, 2.4 kHz for SSB, and 6 kHz for AM. Filter selections can be stored in memory!

• ***Multi-Drive Band Pass Filter* (BPF) circuitry.** Fifteen band pass filters are available in the front end to enhance performance.

• **Built-in TCXO for the highest stability.**†

• **Built-in electronic keyer circuit.**

• **100 memory channels.** Store independent transmit and receive frequencies, mode, filter data, auto-tuner data and CTCSS frequency.

• **Digital bar meter.**

Additional Features: • Built-in interface for computer control • Programmable tone encoder • Built-in heavy duty AC power supply and speaker • Adjustable VFO tuning torque • Multiple scanning functions • MC-43S hand microphone supplied

Optional Accessories

- DSP-10 Digital Signal Processor*
- SO-2 TCXO* • VS-2 Voice synthesizer
- YK-88C-1 500 Hz CW filter for 8.83 MHz IF*
- YG-455C-1 500 Hz CW filter for 455 kHz IF*
- YK-88CN-1 270 Hz CW filter for 8.83 MHz IF
- YG-455CN-1 250 Hz CW filter for 455 kHz IF*
- YK-88SN-1 1.8 kHz SSB filter for 8.83 MHz IF
- YG-455S-1 2.4 kHz SSB filter for 455 kHz IF*
- SP-950 External speaker w/AF filter
- SM-230 Station monitor w/pan display
- SW-2100 SWR/power meter
- TL-922A Linear amplifier (not for QSK)

* Built-in for the TS-950SD
† Optional for the TS-950S

KENWOOD U.S.A. CORPORATION
COMMUNICATIONS & TEST EQUIPMENT GROUP
P.O. BOX 22745, 2201 E. Dominguez Street
Long Beach, CA 90801-5745

KENWOOD ELECTRONICS CANADA INC.
P.O. BOX 1075, 959 Gana Court
Mississauga, Ontario, Canada L4T 4C2

KENWOOD
...pacesetter in Amateur Radio