Index to Volume LXXVIII—1994

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QST *Features Editor*

How to Use This Index

Items are listed according to the regular column under which they appeared, and/or under any category (or categories) that generally describes them. The name of the author(s) appears in parentheses, followed by a colon, then the issue and page number. Example: "Epics from the Epicenter" appears under Features—General Interest and under Public Service. Some items have additional subject-description information or the name of a department in which it appeared, in brackets after the title.

EXAM INFO—Column (Jahnke)

FEATURE ARTICLES

General Interest

Why I Like Field Day (Miccolis): Jun, 55
Wire Antenna, The (Schmidt): Nov, 55
Wired for Wireless: Ted Rappaport, N9NB, and His Vision (Cain): Jul, 48
World's Greatest DXer, The [W4DR] (Cain): Nov, 43
X-Banding From an XYL's Perspective (Lee): Mar, 49
Yukon DXing with Flair (Reisenauer): Apr, 24

Technical

160-Meter Sloper System at K3LR, The (Christman, Duffy & Breakall): Aug, 36
Amateur Use of Telescoping Masts (Haviland): May, 41
Anatomy of a Club Homebrew Project: Nov, 39
ATV Station for 915 MHz, An—*Part 1:* A Miniature 1.5-W ATV Transmitter (Graf & Sheets): Nov, 23
ATV Station for 915 MHz, An—*Part 2:* A Tunable ATV Downconverter (Graf & Sheets): Dec, 28
Automatic Temperature-Controlled Fan, An (Kolts): Jun, 41
Beginner's Boomers: Two Phased Vertical Arrays for 30 Meters (Borich & Logan): Jun, 37
Build Your Own *Lowfer* Transceiver (Curry): Apr, 26
Calibrated Noise Source for Amateur Radio, A (Sabin): May, 37
Computer-Controlled Electronic Test Equipment—*Part 2* (Portugal): Jan, 35 (see Dec '93, 42)
CW "Stamp" Identifier, A (Bunn): Oct, 41
Direct Digital Synthesis—An Intuitive Introduction (Cahn): Aug, 30
Easy-to-Build 25-Watt MF/HF Amplifier, An (Breed): Feb, 31
Elkhart County Tone Alert, The (Drudge): Apr, 43
Exploring Intermodulation Distortion in RF Switching and Tuning Diodes (Thompson): Dec, 25
Function Generator with a Frequency-Counter Digital Readout, A (Spencer): Apr, 35
Home-Brew Loop Tuning Capacitor, A (Jones): Nov, 30
Inexpensive Interference Filters (Bloom): Jun, 32
Inexpensive SSTV System, An (Vester): Jan, 27 (see Feedback: Feb, 77)
Inexpensive SSTV System Continues to Grow, An (Vester): Dec, 22
Introducing *ARRL Radio Designer*: New Software for RF Circuit Simulation and Analysis (Newkirk): Oct, 21
Just Enough Radio—The SP-750 Spider Junior (Agsten): Nov, 33
KD7IK's "Quad Lite" (Bock): Oct, 39
Key Components of Modern Receiver Design—*Part 1* (Rohde): May, 29
Key Components of Modern Receiver Design—*Part 2* (Rohde): Jun, 27
Key Components of Modern Receiver Design—*Part 3* (Rohde): Jul, 42
Key Components of Modern Receiver Design: A Second Look (Rohde): Dec, 38
Lead-Acid Battery Charger, A (Spencer): Mar, 25
Mother Nature's Radio (Schneider): Jan, 49
Nearly Perfect Amplifier, The (Measures): Jan, 31
Nickel-Metal-Hydride Batteries in Amateur Radio Applications (Kuusisto): Sep, 38
Null Steerer Revisited, The (Michaels): Jul, 29
On Center-Fed Multiband Dipoles (Belrose & Bouliane): Mar, 35
Overvoltage Protection for AC Generators (Paquette): Jun, 43
Pfeiffer Quad Antenna System, The (Pfeiffer): Mar, 28
Pocket-Sized, *Talking* Morse Code Practice Computer, A (Staton): Jul, 26
QSOcorder, The (Reyer): Feb, 45
QST Product Reviews: A Look Behind the Scenes (Gruber): Oct, 35
Quick Powerhouse, The (Miller): Dec, 33
Radio Gear of Yesteryear (Shrader): Mar, 41
Reevaluation of the Caron RF Impedance Bridge, A (Camillo): Sep, 28
Revisiting the RF Ammeter (Stanley): Feb, 35
Ringmaster Ring Decoder, The (Rumbolt): Apr, 40
Simple Dummy Antenna for ATV Transmitter Testing, A (Sheets): Nov, 27
Simple, Effective Elevated Ground-Plane Antennas (Russell): Jun, 45
Simple Equipment for HF Fox Hunting (Villard, Hagn & Lomasney): Aug, 33
Simple, General-Purpose AF Amplifier (Spencer): Dec, 43
Single-Board Superhet QRP Transceiver for 40 or 30 Meters, A (Benson): Nov, 37
Smart Charger for Nickel-Cadmium Batteries, A (Avritch): Sep, 40
So What's New in the *ARRL Antenna Book*? (Straw): Sep, 24
Stacking Tribanders: A Super Station—Sorta (Straw & Hopengarten): Feb, 38
SWR Detector Audio Adapter, An (Spencer): Jul, 24
Thermoelectric Power for QRP Transmitters (Sayre): Sep, 33
Those New *QST* Propagation Charts (Hall): Oct, 27
Trigonal HF Beam, A (Bird): Apr, 32
Two New Multiband Trap Dipoles (Buxton): Aug, 26
Uncle Albert's Unique Keyer (Ulbing): Jan, 43
Under the Hood III: *Capacitors* (Bergeron): Jan, 45
Under the Hood IV: *Inductors* (Bergeron): Mar, 37
Under the Hood V: *Solid State Devices* (Bergeron): May, 46
Under the Hood: *Lamps, Indicators and Displays* (Bergeron): Sep, 34
Using a VU Meter for Phone-Patch Adjustment (Lorona): Mar, 32
You Can Build: A Compact Loop for 30 through 12 Meters (Capon): May, 33

FEEDBACK

1993 Simulated Emergency Test (SET) Results (Ewald): Sep, 95 (see Jul, 54)
Amateur Use of Telescoping Masts (Haviland): Jul, 83 (see May, 41)
Calibrated Noise Source for Amateur Radio, A (Sabin): Jul, 83 (see May, 29)
Computer-Controlled Electronic Test Equipment—*Part 1* (Portugal): Feb, 77 (see Dec '93, 45)
Function Generator with a Frequency-Counter Digital Readout, A (Spencer): Dec, 85 (see Apr, 38)
Inexpensive SSTV System, An (Vester): Feb, 77 (see Jan, 27)
New Products: Oct, 45 (see Aug, 70, and Sep, 39)
Null Steerer Revisited, The (Michaels): Sep, 95 (see Jul, 30)
Plug Into PacTOR: Jun, 78 (see Mar, 67)
Product Review: JPS SSTV-1 DSP Filter: Dec, 85 (see Nov, 81)
Radio Gear of Yesteryear (Shrader): Jun, 78 (see Mar, 41)
Reevaluation of the Caron RF Impedance Bridge, A (Camillo): Nov, 88 (see Sep, 28)
Smart Charger for Nickel-Cadmium Batteries, A (Johnson): Nov, 88 (see Sep, 41)
So What's New in the ARRL Antenna Book (Straw): Oct, 80 (see Sep, 24)
SWR Detector Audio Adapter, An (Lones): Aug, 69; Nov, 88 (see Jul, 24)
Wise Owl Worldwide Publications: Jun, 78 (see Strays: Apr, 107)

FM—Column (Battles)

ATV: Not as Easy as it Looks, But Still Fun (Lee): Oct, 95
Crossband and Linked Repeaters: Their Potential in Emergencies—*Part 1* (Magness): Jun, 98
Crossband and Linked Repeaters: Their Potential in Emergencies—*Part 2* (Magness): Jul, 107
Love of Amateur Radio and His Fellow Hams (Gartenberg): Aug, 88
One Cool Radio (Staponski): Aug, 88
PC Voice on FM (DCI): Apr, 109
Santa is a Ham: A Christmas to Remember (Foble): Dec, 103
VHF Contesting: Jan, 113
Wilderness VHF FM Protocol, A (Alsup): Feb, 99
Wilderness VHF FM Protocol, A: Your Comments—*Part 1:* Apr, 109
Wilderness VHF FM Protocol, A: Your Comments—*Part 2:* May, 103
With a Little Bit of Planning: A High-Speed QSO (Lee): Nov, 64
Your Dream Rig? Here's What You Said...: Sep, 113
Your Dream Rig?: Mar, 109

HAPPENINGS—News (Cain)

222 MHz: Novices Get More; Weak-Signal Band Okayed: Jan, 83
1993 *QST* Cover Plaque Award Winners Named: Feb, 83
1994 Dayton HamVention Names Award Winners: May, 90
Action Urged on Call Sign Selection: Jul, 84
Amateur's "Black Box" is Government Gold Mine: Mar, 93
Amateur Radio Bills Hang in Balance as 103rd Congress Nears its End: Nov, 93
Amateur Radio Resolution Gains Majority in House: Jun, 83
Amateur Radio Show Goes National: Feb, 81
Appeals Court Sides with Minnesota Amateur; Antenna Victory Reinforces Federal Declaration: Mar, 91
AMSAT-NA Annual Meeting Draws 240 to Orlando: Dec, 88
ARRL and FCC Agree to New Cooperative Agreement: May, 91
ARRL Congress Watchers Remain Optimistic: Oct, 85
ARRL, Lambda ARC Agree, Issue Joint Statement: Dec, 88
ARRL Members Pick New Board Representatives: Jan, 83
ARRL Ombudsman Debuts: Jul, 87

Build Your Own Shortwave Antennas [Yoder] (Kleinschmidt): Oct, 74
Code Book, The: Amateur Radio CW Operating [Halprin] (Battles): Feb, 53
Code Book, The: Morse Code Instruction Manual [Butt] (Battles): Jun, 74
Communications Receivers, The Vacuum Tube Era: 1932-1981 [Moore] (Kleinschmidt): Oct, 74
CQ *1994 Amateur Radio Almanac* [Grant] (Battles): May, 54
DOS for Dummies [Gookin] (Battles): Jul, 41
Easy Target [Wall] (Dunn): Nov, 36
Emergency Radio! [Schrein] (Battles): Sep, 42
Encyclopedia of Electronic Circuits, Vol 4 [Graf & Sheets] (Hale): Mar, 31
Ham Radio Contesting [Halprin] (Kearman): May, 115
Hidden Ham Antennas [Hughes] (Kearman): Sep, 58
How to Get Started in QRP [Ingram] (Kearman): Jun, 74
Internet for Dummies, The [Levine & Baroudi] (Battles): Jul, 41
Just When Tom Had Jean Convinced That His Friends Were Normal, He Took Her to a Hamfest! [Irwin] (Battles): Jun, 44
Low Power Communications Volume 2, Advanced QRP Operating [Arland] (Kearman): Jun, 42
MARS: Calling Back to "The World" from Vietnam [Scipione] (Kearman): Nov, 102
Mastering Radio Frequency Circuits Through Projects and Experiments [Carr] (Kearman): Sep, 37
Mobile 2-Way Radio Communications [West] (Battles): Jan, 59
More About Cubical Quads [McCarthy] (Cain): Oct, 34
National Radio Club *AM Radio Log* [Chatterton] (Kleinschmidt): Sep, 45
NOSIntro [Wade] (Battles): Mar, 107
OS/2 for Dummies [Rathbone] (Battles): Jul, 41
Power Up! [Strom] (Battles): Jul, 47
Receiving Antenna Handbook, Joe Carr's [Carr] (Kearman): Oct, 82
Riding the Airwaves with Alpha and Zulu [Abbott] (Battles): Feb, 37
Sam and Erin Go to a Hamfest [Wall] (Battles): Jul, 108
Satellite Communications and DBS Systems [Wood] (Ford): Apr, 31
Space Satellite Handbook [Curtis] (Ford): May, 36
Tune to Satellite Radio on Your Satellite System [Harrington] (Kleinschmidt): Oct, 80
UNIX for Dummies [Levine & Young] (Battles): Jul, 41
W1FB's Help for New Hams [DeMaw] (Ford): Sep, 64
What is Your TNC Doing? [Medcalf] (Ford): Apr, 34
Windows for Dummies [Rathbone] (Battles): Jul, 41

NEW HAM COMPANION

10 Meters is *Not* Dead! (Ford): Apr, 68
2-Meter FM DXing (Ford): Sep, 60
Amateur Radio 101 (Hughes): Jul, 71
"At the Tone..." (Ford): Jun, 62
Backpacking, Troop 404 Style (Rowlett): Oct, 60
Be a Bone Yard PhD (Keith): Jun, 64
Build a 12-V Junction Box (Capon): Aug, 54
Build a One-Watt Transmitter in a Kodak Film Box (Capon): Oct, 64
Build an HF Walking Stick Antenna (Capon): Dec, 72
Building and Adjusting Trap Dipole Antennas (Edwards): Nov, 72
Building Your Own Station Accessories (Gold): Feb, 61
But How Do I Use It? (Danzer): May, 66
Cheap Way to Hunt Transmitters, A (Rickerd): Jan, 65
Climatological Analysis of the Dayton HamVention, A (Friedman): Apr, 69
Conquering the Code (Bellamy): May, 64
Design Your Own Photographic QSL Card (Weisman): Sep, 63
Do I Need a Linear Amplifier? (Aurick): Apr, 73
Do You Need an Antenna Tuner? (Ford): Jan, 71
DXing with 2-Meter Packet Mail (Smith): Feb, 66
Easy Dual-Band VHF/UHF Antenna, An (Reynante): Sep, 61
El Dipolo Criollo (Meara): Aug, 58
Fishing Tackle HF Station to Go!, A (Capon): Nov, 67
FM Contesting on Taylor Mountain (Parmley): Dec, 71
Gain Antenna for 28 MHz, A (Beezley): Jul, 70
Get Out Your Calculators or Computers (Straw): Apr, 71
Getting More Replies to Your CW Calls (Brogdon): Jun, 61
Getting Started on the Magic Band (Neubeck): Mar, 61
Go Digital! (Kleinschmidt): Sep, 67
Hams 'R Us Kids Net (Bernotas): Nov, 69
Importance of Zero-Beating, The (Shrader): Oct, 58
Indestructible Dipole for 10 Meters, An (Bowles): Apr, 67
Interference in Reverse (Freedom): May, 65
Make Your Mobile More Portable (Mendelsohn): May, 62
Modest Multiband Antenna, A (Brogdon): Jul, 68
NAVTEX and Your Multimode TNC (Ford): Aug, 52
NiCd Never Forgets, A. Or Does It? (Gruber): Nov, 71
On the Wings of a DOVE (Ford): Sep, 65
Over-the-Dash H-T Mount, An (Leyson): Mar, 70
Over-the-Keyboard Desk Top (Leyson): Jul, 75
PACSATs from an Apartment! (Schliemann): Mar, 65
Packet Without Computers (Wolf): Apr, 64
PC Shopper's Guide, A (Ford & Kleinschmidt): May, 69
Plug Into PacTOR (Gold): Mar, 67
QSL Cards? Before You Write That Check... (Bowles): Dec, 64
QSLing Through a Manager (Miller): Aug, 56
Repeater Eater, The (Murphy): Nov, 74
RS-12 Worked All States (Peschka): Oct, 67
Six-Meter Spectrum, The: Mar, 63
Such a Deal! (Ford): Jul, 73
SWR Obsession, The (Ford): Apr, 70
Test Day (Bowles): Jan, 62
Trusty Slingshot, The (Ford): Jul, 70
Understanding Signal Strength (Wilson): Jul, 67
"Universal" VHF/UHF Antenna, A (Miller): Dec, 75
View From Above, The (Ford): Dec, 68
Where's My Mail? (Patterson): Jan, 67
Worked All Palm Beach (Penn): Mar, 60
Working Satellite RS-12—The Ultimate Satellite Primer (Capon): Feb, 58
You Never Forget the First Time (MacDonald): Jun, 58

Doctor is IN, The—Column:

Jan, 64; Feb, 65; Mar, 64; Apr, 66; May, 72; Jun, 60; Jul, 66; Aug, 61; Sep, 70; Oct, 59; Nov, 66; Dec, 67

Radio Tips

8-Band Backpacker Special (Andera): Jun, 68
10-Meter Band Plan, The: Apr, 68
Abbreviated Packet-Speak Glossary, An (Ford): Feb, 68
Activity Nights (Owen): Mar, 63
Buying on the Packet Network (Ford): Jun, 67
International Third-Party Traffic—Proceed With Caution: Dec, 70
Join the QRP Craze (Kleinschmidt): May, 68
Log it or Lose it (Kleinschmidt): Jan, 72
Packet Snooping (Kleinschmidt): Jan, 69
Selective Calling (Ford): Nov, 73
Those Versatile Hand-Helds (Ford): Jan, 66
Traffic Handling is for Everyone (Kleinschmidt): May, 63
UHF and Microwave Equipment (Healy): Aug, 60
Wave Angles (Ford): Aug, 57
What's DSP? (Ford): May, 71
Working the DX Pileups (Ford): Oct, 66

NEW PRODUCTS

300/1200/9600 bit/s TNC: Nov, 48
1994 *CQ* Ham Calendars: Jan, 128
Add-Ons for Yaesu FT-1000: Jul, 28
Affordable FM Monitor Receiver Kits: Dec, 52
AMTOR/PacTOR Software: May, 104
Amateur Radio Mail Order Catalog [Thompson] (Ford): Jul, 61
Amiga Code-Practice Software: Jun, 49
ATV Scan Converters, TNC: Sep, 31
Bases for Hand-Held Rigs: Jul, 45
Boom Mike/Headset: Apr, 42
Booster Speaker: Jan, 81
Code Keys: Sep, 39
Compact Crossband Repeater Controller: Mar, 36
Complete *NCJ* Reprints: Apr, 59
Computerized Rotator Controls: Sep, 39
CT Keyboard Labels: Oct, 45
CT Keyboard Templates: Sep, 39
CT Version 9: Jul, 25
Digital Antenna Meter: Aug, 47
Easy Way to Build Classic Transmitters, The: Feb, 34
Free Antenna Software [The Antenna Specialists]: Apr, 46
Getting Started in Contesting [video] (Stankiewicz): Oct, 77
Getting Started on VHF [video] (Ford): Oct, 30
Globe on Your PC: Jun, 40
Grove SDU 100 Spectrum Display (Kleinschmidt): Jun, 36
Hamcall CD-ROM: Jul, 62
Help for Shortwave Listeners: Mar, 111
HF and VHF/UHF Antennas: Feb, 91
Home-Brewer's Newsletter: Feb, 7
ISS Software for PacComm's PacTOR Controller [PTC]: Nov, 56
Jade Products "Jade Pole" Portable J-Pole VHF Antenna: Nov, 90
Jade Products Keyer Kit: Nov, 63
Jones Straight Key: Jan, 51

R. L. Drake SW8 General-Coverage Receiver (Kearman): Oct, 68
Radio Shack DSP Communication Noise Reduction System (Kearman): Jul, 78
Radioware SSTV Explorer (Pagel): Apr, 80
S & S Engineering ARK 40 CW QRP Transceiver Kit (Gold): May, 83
Solder-It Soldering Kit (Gruber): Apr, 78
Timewave Technology DSP-9+ and DSP-59+ Digital Signal Processors (Healy): Oct, 71
Trimble Scout GPS Hand-Held Global Positioning System Receiver (Wilson): Mar, 77
Watkins-Johnson HF-1000 General-Coverage Receiver (Newkirk): Dec, 77
Willco Electronics ICM-1024 Memory Replacement Board (Kearman): Jul, 79
Yaesu FRG-100 General-Coverage Receiver, The (Newkirk): Jan, 73
Yaesu FT-840 MF/HF Transceiver (Ford): May, 80

Solicitation for Product Review Equipment Bids:

Jan, 77; Feb, 73; May, 85; Aug, 65; Oct, 73

PUBLIC SERVICE

1993 Simulated Emergency Test (SET) Results (Ewald): Sep, 95 (see Jul, 54)
California's Burning (Palm): Feb, 23
Epics from the Epicenter (Palm): Aug, 21
Test of Performance, The [SET] (Ewald): Oct, 49

Public Service—Column (Palm)

ARES in the Big Apple (Schwartz): Dec, 97
Behind the Diamond: Meet New Western Massachusetts Section Manager Dan Senie, N1JEB (Palm): Feb, 94
Colorado Amateurs Win Award for Support During Papal Visit (Williams): Oct, 93
Disaster in Burlington (Feist): Aug, 81
Drill Turns into Real Thing, Tests New State Plan in Northern Florida (Thurston): Mar, 100
From the Ashes of the California Fires (Reinhardt): Mar, 100
Ham Radio Rides the Great Circus Train (Romelfanger): May, 94
How I Got Hooked on Traffic Handling (Brogdon): Sep, 104
K1CE's NTS Notes: Dec, 97
Ku Klux Klan, the Black Panthers and Amateur Television, The (Mallette): Feb, 93
NTS Notes: New Hams: Try Traffic Handling! (Brogdon): Mar, 101
NTS Wide-Area VHF Net Attracts New Hams (Booth): Apr, 95
Oakland Fire Department Recognizes ARES/RACES: Jun, 88
Public Service Awards Program: Jan, 109
Rewards of Being a Ham, The (Pachaly): Sep, 103
SATERN Rings Salvation Army (Shaver): Jun, 87
Simulated Emergency Test Results (Ewald): Jul, 54
Sky King Saves the Day: Apr, 96
SKYWARN Award Presented: May, 95
Special Olympics: The Supreme Public Service Event (Finch): Sep, 103
Supporting Public Service Events—*Part 1* (Fishman): Oct, 92
Supporting Public Service Events—*Part 2* (Fishman): Nov, 96
Trial By Fire: Washington State Firestorms (Price & Johnson): Dec, 96
Truly Worst "Worst-Case" Scenario, The (McCallum): Jul, 89
Word Up, NTS? (Palm): Sep, 103
World University Games (Weir): Apr, 95

STRAYS

Attention WW II ETO Vets: Sep, 117
Bicycle Mobile Event: May, 135
Coast Guard Net: Apr, 107
Coincidence: Dec, 111
Columbia Alumni Wanted: Aug, 50
County Hunters Convention: Jun, 105
CQ Girl Scout Hams: Jul, 112
Crank 'Em Down: Jan, 41
Diabetes Bike-A-Thon (Adamovich): May, 112
Digital Audio Broadcasting: Oct, 128
Drive-Through QSO (Kremer): May, 115
Emerald Isle Event (Walsh): Sep, 75
Extra Class Couples: May, 115
Fast Code in Florida: Jun, 40
Florida Ham Hotline (Holcomb): Dec, 85
Flying, With and Without Wings (Booen): Mar, 24
Glad to Meet a Ham (Leyson): May, 115
Ham BBS: Dec, 111
Ham Radio in the Blood: Jan, 56
His Master's VOX (Vilms): Sep, 112
Hospital Alarm QRM? Jun, 80
Iceland Awards: Jul, 114
League Anniversary Pins: Sep, 121
Magazine for Classic Radio Fans [*Electric Radio*]: Jun, 108
Northwestern Trail Bicycle-Mobile Adventure: Jun, 51
Novel Features Ham Hero (Craswell): Jun, 92
Nursing Home Resident Goes to the Top (Friesner): Mar, 116
Perseverance Pays [Project PENEX] (Rose): Mar, 99
Portuguese Handbook: May, 109
QCWA Banquet at Dayton (Dingle): May, 135
QRO Maritime Mobile (Terry): Oct, 34
QRP Wallpaper (Wendell): Mar, 83
QST de Gallery: Apr, 34
Radio Magazines: Apr, 107
SAREX QSLs and Upcoming Missions: Feb, 101
Secret Signals (McCallum): Sep, 108
Shuttle Successes: Jul, 36
Some Things Never Change: Jun, 54
Theramin, Leon, SK (Egelberg): Oct, 91
Transatlantic Tests: Jan, 96
US Islands Awards: Dec, 104
VHF/UHF Conference: May, 112
Yo! Getcha Red-Hot Radio-Astronomical Data!: Sep, 75
Youth Net on Internet: Jun, 51; Correction: Jul, 114

TECHNICAL CORRESPONDENCE—Column (Pagel)

Automatic Antenna Tuners for Wire Antennas (Belrose): Apr, 84
Beginner's Boomers Revisited (Lewallen): Oct, 78
Computer-Aided Design of Loaded Short-Doublet Antennas (Ellers): Oct, 79
Correlating Solar Flux and Sunspots (Hall): Feb, 76
Eclipse-Enhanced HF Propagation (Lewis): Nov, 87
Elevated Radials Work! (Bowen): Dec, 84
HF Mobiling: Mag Mounts and Grounding (Brogdon): May, 88
Large Antenna Coils (Eddy): Jan, 80
Maximum Bandwidth Monopole Antennas (Formato): Mar, 80
MININEC Bugs: K6STI Plays Exterminator (Beezley): Feb, 76
Modem RFI (Applegate): Aug, 69
More on Current Distribution and Pattern Nulls (Michaels): Oct, 78
More on Electrical Protection Devices (Hart): Jan, 81
More on Radio Gear of Yesteryear (Byron): Oct, 79
More on the SI8901/SD8901 (Carver): Feb, 77
More Tower Safety Tips (Burkhart): Dec, 85
North Shadow (McNally): Jul, 83
Practical, Compact Multiturn Transmitting Loop Antenna for 80 Meters, A (Jones): Nov, 87
Relay Chatter (Huston): Jun, 78
Revisiting the "Nearly Perfect Amplifier" (Telewski, Brandon, Clemow, Fakan, Katz, Rauch): Sep, 71
RFI From Household Electronics (Crabil): Aug, 68
Sniff! Sniff! I Smell Line Noise! (Thorington): Apr, 85
Sound of a Spark transmitter, The (Belrose): Aug, 68
Subject: MOVs (Sandoz): Jul, 82
Telling Time (Bessette): Oct, 79
Terminated Folded Dipole (Belrose): May, 88
Toroidal-Core Color Codes (Czuhajewski): Mar, 81
Vester SSTV/FAX480/Fax System Upgrades (Vester): Jun, 77
VLF Listening Can Be Rewarding (Fischer): Jun, 78

World Above 50 MHz, The—Column (Pocock)

50-MHz Standings: Dec, 102
144-MHz Standings: May, 109
222-MHz Standings: Jul, 106
432-MHz Standings: Oct, 99
Beacons on 144 MHz and Higher: Aug, 84
CATV Cable on VHF and UHF: Jun, 94
Claimed North American Distance Records (Ward): Apr, 105
EME Annals: Mar, 104
How's Your Horizon?: Mar, 103
Invitation to an Aurora Experiment: Dec, 100
Limits of DX, The: Apr, 103
Meteor Season, The: Jul, 104
Microwave Standings: Jan, 107
One Year and Counting!: Jan, 105
Reallocation of the Microwave Bands: Nov, 105
Sporadic-E Mapping: May, 107
Strange Doings in the E Layer: Sep, 109
Transpacific Duct Breaks the Record Book: Oct, 97
VHF and UHF in Cuba: Feb, 97
VUCC Awards Update: Nov, 107

W1AW SCHEDULE:

Jan, 101; Feb, 44; Mar, 105; Apr, 103; May, 135; Jun, 119; Jul, 33; Aug, 80; Sep, 117; Oct, 55; Nov, 115; Dec, 120

YL NEWS—Column (Dunn)

Ain't Nothin' Wrong with Radio: Dec, 109
ATV: What is It, and What Can YLs Do with It?: Jan, 114
Becoming an Elmira: Jun, 101
CQ Cyberspace! Is There a YL Connection?: Nov, 111
Falling in Love with QRP: Apr, 113
Mini-History of YL Operations, A: Mar, 113
Mom, Morse & Language Skills: Sep, 119
YLs: You've Come a Long Way: Oct, 104

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OUR COVER

The article by Bill Welsh, W6DDB, on page 22 tells you how to start your own collection of Amateur Radio postage stamps. The cover photo and those accompanying Bill's article show a complete collection of Amateur Radio stamps and other postal items from around the world. The collection shown has been donated to the ARRL by Southwestern Division Director Fried Heyn, WA6WZO, and his wife Sandi, WA6WZN, for display at HQ. *(photo by Kirk Kleinschmidt, NT0Z)*

CONTENTS

January 1994
Volume LXXVIII Number 1

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

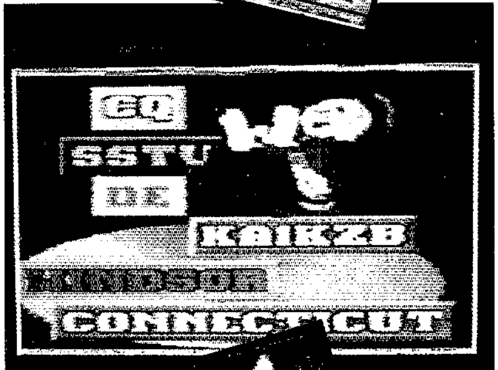
An Inexpensive SSTV System

Simple hardware and flexible software can provide you with an SSTV system capable of excellent results!

By Ben Vester, K3BC
4921 Bonnie Branch Rd
Ellicott City, MD 21043

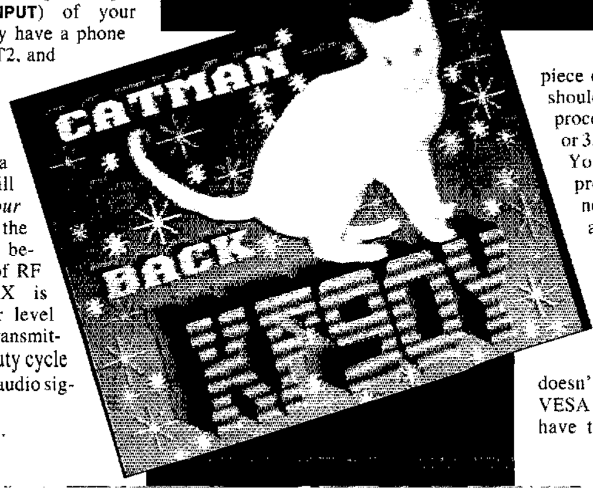
John Langner's great article on Slow-Scan TV[1] really got my interest. John's statement"...be sure to use an external crystal-controlled timing source...any attempt to use software timing loops is doomed to failure" was just too much of a challenge to ignore! I'd made some respec-table 64-color (gray shade) weather-fax receiving systems[2,3] for the Commodore and IBM computers using a simple interface between the receiver and computer. It didn't appear that demodulating an SSTV signal would be that much more difficult. What I wound up with is a color SSTV/FAX480/weatherfax system for IBM PCs and compatibles that is essentially 99% software! And this system *transmits*, too![4] I heartily recommend that you read John's article to fill in the gaps and learn something about the techniques and history of SSTV.

My work is aimed at the experimentally inclined, so if you're not familiar with BASIC programming, be prepared to learn a little about it if you want to maximize the utility of this system.



Hardware

Fig 1 (next page) shows a simple circuit used for receiving and transmitting. Connect the output of T2 to the phone patch input (often labeled **LINE INPUT**) of your transceiver. If you already have a phone patch, you can eliminate T2, and connect the line directly to the patch's phone-line terminals. All patches I know of employ transformer isolation, but a simple ohmmeter check will verify that is true of *your* patch. (I avoid using the transceiver's mike input because of the possibility of RF feedback problems.) RX is chosen to set the proper level for the audio going to the transmitter. We're using a 100% duty cycle signal, so you must set the audio signal to the transceiver at a level it can handle without overheating. With my transceiver, I went directly to the phone patch with an RX value of 43 kΩ.



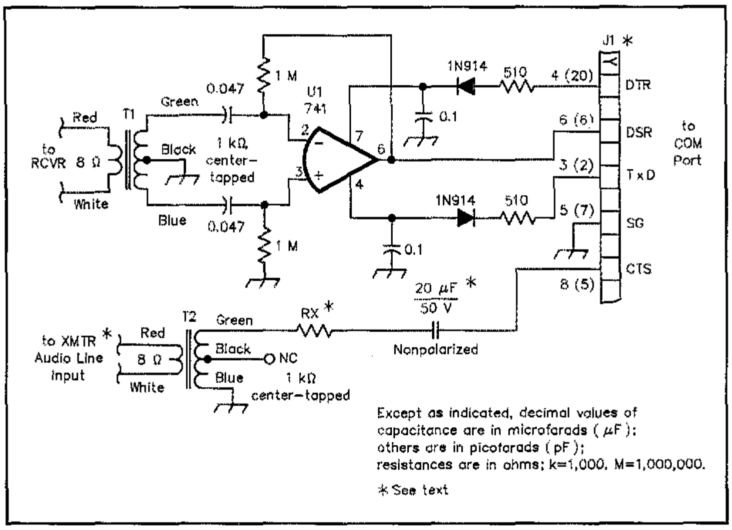
I've not included any low-pass filtering in the audio line between the computer output and transmitter audio input. My on-the-air checks with many stations reveal no additional external filtering is required when using SSB transmitters equipped with mechanical or crystal filters. If you intend to use this circuit with an AM or phasing-type SSB rig (or with VHF/UHF FM transmitters), audio filtering is required to provide the required spectral purity. An elliptical low-pass filter such as described by Campbell[5] should be adequate for most of these cases, but I have not specifically addressed this.

Circuit component values aren't critical nor is the circuit's physical construction. *Do* use a socket for the IC. A PC board is available from FAR Circuits,[6] but perfboard construction employing short leads works fine.

The Computer

The most important piece of hardware is the computer, which should have an 80286 (or better) microprocessor; a '386 machine running at 16 or 33 MHz definitely gives better results. You need a VGA color monitor that can provide a 640 × 480, 256-color noninterlaced display and a VGA (usually identified as SVGA) video adapter card that offers a 640 × 480 × 256-color mode.[7] The software directly addresses six of the most common SVGA chip types and also includes a VESA standard choice. If your video adapter card doesn't match one of the six, you'll need a VESA driver for your specific card. If you have trouble finding a driver, try some

[1]Notes appear on page 29.



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

* See text

Fig 1—Schematic of the simple SSTV receive and transmit circuit. This circuit is based on one that appears in the September 1991 Technical Correspondence column (see Note 3). T1 and T2 are Radio Shack 273-1380 audio-output transformers; The 20-µF, 50-V capacitor is a parallel combination of two Radio Shack 272-999 10-µF, 50-V nonpolarized capacitors; equivalent parts can be substituted. See text for value of RX. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. An optional low-pass filter can be used between the output of the computer and the transmitter's audio-input line (see text). At J1, numbers in parentheses are for 25-pin connectors; other numbers are for 9-pin connectors.

computer BBSes; I'm told that the Steve Rimmer BBS (tel 416-729-4609) has many video drivers.

Software

As with my earlier weatherfax programs, I've used GWBASIC as the programming tool. Although the guts of the program are contained in assembly language code (.ASM files), this code is available to the program (and you) through BASIC. All of the modifications to the core programs (.ASM files) that adapt them to the multitude of SSTV/FAX modes are accomplished using BASIC POKEs. This allows experimenters with even a limited knowledge of BASIC programming to make modifications that add other modes, etc. In deference to a few of my friends who complained about learning any BASIC, I have included a system configuration list in the programs. The program uses this list to determine which POKEs to make. This system is strictly keyboard controlled. The software uses a unique technique to get wider color definition than is normally available with a 256-color video card. The pictures in this article illustrate its effectiveness!

FAX480 Influence

Ralph Taggart, WB8DQT, introduced his FAX480 mode while I was in the midst of developing this program. With the flexibility of my software structure, I was able to add FAX480 capability to the available modes within a couple of hours. Since my processing was set up for 64 shades of gray, I kept it for this mode.

Some Program Details

One of the common SSTV practices is to retransmit a picture you just received so other SSTVers not copying the originating station can see the image. This capability is included.

RT.BAS is the receive and retransmit program. On receive, you simply choose the mode from a menu, and wait for the picture transmission to complete. As of this writing, Robot 36 and 72 modes are available in either a synchronous or a line-synced mode. Other modes (all synchronous) are Scottie 1 and 2, Martin 1 and 2, AVT90, AVT94, Wraase 96, FAX480 and weatherfax.

When receiving, if you fail to get the mode selection made in time to catch the frame sync, you can go directly to copying by pressing the keyboard's spacebar. On all but the AVT modes, the next line sync is picked up and starts the picture. The AVT modes copy out of sync. Because the program allows you to scroll horizontally across the RGB color frames, you can resync after the picture has been received. A few images I've copied have nonstandard color registration, so I also included the ability to adjust color registration after the picture is received. You also can save the picture—usually after you have scrolled the picture so the CRT screen frames just the part you want to keep.

TX.BAS is used for transmitting any picture file. When queried, you provide the mode and the file name, and after a brief pause while the picture loads, press G(o) to transmit. To avoid additional switching complexity, VOX transmitter switching is used.

VU.BAS allows you to view a picture. It has the same adjustments available as *RT.BAS*. One feature (applicable only to the Robot modes) is the ability to "retune" the picture (in 10-Hz increments) as you view its color balance.

SLIDESHO.BAS gives you the vehicle to display a bunch of pictures as a slide show. Place *SLIDESHO.BAS* in a directory contained in your PATH statement so it can be called up from anywhere.

TIFCONV.BAS converts 640 × 480, 24-bit color, TIFF pictures into a format that can be transmitted by any of the supported SSTV modes except Robot. In my experience, TIFF is the most common format used to transfer higher-resolution pictures between programs. I've used this program with the Computer Eyes/RT[8] and Software Systems Consulting[9] frame grabbers. The picture output from this program can be viewed with *VU.BAS* and, of course, is bound by 320 × 240 with 18-bit color.

LABEL.BAS allows you to add call signs and other text to the SSTV pictures. It takes any black-and-white TIFF (ie, 1-bit) file and creates a mask cutout where the black is. You can superimpose the cutout over an SSTV picture either in any color you want, or transfer a cutout of any background file you find interesting. The letters will then look like they were cut out of the background picture. Obviously, you can use squares or circles in addition to fonts to transfer a piece of one file onto another one. I use a cheap hand scanner to capture interesting fonts I find. You can get a three-dimensional effect by painting a color through the mask, then moving the mask a few pixels and rerunning the data through *LABEL* with a background file or another color. Or, run several different masks through *LABEL* in sequence to obtain different colors or patterns on different letters.

Summary

Here, then, is a brief description of a fundamental color-SSTV/FAX480 weatherfax system constructed almost entirely of software (I'm still working on it). The software is *free*, is not copy protected and can be obtained from the ARRL BBS (203-666-0578). Have fun—you've got lots to experiment with!

Notes

[1] J. Langner, "Slow-Scan TV—It Isn't Expensive Anymore!", *QST*, Jan 1993, pp 20-30.

[2] B. Vester, "C64 WEFAX Improvements," Technical Correspondence, *QST*, Jan 1988, pp 47-49.

[3] B. Vester, "Improved HF Weather Facsimile Programs," Technical Correspondence, *QST*, Sep 1991, pp 40-41.

[4] The software is available free from the author and can be downloaded as VESTER.ZIP from the ARRL BBS (203-666-0578).

[5] R. Campbell, "High-Performance, Single-Signal Direct-Conversion Receivers," *QST*, Jan 1993, pp 32-40. See also Feedback, *QST*, Apr 1993, p 75.

[6] FAR Circuits, 18N640 Field Court, Dundee, IL 60118-9269. The PC-board is $4.50, plus $1.50 shipping.

[7] Picture quality is degraded with an interlaced display. Few, if any, newer displays are interlaced at 640 × 480.

[8] ComputerEyes R/T by Digital Vision, Inc, 270 Bridge St, Dedham, MA, tel 617-329-5400, BBS 617-329-8387.

[9] Software Systems Consulting, 615 S El Camino Real, San Clemente, CA 92672, tel 714-498-5784, fax 714-498-0568.

Ben Vester was first licensed in 1945 and formerly held the call W3TLN. He retired in 1984 after a 34-year career with Westinghouse in the aerospace industry. He holds a BSEE from Virginia Polytechnic Institute, an MSEE from Johns Hopkins and an SEP from Stanford.

Some may remember his article, "Surplus-Crystal High-Frequency Filters" (QST *Jan 1959, pp 24-27), which was followed by the article "Mobile S.S.B. Transceiver"* (QST, *Jun 1959, pp 11-17 and 164) which used one of those filters and had about half the then-normal tube count. Two other articles written by Ben received* QST *Cover Plaque awards: "A Solid-State S.S.B. Transceiver"* (QST, *Jun 1963, pp 27-33) and "The Half-Square Antenna"* (QST, *Mar 1974, pp 11-14). Ben's other joy is sailing.* QST

The Nearly Perfect Amplifier

Have you ever bought a car and thought that everything about it was perfect? Or do you always wish it had a few other nifty or safety features? It's the same with ham power amplifiers. Here are some modifications that will add some nifty and safety features—to the amplifiers, not the cars.

By Richard Measures, AG6K
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Somis, CA 93066

After the article "Circuit Improvements for the Heath SB-220 Amplifier" was published in the November and December 1990 issues of *QST*, I began to receive letters and phone calls from Amateur Radio operators who were contemplating buying an HF amplifier. They wanted to buy an amplifier that needs no circuit improvements—that is, a perfect amplifier. What follows is a discussion of some design features that—in my opinion—would be present in a perfect, or a nearly perfect, amplifier.

Cathode Considerations

In a directly heated cathode, a ditungsten carbide layer on the hot tungsten (alloyed with about 1.5% thorium) filament wire emits electrons. In an indirectly heated cathode, the filament (the heater) heats a nickel cylinder that is coated with strontium oxide and barium oxide. This coating is relatively frangible, but highly emissive.

Ditungsten carbide is commonly made by heating tungsten in an atmosphere of acetylene (C_2H_2) gas. Carbon atoms in the gas break their bonds with hydrogen atoms and bond with tungsten atoms to form ditungsten carbide on the surface of the filament wire. Since it is atomically linked to the underlying wire, the ditungsten carbide layer is very durable. During use, the process reverses. Ditungsten carbide gradually loses carbon and changes back to tungsten. Extra heat exponentially accelerates this process. The cathode is worn out when the carbon is mostly used up.

After their cathodes are worn out, large external-anode amplifier tubes are commonly "recarburized" with acetylene, vacuum pumped, and resealed. This restores full emission. Although it is possible to recarburize a 3-500Z, doing so is not economically feasible.

Each type of cathode has advantages and disadvantages. A nickel cylinder has much-less inductance than a tungsten wire. Directly heated cathodes are relatively poor performers above the low VHF range, whereas some indirectly heated cathodes can perform satisfactorily at 2500 MHz. A directly heated cathode typically warms up in less than one second, while few indirectly heated cathodes can warm up safely in one minute (three to five minutes is not uncommon). However, the major disadvantage of indirectly heated cathode amplifier tubes is cost. In terms of dollars per watt, they are much more costly than 3-500Zs.

Cathodes deserve respect. Filament voltage and filament inrush current are areas of special concern.

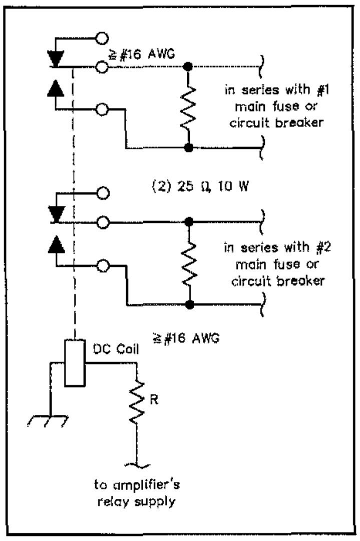


Fig 1—Step-start circuit for amplifiers that are designed for dual-voltage operation (120/240 V). In effect, the timing capacitor for the circuit is the high-voltage filter capacitor. The relay's coil voltage is approximately the same as the dc output voltage of the amplifier's relay supply. A contact rating of 10 A is adequate for 1500-W amplifiers Adjust R for relay pull-in when the high voltage reaches about 65% of normal after turn-on. Typical pull-in time is 1 ± 0.5 second. R will be roughly half of the relay's dc coil resistance.

Filament Voltage

For optimum life from a directly heated cathode, the filament voltage should be just above the voltage where PEP output begins to decrease. As the cathode ages, filament voltage needs to be increased gradually to restore full output. Using this technique, commercial broadcasters typically achieve an operating life of *22,000 hours* from amplifier tubes with directly heated cathodes.

According to Eimac's *Care and Feeding of Power Grid Tubes*, every 3% rise in directly heated cathode filament voltage results in a 50% decrease in life due to carbon loss. Yes, each additional 3% rise in filament voltage decreases the life *by half*. Expressed mathematically, cathode life is proportional to $[E_1/E_2]^{23.4}$, where E_1 is the lowest filament voltage at which normal PEP output is realized and E_2 is the increased filament voltage.

It's easy to make the filament voltage adjustable when the filament is powered by its own transformer. All that's needed is a small rheostat in series with the primary. For dual-voltage, dual-primary transformers, a dual, ganged rheostat is required. When the filament is powered by a winding on the high voltage transformer, making the filament voltage adjustable is more difficult since the rheostats must be connected to the low voltage secondary winding—and dual, ganged 0.01-Ω, 30-A rheostats are not to be found in your local Radio Shack.

An indirectly heated cathode can be permanently damaged by being operated *below* its rated minimum filament voltage. When operated *above* its maximum filament voltage rating, an indirectly heated cathode quickly boils off emissive material. Errant emissive material is bad news when it lands on the grid. For maximum cathode life in HF service, an indirectly heated cathode should be operated at the rated minimum filament voltage. This can be accomplished best with a regulated dc supply. Set it once and forget about fluctuations in the electric mains voltage.

Filament Inrush Current

Directly heated cathode filaments are commonly two vertical meshing helices (coils) of tungsten wire that are suspended by their ends (see September 1990 *QST*, page 15). The conductance of tungsten at room temperature is about 8.33 times the conductance at its normal operating temperature. Thus, the start-up current for a 15-ampere filament can be 125 amperes.

In a high-amplification tube such as the 3-500Z, the filament helices clear the grid cage by a distance of only a few thousandths of an inch. If the position of the filament changes, a grid-to-filament short may result. Therefore, it is prudent to limit filament inrush current in order to minimize thermal and magnetic stresses. For many of its smaller directly heated cathode amplifier tubes—such as the 3-400Z and 3-500Z—Eimac recommends that filament inrush current be limited to no more than double the normal current.

Since the grid-to-cathode clearance in an indirectly heated cathode is not affected by movement of the heater inside the rigid nickel cylinder, indirectly heated cathodes are not affected by inrush current.

Measurements

I tested the filament inrush current in a popular factory-built MF/HF amplifier with a pair of 3-500Zs. Since each tube's operating filament current is 14.7 A, the inrush current should not exceed 29.4 A, but I measured 34 A of inrush current per filament. Eimac rates the filament voltage at 4.75 V minimum to 5.25 V maximum but the filament voltage measured 5.31 V. With 4.8 V instead of 5.31 V, the useful life of the 3-500Zs would be about *ten times* longer. I measured the filament voltage in a factory-built single 3-500Z amplifier. The filament voltage was over 5.7 V. At this voltage, the cathode would probably be worn out in 400 hours of operation—and this one was.

The 8877 has a filament voltage rating of 4.75 V minimum to 5.25 V maximum; for HF communications service, the optimum filament voltage is 4.75 V. One popular commercial amplifier operates its 8877 filament at about 5.95 V when the amplifier is operated at the US-standard 120/240 V. Operating an 8877 at a filament voltage of 5.95 V is recommended for those people who have more money than brains.

A Simple ac Voltmeter

For measuring ac filament voltage, linearity at the low end of the meter scale is not important. With this in mind, designing an ac voltmeter using a dc meter movement is much easier than would otherwise be the case. All that's needed is a half-wave rectifier using a Schottky diode, a capacitor and a few resistors. The meter can be the amplifier's multimeter. All that the operator needs is two marks on the meter scale—one for minimum voltage and the other for maximum voltage.

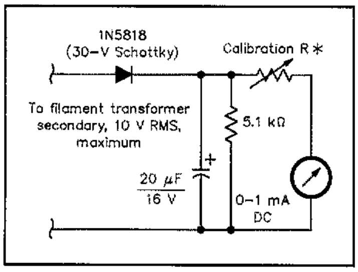


Fig 2—How to measure the ac filament voltage with a dc meter. Calibrate R with a standard voltmeter connected to the filament pins on the tube socket, so as to compensate for the voltage drop in the filament choke.

Grid Protection

I have performed autopsies on too many kaput amplifier tubes that died in HF amplifiers. Some of these tubes had damaged grids—but the damage was the unique type that is caused by VHF or UHF current. I have never found a grid that was damaged by excessive HF grid current. Perhaps this isn't so strange. I'm sure it's possible to roast a grid; tuning up key-down for a couple of minutes on 10 meters with the load capacitor set for 40 meters comes to mind. This would result in high grid current and almost no RF output. However, since most people—myself included—tune a grounded-grid amplifier for maximum RF—and maximum RF virtually coincides with normal grid current—very few people are likely to overheat a grid. Thus, complex electronic grid-protection circuits are unnecessary. A major disadvantage of electronic grid-protection circuits is they are not effective against the major source of grid damage—sudden, large bursts of VHF or UHF grid current. A more foolproof method of protecting the grid is a fuse or fuse resistor. Carbon-film resistors make good grid fuses.

Glitch Protection

During a major glitch, the anode (plate) current meter is subjected to a current surge as the HV filter capacitors discharge. Such a current is typically *several hundred* peak amperes—not exactly courteous treatment for a 1-A meter. However, the peak current will be much higher if a resistor is not used to limit the short circuit current that can be delivered by the HV filter capacitors. The current-limiting resistor is placed in series with the positive output of the filter capacitors. A 10-Ω, 10-W wire-wound resistor is adequate for up to 3 kV and 1 A. Since the current-limiting resistor will be dissipating many kilowatts during a major glitch, it should be a rugged glass-coated (ie, vitreous) type. If a glass-coated resistor opens during a major glitch, it won't be throwing large chunks of shrapnel around—like a rectangular ceramic-cased resistor often does.

If the positive high voltage briefly arcs to chassis ground—because of lint, a tiny insect, an intermittent VHF parasitic, or an errant hair—the negative high-voltage circuit will try to spike to several kilovolts negative. In the real world, this type of glitch is not an uncommon occurrence. Anything that gets in the way of the negative spike may be damaged. Since the grid-current meter is normally connected between chassis ground and the negative high-voltage circuit, the meter can be exposed to kilovolts at hundreds of amperes. I heard about one grid current meter in a homebrew amplifier that *exploded* during a glitch. The glass from the meter landed on the floor.

The easiest way to protect a current meter is to connect a silicon rectifier diode across it or across its shunt resistor. Usually, only one diode (connected with its cathode band to the meter's negative terminal) is needed in parallel with a meter.

It may take more than one diode to protect a meter shunt resistor. A silicon diode begins to conduct at a forward voltage of about 0.5 V. To avoid affecting meter accuracy, the operating voltage per protection diode should not exceed 0.5 V. For example, a 1-Ω shunt, for a reading of 1 A full-scale, has 1 V across it. Thus, two protection diodes would be needed to preserve meter accuracy. If the shunt resistor for a 1-A full-scale meter is 1.5 Ω, three diodes are needed.

Protection diodes should not be petite. Big, ugly diodes with a peak current rating of 200 A or more are best. I have seen smaller diodes—and the meter they were supposed to be protecting—literally blown away by a glitch. After some bitter experiences with lesser diodes, I began using the 1N5401. In small quantities, the 1N5401 costs about 20 cents each. It is rated at 200 A for 8.3 ms, 3 A rms, and 100 PIV. Other diodes from the 1N5400 family will work as well. During an extremely high current surge, a glitch-protection diode may short out—and, by so doing, still protect the precious parts. Replacing a shorted protection diode instead of a blown meter is almost fun.

A brief high-voltage flashover can damage an indirectly heated cathode tube. Here's how: In many amplifiers, one side of the filament/heater is grounded. The cathode is connected to the negative HV circuit. If the negative HV spikes to several kilovolts, the cathode will arc to the grounded filament. At a minimum, this breaks down the insulation between the heater and the cathode. Sometimes the heater wire burns out—and sometimes the cathode arcs to the grounded grid. Either way, the tube is kaput. There have been many 8877s and other indirectly heated cathode amplifier tubes that died this way—all for lack of 60 cents worth of glitch-protection diodes.

So why don't manufacturers of such amplifiers protect the negative HV circuit from spikes? The answer is an electronic

Catch-22. Even though it's likely that no amplifier manufacturer has ever seen a grid that was damaged by HF grid current, they seem to feel that electronic over-current protection for the grid is important. However, electronic over-current "protection" circuits for grids are not compatible with things that limit the voltage from the negative HV circuit to chassis ground. Thus, in an attempt to protect the amplifier tube from one *perceived* problem, designers leave the tube vulnerable to assassination from *common* occurrences.

To prevent the negative HV circuit from spiking to several kilovolts, connect a string of 200-A (or greater) glitch-protection diodes from the negative terminal on the high-voltage filter capacitor to chassis. Each diode will limit the voltage across itself to about 1.5 V. Typically, three diodes are needed—thus limiting the spike to about 4.5 V. The diode polarity is with the cathode band toward the negative high voltage. With one simple wiring change, the same string of diodes can also protect the grid-current meter and the anode-current meter.

Power Supplies

Transformers

Virtually all transformers use paper to separate and insulate each layer of windings, and paper is hygroscopic—it absorbs water vapor from the air. The presence of water reduces the insulating ability of the paper. In time, insulation breakdown is likely. The solution is to pot the windings; plastic resins are best for potting, and petroleum tar is next best. Since potting fills up the air spaces in the windings—and air is a poor heat conductor—potting improves heat transfer, thereby reducing internal temperature and decreasing the likelihood of failure. Potting adds very little to the initial cost of a transformer and subtracts substantially from the long-term cost by increasing the transformer's lifetime.

Filters

Capacitor-filtered power supplies are the norm in Amateur Radio amplifiers, because of their light weight, relative ease of obtaining high dc potentials, good transient-current voltage regulation (a must for SSB), and cost. Since choke and swinging-choke filters can not handle transient current loads, the only alternative for SSB use is a resonant-choke filter, which has advantages and trade-offs. A resonant-choke filter is tricky to tune, heavy, and expensive, and it requires a much-higher bleeder current than a capacitor filter requires. However, a resonant-choke filter demands only about one-sixth as much peak power from the electric mains as a capacitor filter demands. This means that for 120-V operation, where power is much more limited than with 240-V operation, a resonant-choke filter is clearly the better choice.

Rectifiers

The most frequent failure mode for high-voltage power-supply rectifiers is too much reverse current. This problem can be virtually eliminated in full-wave rectifier circuits by making sure that the total peak inverse voltage in each string of diodes exceeds the peak secondary voltage of the high-voltage transformer by a comfortable margin; 50% sounds comfortable to me.

Modern solid-state rectifiers are made differently than they were 30 years ago. In those ancient days, rectifiers did not have uniform capacitance. In an attempt to help equalize the peak reverse currents in series-connected rectifiers, a parallel resistor and capacitor was connected to each rectifier. It was felt that swamping 50 to 100 pF with 10,000 pF (0.01 µF) would help. In practice, it didn't work too well since the tolerance of the capacitors used was typically −20% to +80%. Another problem is that the resistors that were typically used (470 kΩ, 0.5 W) were rated at 250 V absolute maximum. It is hardly safe to use one of these resistors with a 1000-PIV rectifier. As a result, "equalizing" did more unequalizing than anything else. Even after rectifier technology improved, people hung on to the old habit of using parallel resistors and capacitors.

There is a flaw in the logic behind using rectifier equalization. In any series circuit, the currents in all of the elements are exactly equal. Thus, when rectifiers are in series, the reverse-current burden is exactly the same for each rectifier. How is it that something that is already exactly equal needs to be equalized? It should go without saying that series-connected rectifiers should always be of the same type. Mixing rectifiers with different junction capacitances can cause a problem.

There is one instance where equalizing resistors and capacitors are a good idea. Voltage spikes come in two flavors—positive and negative. In a full-wave capacitor-filter rectifier circuit, the energy from positive and negative voltage spikes is simply rectified and harmlessly stored in the filter capacitor. However, in a half-wave rectifier circuit only one polarity is rectified. A voltage spike of the other polarity cannot be absorbed by the filter capacitor. Instead, the potentially destructive spike appears across the rectifiers. Placing a capacitor across each rectifier helps to limit reverse spikes. A better solution is to connect a metal-oxide varistor across each half-wave rectifier—or to use a full-wave rectifier circuit.

Electrolytic Capacitor Equalizing Resistors

A resistor's voltage rating takes precedence over its power rating. One-watt carbon-composition resistors have a maximum voltage rating of 350 V. For example, it takes 469 V to dissipate 1 W in a 220-kΩ resistor ($E=[PR]^{0.5}$). In the past, some amplifier engineers decided that 220-kΩ, 1-W carbon-composition resistors would make good (and cheap) voltage-equalizing resistors for 450-V electrolytic capacitors—considerable overvoltage for 1-W parts! When a carbon-composition resistor is operated above its maximum voltage rating it changes resistance—exactly what you don't want in a voltage divider. Capacitor failure is likely in a string of equal-value electrolytic capacitors connected in series when the voltages across the capacitors is not the same. Even when they are operated within their voltage rating, carbon-composition resistors change resistance with age. Thus, 2-W carbon-composition resistors, which are rated at 500 V, are not the answer. Metal-oxide-film (MOF) resistors are far superior to carbon-composition resistors. A 3-W, 100-kΩ MOF resistor makes an excellent equalizer resistor for 450-V capacitors. Lower values of resistance create extra heat—something that electrolytic capacitors do not tolerate well.

Biasing

The operating bias in most amplifiers is not adjustable; a single Zener diode is typically used. The resulting zero-signal anode current (the idling current) is seldom optimum. Adjustable bias would be nice. The solution: obtain the operating bias from a series string of forward-biased rectifier diodes. By switching the number of diodes in and out with a rotary switch, the bias can be changed in approximately 0.7-V increments.

Another area that could be improved on is the method of bias switching between receive and transmit. In this modern age there is no reason to use a pokey, noisy mechanical relay to switch bias. An opto-isolator coupled to a transistor switch can do this job better and cheaper. An electronic bias switch is more than fast enough to keep up with modern high-speed RF relays.

Electronic bias switches that are RF actuated create two problems. The amplifier tube switches between linear bias and nonlinear bias during softly spoken syllables of speech, causing choppy-sounding audio and splatter. These two problems are eliminated when the electronic bias switch is controlled by the current that passes through the RF-relay's coils.

High-Speed Relays

The switching time of a conventional relay is 15 to 25 ms—switching in a somewhat stentorian manner. Such relays have traditionally been used for RF and bias switching in HF amplifiers. This was acceptable when transceivers also used conventional relays. Currently manufactured transceivers are designed for AMTOR, QSK CW, and reasonably quiet SSB-VOX operation. Such transceivers switch quickly and quietly. Japanese transceivers often use a Matsushita (Panasonic) NR-HD-12V reed relay to switch the RF output. Provided that the SWR isn't ridiculous, this relay can dependably switch 150 W of RF. It works well in an HF amplifier when it is used to switch the input RF signal. This relay is available in the US as a spare part for Kenwood and Yaesu radios.

Although Matsushita refuses to sell its NR-HD-12V relay to US Amateur Radio

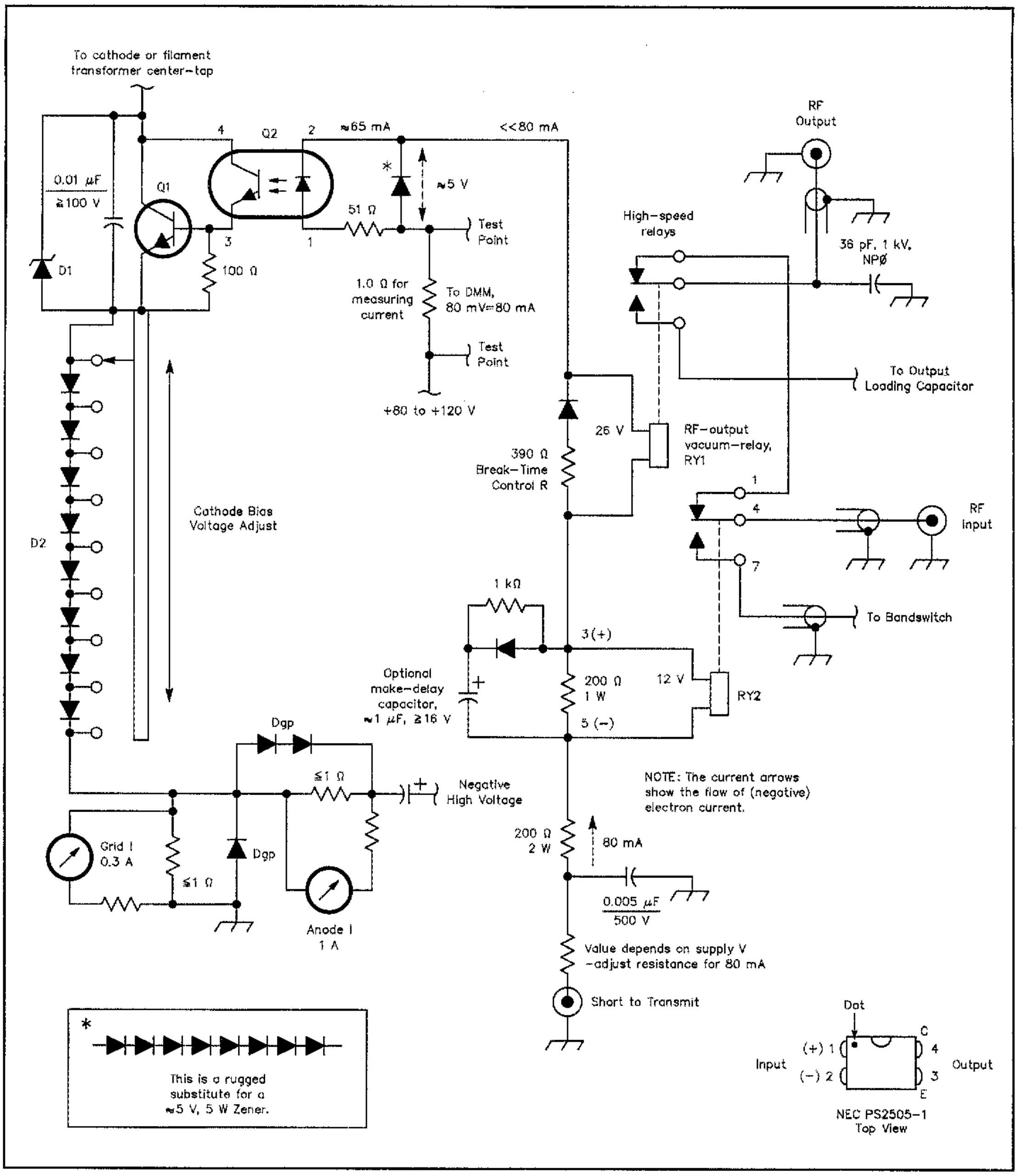


Fig 3—Adjustable electronic cathode bias switch for cathode currents up to 3 A.

D1—30- to 51-V transient voltage suppressor (Digi-Key part 1.5KE51CA-ND or similar; Digi-Key's toll-free number is 800-344-4539).

D2—For collector currents between 1 and 3 A, use 1N5401 diodes. The number of diodes making up D2 may have to be increased in some amplifiers.

Dgp—Glitch-protection diode (1N5401 or similar, 200-A peak or greater).

Q1—NPN, 80 to 100 V, 10 A (TIP 33B, TIP 33C, etc). (The collector-to-emitter voltage drop across Q1 is about 1.4 V on transmit. On receive, the drop depends on the supply voltage to the tubes and also on the particular tubes used; typically the drop on receive is 24 to 30 V. Q1 needs only a minimum heat sink.)

Q2—NEC PS2505-1 optoisolator, input-protected, 80-mA maximum input, Vce=80 V, gain of 300% or greater.

RY1—Kilovac HC-1 or Jennings RJ-1A (7 A maximum at 32 MHz).

RY2—Matsushita NR-HD-12V reed relay (Trio-Kenwood part number S51-1429-05; $21 plus shipping and handling, as of August 1993).

The unspecified diodes are 1-A or greater, 200 V PIV (1N4003, 1N5401, etc).

The unspecified resistors are ¼-W.

equipment manufacturers, a US reed relay manufacturer could probably come up with a similar relay, if there were enough demand. Jennings and Kilovac manufacture high-speed relays that will switch kilowatts of RF. The Jennings relay is the RJ-1A; Kilovac's is the HC-1. I have been using these two relays for many years to switch the output in my amplifiers. When mounted with silicone rubber, they are fairly quiet. When used with a speed-up circuit, either relay can switch in under 2 ms. I use a speed-up/sequencing circuit that prevents them from being hot-switched, and have not had a relay fail in this circuit.

It makes little sense to be currently building—or buying—an amplifier that switches in 25 ms. If such an amplifier is used with a modern radio, it would be technically more correct to say "hot-switches in 25 ms."

VHF Stability

On page 72, the 1926 edition of the ARRL's *The Radio Amateur's Handbook* told us how to build an improved VHF parasitic suppressor—one that provides better VHF stability than ordinary parasitic suppressors. The logic was elementary. A suppressor is supposed to dampen a circuit. Since low Q is synonymous with high dampening, build a suppressor with low Q. Instead of using a conductor with a high Q at VHF—such as copper or silver—use a conductor with low Q at VHF—resistance wire. The 1926 *Handbook* said, "The combination of both resistance and inductance is very effective in limiting parasitic oscillations to a negligible value of current."

After 1929, someone forgot to include this information in the *Handbook*. In those days, the oversight probably didn't matter very much. Electron tubes generally had poor amplification at VHF, so VHF instability was not much of an issue. During the ensuing decades, Amateur Radio operators and amplifier manufacturers got into the habit of using parasitic suppressors made from copper—or, even worse—silver-plated copper. This was an easy habit to get into; copper and silver can be soldered easily and cheaply. Meanwhile, the VHF amplification of electron tubes kept improving. Modern tubes need 1926-vintage parasitic suppressors with a low Q at VHF!

VHF parasitic oscillation can cause bandswitch arcing, tuning-capacitor arcing, and a large pulse of grid current. The pulse of grid current is so large that a powerful magnetic force is exerted between the grid and filament. In a 3-500Z, this force is capable of bending the hot tungsten filament helices and causing a filament-to-grid short. Of course, there is a trade-off to using suppressors with a low Q at VHF. On the 10-meter band, they reduce amplifier output by roughly 0.08 dB. This should come as no surprise: anything that dampens VHF resonance is bound to have some effect at 29 MHz. A VHF suppressor that does not get hot on 10 meters isn't doing its job.

Even though much has been published about VHF parasitic oscillation, amplifiers are still being built without the benefit of suppressors with a low Q at VHF. Some amplifier manufacturers presently take a dim view of using such suppressors. Even if you have signs of parasitics, such as intermittent bandswitch arcing, they will void the warranty if you remove their suppressors with a *high* Q at VHF and install replacements with a *low* Q at VHF.

Step-Start

Large power supplies need something to soften the shock of start-up. A 10-A DPST (normally open) or a 10-A DPDT relay and two 25-Ω, 10-W resistors are just about all that's needed to add a good step-start circuit to the average 1500-W amplifier. The step-start circuit belongs in series with the main fuses or circuit breakers. This way the filaments will also enjoy the benefit of a gentle start-up.

Is More Gain Always Better?

Today, the more or less standard in transceiver output is 100 W. There are amplifier tubes that can easily be ruined by 100 W of drive. A good example is the 3CX800A7. Using 100 W of drive will eventually strip flakes off the cathode. The flakes can become lodged between the cathode and the grid cage—creating a short. Even a pair of 3CX800A7s is clearly overdriven by 100 W. Doing so probably won't flake the cathodes, but it can cause rotten splatter. The fix is simple: connect a 40-Ω resistor in series with each 3CX800A7 cathode. These (cathode RF negative feedback) resistors reduce gain. As a result, the amplifier won't be driven above its absolute maximum ratings—and into nonlinearity—by a 100-W transceiver.

Cathode RF negative feedback resistors are better than having a matched pair of 3CX800A7s—the cathode currents automatically equalize themselves. And unlike all ALC circuits, cathode feedback resistors work instantaneously—eliminating ALC's generic flaw—leading-edge splatter on SSB. (Amplifier-to-transceiver ALC only works properly on modes with a constant signal level, such as RTTY and FM.) When a single 3-500Z is driven by 100 W, it also splatters. Although the rated drive is around 55 W, manufacturers of amplifiers with a single 3-500Z give the green light to driving their amplifiers to an anode current of up to 550 mA—a feat that requires using about 100 W of drive. On SSB, this produces distortion and splatter. Eimac rates the anode current at 400 mA *absolute maximum*. It takes a 25-Ω cathode feedback resistor to make a 3-500Z happy with 100 W of drive. The resistor goes in series with the cathode coupling capacitor.

It would be nice if amplifiers were designed to be compatible with 100 W of drive.

Adjustable Tuned Inputs

Modern MF/HF transceivers with a solid-state output stage use an untuned push-pull RF output stage. In order to meet FCC requirements on spurious emissions, passband LC filters are used. Such filters introduce inductive and capacitive reactance at various frequencies within their passbands. In other words, the output impedance of a modern transceiver is seldom 50 ± *j*0 Ω. This is of no consequence unless you happen to be driving a tuned input in a grounded-grid amplifier. In this case, the filter reactance interacts with the reactance of the input capacitor in the pi-network tuned input. The length of the coax between the filter and the tuned input affects the way the reactances interact. As a result, the input SWR can suffer. If the SWR is too high, the transceiver will automatically cut back on power.

For example, to obtain Eimac's recommended Q of 2, approximately 200 pF of input capacitance is needed for a tuned input on the 10-meter band. In actual practice, however, a 50-pF input capacitor may produce the best SWR with a particular model transceiver and a particular length of coax. A different model transceiver or a different length of coax may require a different-value input capacitor.

It would be nice if an amplifier's input capacitors on the tuned inputs could be readily adjusted.

Summary

Here is a list of the various features discussed in this article and the approximate cost of including them in an amplifier design.

- Adjustable filament voltage and a simple filament voltmeter using the existing multimeter, $10 to $20.
- Step-start relay circuit, $7.
- Adjustable electronic bias switch, $2.
- High-speed RF relays, $90.
- Parasitic suppressors with a low Q at VHF, $1.
- Separate high-voltage and filament transformers—preferably potted; metal-oxide film equalization resistors for electrolytic capacitors; no so-called equalizers on high-voltage rectifiers, $60.
- Make the amplifier compatible with 100 W of drive, $1.
- Glitch-protection diodes for the meter circuits and the negative high-voltage circuit, 30 cents.
- A 10-Ω, 10-W glass-coated resistor in series with the positive lead of the high-voltage power supply, 76 cents.
- Adjustable input capacitors on the tuned inputs, $6.

Obviously there are other important elements to consider in amplifier design. The elements discussed above are ones that are frequently overlooked. If you have any questions or comments, please call. My new telephone number is 805-386-3734. QST

Computer-Controlled Electronic Test Equipment

Part 2[1]—This month, we build the motherboard and power supply—the physical foundation for all the test-equipment projects to come. As promised, the first of those projects is an L-C meter, also in this installment. We've got lots to do!

By Ron Portugal
52 Susan Lane
North Haven, CT 06473

In Part 1, we laid the groundwork for a number of pieces of computer-aided test equipment that work *with* the computer, but aren't installed *inside* the computer. This approach avoids using the much-needed expansion slots of the host PC. It also eliminates loading the PC's power supply and eliminates the potential of a test-equipment accident from crippling the computer. In this installment, we'll cover *two* projects: the Z8671 microcomputer's motherboard/power supply combination and an inductance/capacitance (L-C) meter.

Z8671 Microcomputer Motherboard and Power Supply

Connecting components to a microprocessor is frequently a chore involving a mess of wires, cables, power supplies and sundry items. In an effort to *almost* eliminate these inconveniences for the Z8671 test-equipment system, I developed a motherboard/power-supply platform. See Figs 1 and 2. The 12.5- × 7-inch single-sided PC board[2] supports nine male 50-pin headers (JP3-JP11) to accept 5- × 6-inch PC project boards and includes one 50-pin female socket (JP12) for bus expansion.

Also on the motherboard are three power supplies: +9 V @ 2.5 A unregulated, and regulated ±15 V @ 0.45 A; see Fig 3. A capacitor-input full-wave bridge configuration is used for the +9-V supply, and a dual output, step-up, flyback, switching regulator provides the ±15-V sources. Other than the transformers, 50-pin headers, 50-pin socket and the angle brackets, all the parts can be obtained from electronic component suppliers such as Mouser and Digi-Key.[3]

U1 is an adjustable switching regulator. Input voltage and ground are brought to U1 on pins 5 and 3, respectively. Pin 4, the switch output, drives the primary of T2, a 200-µH flyback transformer. T2's center-tapped secondary winding feeds two half-wave rectifier circuits (D3-D4, C5-C6 and R5-R6) that provide the ±15-V outputs. The +15-V output feeds a voltage divider (R3 and R4, a trimming pot) that connects to U1's feedback input (pin 2). R4 (**±15 V DC ADJ**) adjusts the amplitude of the ±15-V supply outputs. R2 and C4 stabilize U1.

U1 has built-in thermal-shutdown and current-limiting protection. Switching regulators usually don't dissipate much power. When I tested the circuit with a 0.55-A load attached to both 15-V outputs with a 120-V ac line input, however, U1's case got a little hot. A simple U-shaped heat sink keeps it cool.

Mechanical Assembly

Six power-cord strain-relief holes are provided on the PC board. Each of the ac power cord's three wires is woven through two of the holes provided it, then fed into the third row of solder-pad holes for attachment. Two holes marked **POWER SWITCH** connect the ac line to a front-panel power switch. Holes labeled **LED+** and **LED−** provide for connection of D2, the **POWER ON** LED.

For stability and strength, the motherboard's size demands that it be mounted on a chassis or frame. Plexiglas, aluminum or

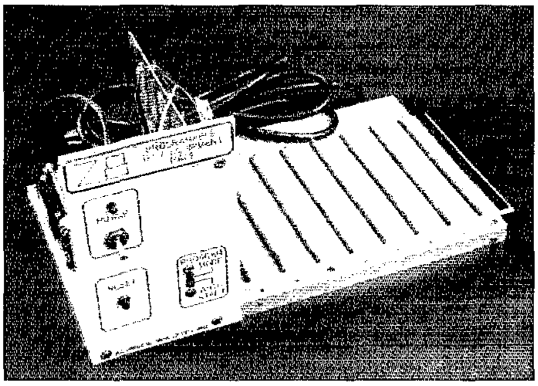


Fig 1—The panel supports the **POWER SWITCH, POWER ON** LED and the Z8671 CPU board's **RESET** and **AUTO/PROGRAM** controls. When finished, the L-C meter described in this month's installment will plug into the motherboard socket next to the CPU board and have a narrow front panel of its own. A section of split tubing added across the tops of the panels adds rigidity.

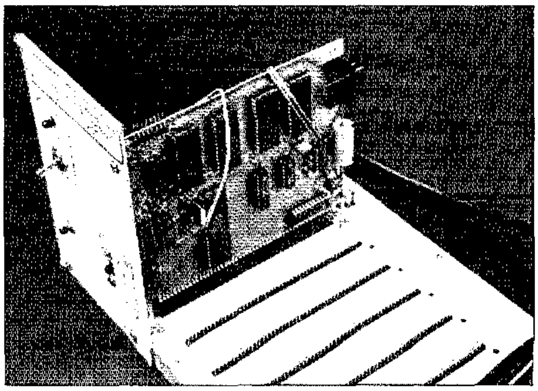


Fig 2—A side shot showing the power supply section on the motherboard.

[1] Notes appear on page 41.

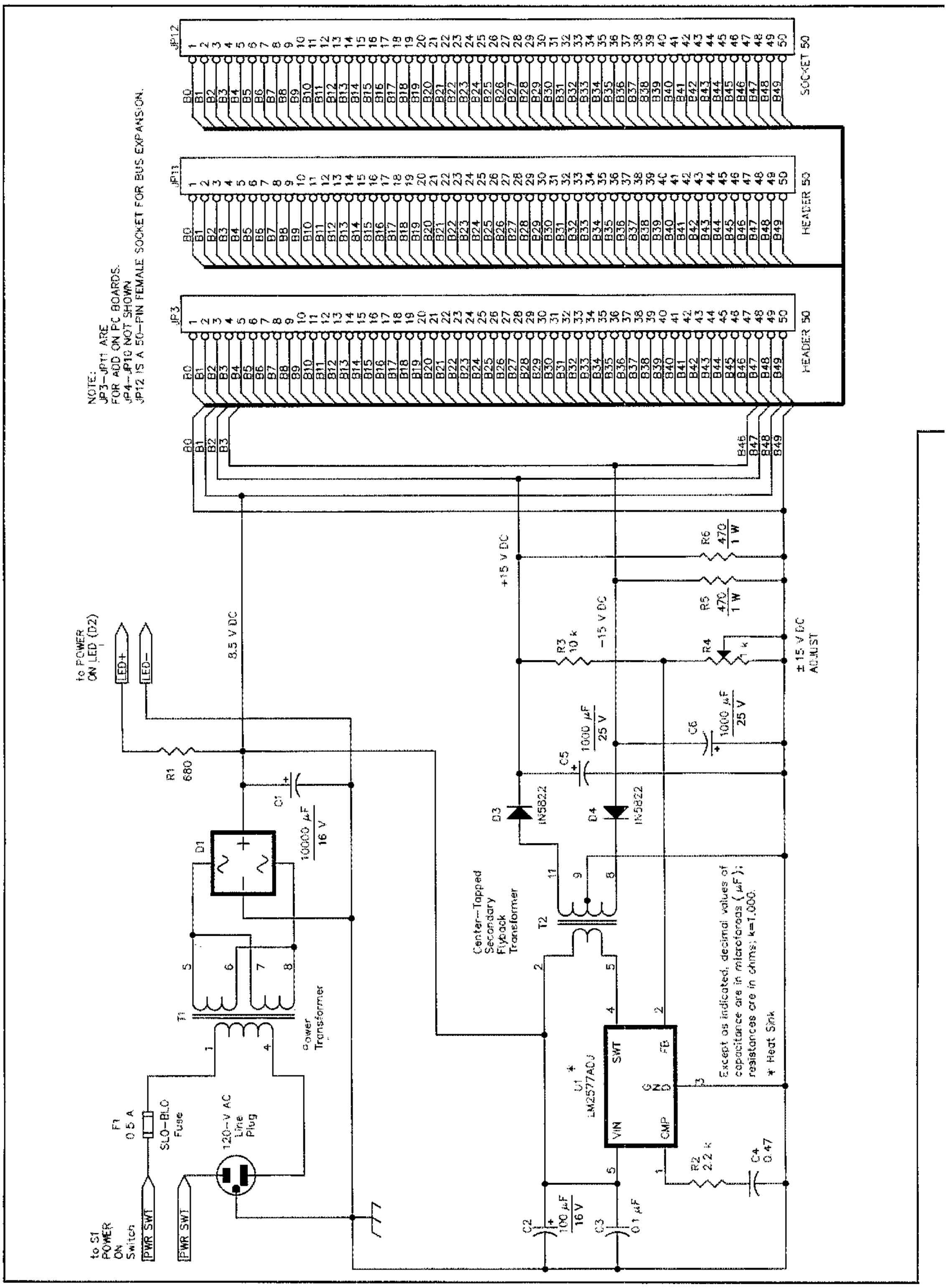

NOTE:
JP3–JP11 ARE
FOR ADD ON PC BOARDS.
JP4–JP10 NOT SHOWN
JP12 IS A 50–PIN FEMALE SOCKET FOR BUS EXPANSION.
JP3
JP11
JP12
B0 B1 B2 B3 B4 B5 B6 B7 B8 B9 B10 B11 B12 B13 B14 B15 B16 B17 B18 B19 B20 B21 B22 B23 B24 B25 B26 B27 B28 B29 B30 B31 B32 B33 B34 B35 B36 B37 B38 B39 B40 B41 B42 B43 B44 B45 B46 B47 B48 B49
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20 21 22 23 24 25 26 27 28 29 30 31 32 33 34 35 36 37 38 39 40 41 42 43 44 45 46 47 48 49 50
HEADER 50
HEADER 50
SOCKET 50
B0
B1
B2
B3
B46
B47
B48
B49
to POWER
ON LED (D2)
LED+
LED–
8.5 V DC
R1
680
C1
10000 µF
16 V
D1
T1
Power
Transformer
1
4
5
6
7
8
F1
0.5 A
SLO–BLO
Fuse
120–V AC
Line
Plug
to S1
POWER
ON
Switch
PWR SWT
PWR SWT
+15 V DC
–15 V DC
R3
10 k
R4
1 k
± 15 V DC
ADJUST
R5
470
1 W
R6
470
1 W
C5
1000 µF
25 V
C6
1000 µF
25 V
D3
1N5822
D4
1N5822
T2
Center–Tapped
Secondary
Flyback
Transformer
11
9
8
2
5
U1 *
LM2577ADJ
SWT
FB
GND
VIN
CMP
4
2
3
5
1
R2
2.2 k
C4
0.47
C2
100 µF
16 V
C3
0.1 µF
Except as indicated, decimal values of
capacitance are in microfarads (µF);
resistances are in ohms; k=1,000.
* Heat Sink

Fig 3—Schematic of the power-supply/motherboard circuit. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. The 50-pin male headers (JP3-JP11) are on 1-inch centers and two grounded mounting holes are provided for securing each PC board to the motherboard. Use Keystone Electronics Corp mounting brackets (part numbers 612 or 621).

D1—50-PIV, 3-A bridge rectifier.
D2—LED.
D3, D4—1N5822.
F1—0.125-A, 125-V fuse.
JP3-JP11—Male 50-pin header (Samtec TSW-150-14-L-S).
JP12—Female 50-pin socket (Samtec BCS-150-L-S-HE).
S1—SPST toggle switch (not shown).
T1—Power transformer (Magnetek-Triad part number F16-2250, 36-VA, 115-V, 50/60-Hz primary, dual 8-V RMS @ 2.25-A secondaries.
T2—Dual-secondary flyback transformer. (According to National Semiconductor, T2 is available from at least three sources: Renco Electronics Inc, 60 Jeffryn Blvd, Deer Park, NY 11729, tel 516-586-5566, part no. RL-2581; Pulse Engineering, PO Box 12235, San Diego, CA 92112, tel 619-268-2400, part no. PE-65301; AIE Magnetics, 2801 72nd St N, St Petersburg, FL 33710, tel 813-347-2181, part no. 330-0202.)

plastic angle brackets or an aluminum picture frame are a few options. If you mount the motherboard on a metal frame, *don't short the bus lines!* Insert a piece of Mylar or other insulating material between any metal chassis supports and the motherboard. T2 is relatively hefty, so support it with a couple of standoffs affixed to the right-hand transformer mounting screws.

The **POWER SWITCH, POWER ON** LED and the Z8671 CPU board's **RESET** and **AUTO/PROGRAM** controls are mounted on a small front panel. This panel is a 4-inch wide plate affixed to the left-front section of the motherboard. The panel's vertical dimension depends on the height of the chassis or frame used. An angle bracket secured between the upper mounting hole of the Z8671 board and the front panel keeps the assembly rigid and stable.

Power-Supply Adjustment

The 9-V supply requires no adjustment. When you're ready to test the switcher supply, however, first *set R4 to its maximum value,* 1 kΩ. The feedback-input pin voltage is compared to U1's 1.23-V reference. When R4 is set to 1 kΩ, the ±15-V supplies deliver about ±13.5 V. As R4's resistance is *decreased,* the output voltages *increase.* Meter the +15 V line and adjust R4 for a reading of +15 V. Then check the −15 V line to ensure that its voltage is also correct.

The L-C Meter

L-C meters are handy to have.[4] Nowadays, you can find commercially made L-C meters priced at about $60 and more. Usually the meters don't cover the full range of L-C values of interest to amateurs. Also, there's always the question of calibration: How accurate is the reading and what's the resolution of the reading?

Our computer-aided L-C meter has a resolution of better than four significant digits over the following ranges: capacitance, 1 pF to 10,000 µF; and inductance, 1 µH to more than 5 H.

The meter's accuracy depends on the components used during calibration. Importantly, both instruments can be calibrated (via programing) to compensate for variables that affect the instruments' readings.

Capacitance is measured in two ranges: 1 pF to 1 µF and 1 µF to 10,000 µF. Inductance is measured using a single range, 1 µH to 5 H. The capacitance meter can be made to be autoranging.

Inductance Measurement

Fig 4 is the L-C meter schematic. Let's start by examining the inductance-meter oscillator circuit in the lower left-hand corner of the drawing. Our inductance-measurement procedure first measures the frequency of an L-C oscillator having a known capacitance value, then, using the resonant frequency equation:

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

we compute the value of L:

$$L = \frac{1}{(2\pi f)^2 C} \qquad \text{(Eq 2)}$$

With the number-crunching power of a Z8671 microcontroller and a PC available, the value of L can be computed because we know the values of pi, the two capacitors and the oscillator frequency. If, instead of frequency, the *period* of the oscillator is known, Equation 2 becomes:

$$L = \left(\frac{T}{2\pi}\right)^2 \times \frac{1}{C} \qquad \text{(Eq 3)}$$

Measuring the period of a signal is done with three 4-digit counters: U9A, B, and C (the 82C54 triple-counter chip).[5] Two of the counters, U9C and U9B, are series connected and count a gated train of 7.3728-MHz clock pulses generated by the crystal oscillator located on the Z8671 CPU board. U9A counts 1000 periods of the oscillator and, in conjunction with U11A, a J-K flip-flop, generates the clock gate for cascaded counters. The number that ends up in U9C and U9B represents 1000 periods of the inductance-meter oscillator frequency.

At the end of the period measurement, the contents of the cascaded counters are massaged by the Z8671 and then sent off to the PC for further processing, correction and display. The conversion operations change the BCD numbers in the two 4-digit counters to ASCII format. Since the 82C54's counters count down from 9999d before ASCII conversion, each digit has to be subtracted from 9d. The whole process is done with a few lines of a BASIC/DEBUG program; more about this later.

The inductance and capacitance measurement processes are controlled by U14, a "subinstruction register." Appropriate bit combinations are sent to U14 from the Z8671 CPU under program control.

U12A (¼ of a CMOS quad **NAND** gate) is used as the amplifier portion of the Colpitts L-C oscillator. By connecting a high-value resistor, R23, from the input to the output of the gate, a linear gated amplifier is formed. Gated? Yes, the second input to the gate still controls the output; a zero on pin 1 of U12A disables the gate and causes its output, pin 3, to go high (+5 V).

This L-C oscillator configuration generates a square wave that complies with the resonant frequency equation:

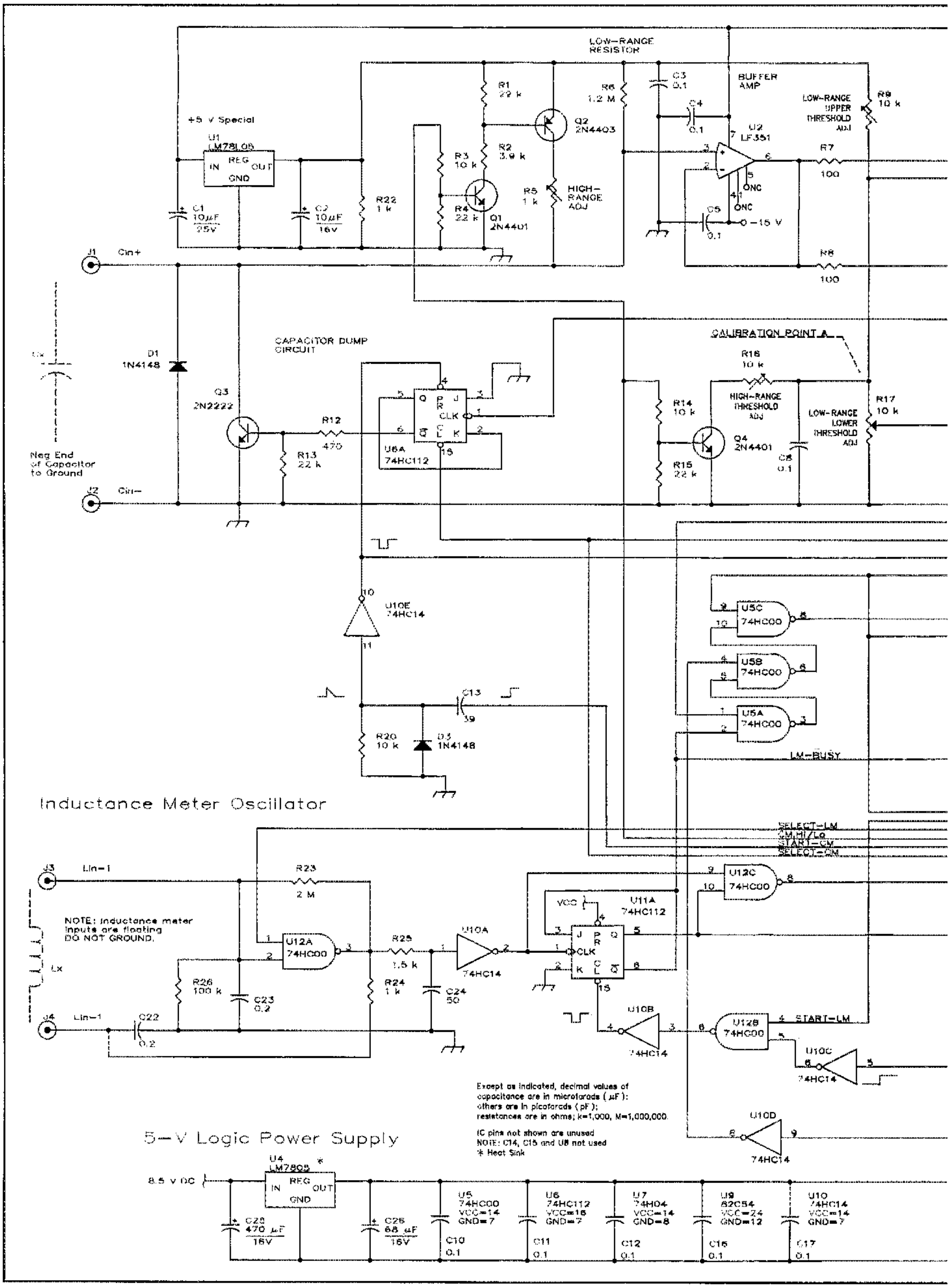
$$f = \frac{1}{2\pi\sqrt{L(C/2)}} \qquad \text{(Eq 4)}$$

Equation 4 assumes that C22 and C23 are identical in value. When C22 and C23 are 0.2-µF capacitors, the oscillator operates from about 100 Hz to over 1.8 MHz, enabling inductance measurements to be made from 1 µH to over 5 H. R24 is a drive-limiting resistor; R26 ensures that the oscillator is quiescent when the L_X inputs are floating. A low-pass filter consisting of R25 and C24 (with a 2.1-MHz −3 dB point) and a Schmitt trigger circuit, U10A, prevent high-frequency ringing from generating spurious trigger pulses.

Capacitance Measurement

The capacitance-measurement strategy depends on the fact that when a capacitor is charged from a known, fixed voltage through a known, fixed resistance, the time interval between two predetermined threshold levels along the charging curve is directly proportional to the value of the capacitor (see Fig 5).

The time interval between V_L and V_U is:



Inductance Meter Oscillator

5-V Logic Power Supply

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

IC pins not shown are unused
NOTE: C14, C15 and U8 not used
* Heat Sink

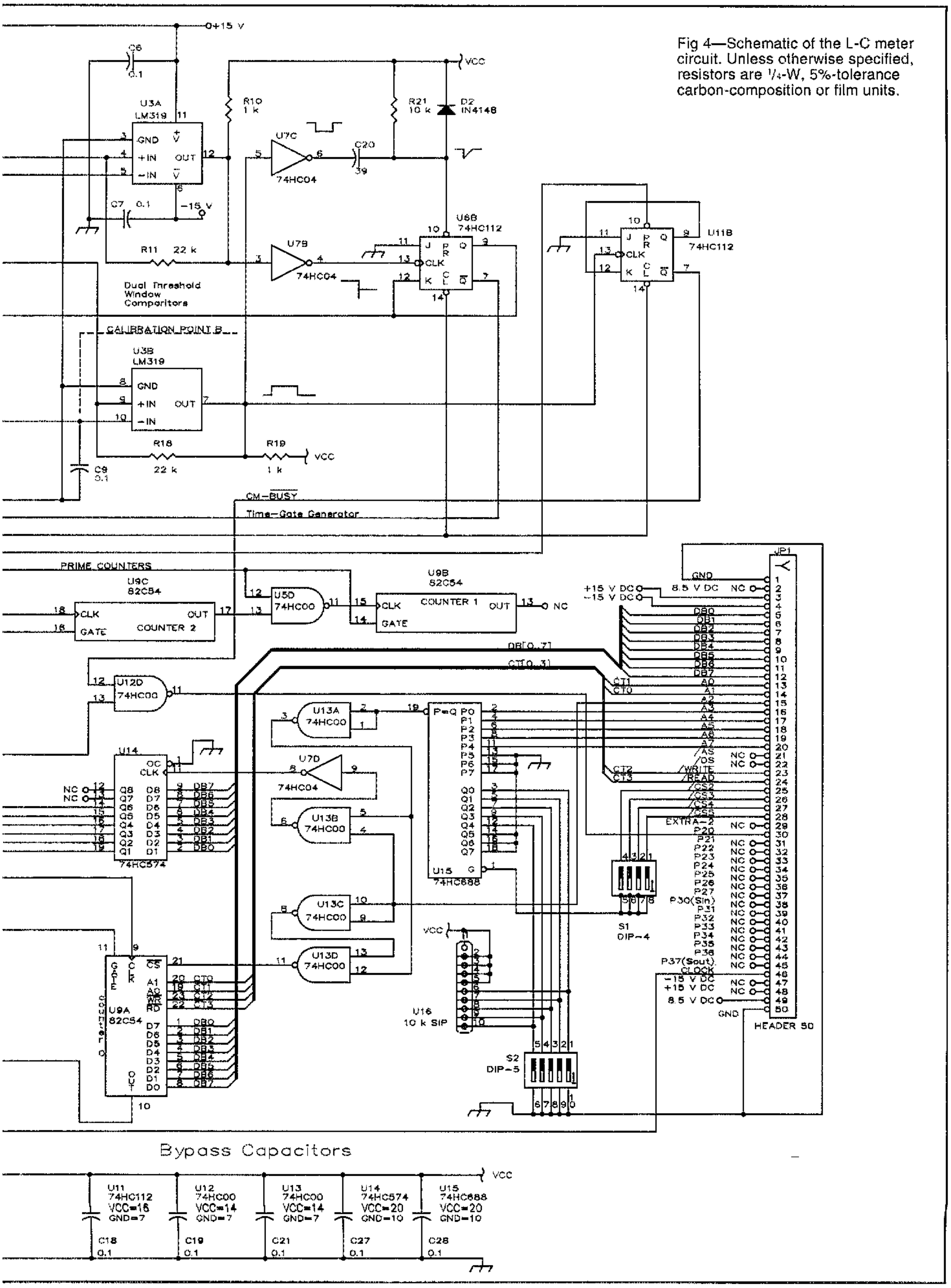


Fig 4—Schematic of the L-C meter circuit. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

$$\Delta T = RC \ln\left(\frac{E - V_L}{E - V_U}\right) \qquad \text{(Eq 5)}$$

Since R, E, C, V_L and V_U are all constants,

$$\Delta T = K \times C \text{ and } C = \frac{\Delta T}{K} \qquad \text{(Eq 6)}$$

where

$$K = R \ln\left(\frac{E - V_L}{E - V_U}\right) \qquad \text{(Eq 7)}$$

All we do is pick values for E, R, V_L and V_U, devise a way of measuring the time interval between the two threshold points along the charging curve, and we're all set.

The upper part of Fig 5 contains the circuit details for performing the time-interval measurement. The unknown capacitor must be discharged below the lower threshold voltage; Q3 and J-K flip-flop U6A; perform that discharge. R5 and R6 take care of the two known charging resistors for the two capacitance ranges and the output of U1, a LM78L05 voltage regulator, generates the fixed known voltage (E) needed for the measurement. By turning off Q3 via flip-flop U6A, the unknown input capacitor, C_X, is allowed to charge through either R5 or R6, depending on the level of the CM, Hi/Lo control line.

Since the charging current through R6 (1.2 MΩ) is low (<4.2 µA), any loading on the capacitance-charging circuit will cause errors in the final reading. By buffering the voltage across C_X with a BiFET op amp (U2, an LF351), the loading problem is avoided. R9 and R17 take care of the low-range thresholds; R16 is used to adjust the high-range thresholds. These threshold divider resistors are referenced to E, the fixed charging voltage.

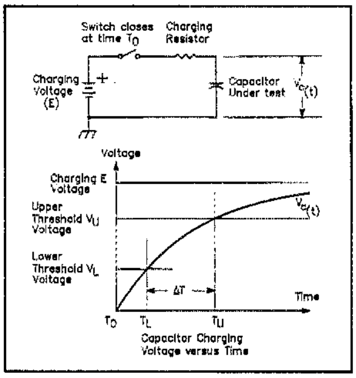


Fig 5—Capacitor charging curve.

The threshold voltages and the output of the buffer amp, U2, are fed to a dual voltage comparator chip, U3A and U3B (LM319). Both comparators are set up as Schmitt triggers and trip (change from the low to high state) when the output of the buffer amplifier exceeds their respective threshold levels. When the lower threshold is exceeded, U3B's output goes high turning on J-K flip-flop U6B (the time-gate generator). When the upper threshold is exceeded, U3A's output goes high and through inverter U7B, turns off U6B.

Flip-flop U6B's output gates the Z8671 CPU clock into the two cascaded 4-digit counters, almost the identical procedure followed for the inductance-meter period measurement. In this case, though, the time interval between two voltage thresholds (not the period of a square wave) is measured. The resulting numbers in the cascaded counters are BCD corrected and converted to ASCII, as before, and sent off to the PC for further processing.

Control/Addressing Decoding Logic

Before we can program and use the L-C meter, one more block of circuitry is needed. The lower right-hand portion of Fig 4 does the trick. First, how many I/O ports are used by the L-C meter? Five: four for the 82C54 counter chip and one for the sub-instruction-register, U14. U14 is an eight-bit latch that provides the L-C meter with the appropriate control signals needed by the L and C portions of the instrument. It obviously has to have a unique port name, or number, so the Z8671 CPU knows where to send the instruction byte. The same is true for the 82C54 counter. However, as mentioned earlier, the counter uses up four ports, one for setting up the operating modes of each counter and three for presetting and reading each 16-bit counter.

Where do these I/O ports come from? Remember the memory map of the Z8671 CPU board? The memory, 64 kbytes, is divided into eight groups of 8 kbytes each. Four of the groups are assigned to the four Z8671 CPU boards' memory chips; the remaining four are fed to the 50-pin I/O connector and labeled $\overline{CS2}$, $\overline{CS3}$, $\overline{CS4}$ and $\overline{CS5}$. When one of the CS lines is asserted it directs the Z8671 data bus (DB0-DB7) to any one of 8192 addresses (ports).

By decoding all or some of the address lines (A0-A12), we can select a single port, or a group of contiguous ports, from the 8192 designated by the activated CS line. In this particular address-decoding scheme, I chose to ignore the upper eight address lines, A8-A12. As a result of this choice, each CS line can select only 256 I/O ports instead of 8192 ports.

So how do we select the five ports we need for the L-C meter? S1 selects one of the CS lines, S2 selects one of the 32 possible groups of eight ports selected by the CS line. A2 divides the selected group of eight ports into two groups of four. Finally, A0 and A1 divide each group of four into four individual ports. Three of the second group of four ports, in this case, are superfluous; we only need one port to address U14. The selection of the CS line is simply a dip-switch setting.

The group of eight addresses is more complicated, using an 8-bit comparator chip (U15). Only five of the eight comparator lines on each input byte to U15 are used to perform the comparison operation: Remember, $2^5 = 32$, one of 32 possible groups of eight ports. When the two groups of five bits are equal, pin 19 of U15 is asserted low and is fed to a **NAND** gate wired as an inverter (U13A). The remainder of the logic, U7D, U13B, U13C and U13D, combine the group-enable signal (from U15) and address line A2 to form chip-select signals for U14 (instruction latch) and the counter chip. The counter chip has an internal 2-line to 4-line decoder that selects one of the four possible states of A0 and A1 to select the chip's registers.

The range of I/O ports that can be activated when $\overline{CS2}$ is selected by S1 while S2 is set to all zeros is shown in Table 1.

Now let's specify the actual addresses of the five ports needed to operate the L-C meter, assuming that S1 selects $\overline{CS2}$ and S2 is set to zero; see Table 2.

Since any address from 4004h to 4007h activates the instruction latch, let's arbitrarily assign address 4004h to the latch. Now that the port addresses are known and assigned, what do they do? Consider the easy one first, the instruction latch. Fig 4 provides us with the following information:

Instruction Register

Bit Number	*Function*
Q1: DB0 Select-CM	Selects capacitance meter function.
Q2: DB1 Start-CM	Starts capacitance-meter measurement.
Q3: DB2 CM,Hi/Lo	Selects high/low capacitance range.
Q4: DB3 Select-LM	Selects inductance-meter function.
Q5: DB4 Start-LM	Starts inductance-meter measurement.
Q6: DB5 Prime-Counter pulse	High to low to high.
Q7: DB6 and Q8: DB7	not used

Each of the control lines is self-explanatory, with the exception of Q6. Due to a few quirks in the structure of the 82C54 chip, a pulse has to be sent to the clock and gate inputs of counters connected in series, and this control line provides the required pulse under program control. To select the L meter, first Q4 of U14 is brought to a logic 1 and to start the measurement, Q5 of U14 is brought high.

Programing Considerations

A few more bits of information are needed before you're ready

Table 1
I/O Port Range Selection

Z8671 CPU Address Lines																Port
A15	*A14*	*A13*	*A12*	*A11*	*A10*	*A9*	*A8*	*A7*	*A6*	*A5*	*A4*	*A3*	*A2*	*A1*	*A0*	*Address*
0	1	0	0	0	0	0	0	0	0	0	0	0	0	0	0	4000h
S1 (/CSn)			These bits are					S2 selects					0	0	1	4001h
selects			ignored by the					these bits					0	1	0	4002h
these bits			decoding logic										0	1	1	4003h
													1	0	0	4004h
													1	0	1	4005h
													1	1	0	4006h
0	1	0	0	0	0	0	0	0	0	0	0	0	1	1	1	4007h

Table 2
Port Addresses Needed to Operate the L-C Meter

Port Address	*Port Function*
4000h	82C54, Counter zero
4001h	82C54, Counter one
4002h	82C54, Counter two
4003h	82C54, Control word register
4004h	Instruction latch U14
4005h	Instruction latch U14
4006h	Instruction latch U14
4007h	Instruction latch U14

to program the L-C meter. We have to program the 82C54 triple-counter chip. First, the three 16-bit counters are down counters. Starting at 0000d, following the first input pulse, they go to 9999d, then 9998d, and so on. Each counter can be independently programed to count in BCD or binary format, and each counter has six possible modes of operation. They can be preset to any number starting with 0, 1 or 2, depending on the operating mode, up to 10,000d or FFFFh, depending on the counting format. The contents of each counter can be read and sent off to the Z8671 MCU.

Let's talk about counter 0, the one used to accumulate 1,000 periods of the inductance-meter oscillator. This counter has to count down from 1,000 to 0. When it reaches zero, it provides a signal indicating that the measurement is complete. This type of operation falls into the "Interrupt on Terminal Count Mode," Mode 0. Here is the programing sequence: (1) Send a control word to the Control Word Register that specifies counter 0. (2) Send the counter preset value, 1000d, to counter 0's least-significant byte first, then the most-significant byte.

Here's what the BASIC/DEBUG instructions look like:

Line Number	*Instruction*	*Comments*
100	@%4003 = 31h	Sets mode for CT(0)
110	@%4000 = 00d	Sets ls byte of CT(0) to 00d
120	@%4000 = 10d	Sets ms byte of CT(0) to 10d

The same type of instruction sequence has to be sent to the Control Word Register for Counters 1 and 2. However, the operating mode for Counter 2 has to be changed to Rate Generator Mode: MODE 2. Counter 1 is set to the Terminal Count Mode, Mode 0, the same as counter 0. The instruction sequence looks like:

Line Number	*Instruction*	*Comments*
130	@%4003 = 71h	Sets mode for CT(1)
140	@%4001 = 99d	Sets ls byte of CT(1) to 99d
150	@%4001 = 99d	Sets ms byte of CT(1) to 99d
160	@%4003 = B5h	Sets mode for CT(2)
170	@%4002 = 99d	Sets ms byte of CT(2) to 99d
180	@%4002 = 99d	Sets ls byte of CT(2) to 99d

To send an instruction word to the L-C's Instruction Register (U14), use the following instruction format:

Line Number	*Instruction*	*Comments*
190	@%4004 = xxh	Sets bits of U14 to xxh (Range of xxh is 0 to FFh or 0 to 255d)

Armed with this information, we can start to develop a "verbal" description of the *two* programs that are needed to run the L-C meter. Why two programs? One for the PC and one for the Z8671 CPU. Details of the PC and Z8671 operating and calibration programs are included with the L-C meter parts kit.

Wrap Up

You've got enough to keep you busy for awhile. Our next installment, which will appear in an upcoming issue, is a frequency-synthesized, 1- to 10-MHz sine- and square-wave generator.

Notes

[1] Part 1 appears in the December 1993 issue of *QST.*

[2] PC boards and parts kits are available from the author. The PC motherboard is $30. Add $3 for shipping charges in the continental USA; elsewhere, add $5.

A PC-board template, drilling template and part-placement diagram are available for $10. Power-supply/motherboard parts kit containing the PC board and all components *less:* 3-wire ac line cord, PC-board mounting frame, power switch and power LED, $95. Shipping charges in the continental USA, add $5; elsewhere, add $10. Address all orders to: Ronald J. Portugal, 52 Susan Lane, North Haven, CT 06473. Payment: checks or postal money orders payable in US funds to Ronald J. Portugal.

[3] A part suppliers' list is presented on pages 35-38 of the 1994 *ARRL Handbook.*

[4] The following L-C meter project items are available from the author: (1) 1:1 positive PC-board artwork including the component-side silkscreen, component- and solder-side artwork, part-placement and drilling diagrams, $10; (2) double-sided, silkscreened, solder-masked PC board with plated-through holes, $20; (3) a complete kit including the PC board and part-placement diagram, all on-board components, ICs, IC sockets, connectors and hardware; price $75. On orders up to $50, please add $4; on orders over $50, please add $6. Foreign orders add an additional $3. Please make your checks or money orders payable in US funds to Ronald J. Portugal. Mail all orders to: Ronald J. Portugal, 52 Susan Lane, North Haven, CT 06473, tel 203-239-0942. Allow 4 to 6 weeks after receipt of order for delivery.

[5] For full details of the operation of the 82C54 triple counter, see the device manufacturer's specification sheets.

Strays

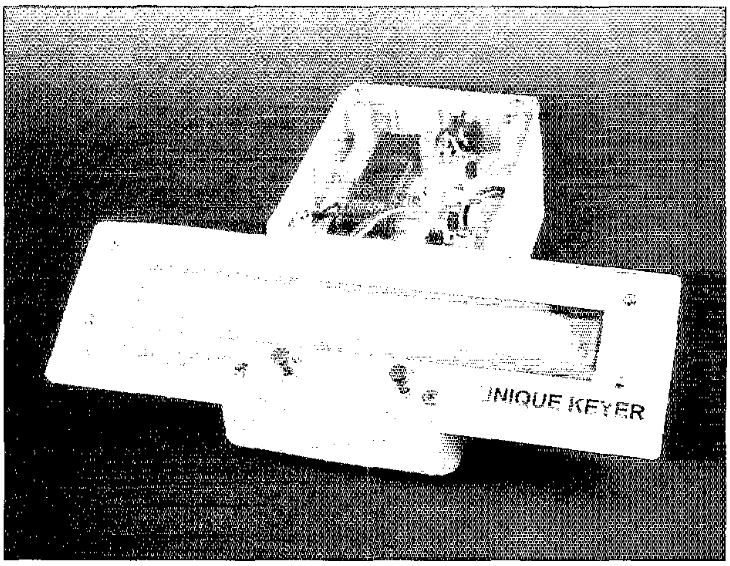
CRANK 'EM DOWN

◊ It pays to crank 'em down: During a winter Nor'easter, a single heavy gust of wind brought down the top half of this crank-up tower. Mike DiPersio, KC2Q, of Bradley Beach, New Jersey, hadn't fully lowered his tower, and his home-brew boomless quad and VHF antennas were virtually destroyed, after nine years of faithful service. *(KC2Q photo)*

Uncle Albert's Unique Keyer

This feature-packed keyer project can help you improve your CW sending and receiving whether you use a straight key, bug or keyer.

By Sam Ulbing, N4UAU
5200 NW 43rd St, Suite 102-177
Gainesville, FL 32606



(photos by Kirk Kleinschmidt, NT0Z)

Many code-practice aids await you if you just want to learn to copy Morse code to pass a test. Audio tapes, computer programs, subliminal approaches—you name it, and it probably exists. Once you've learned to copy enough CW to pass the test, though, you're largely on our own—you just grab your key and pound away. With enough practice, you can send pretty good code—or can you?

I recall that when I was a Novice, other operators kept begging off with "SRI QRN" and bidding me a hasty 73—even when the copy was good on my end. Did the other guy really have QRN or was my fist just hard to copy? After I joined a CW net and heard the wide variety of fists represented there, I *really* started to wonder how my sending sounded. Then, after taking a community college course in microcontrollers, I found that I knew enough to design a circuit that could send code and tell me how well I was sending. The result is Uncle Albert's Unique Keyer.[1]

What It Can Do

An 87C51 microcontroller and a two-row, dot-matrix liquid-crystal display (LCD) serve as the basis for the keyer. I wrote a program for the 87C51 that allows it to function as:

- An 8- to 40-WPM iambic keyer with one 47-character memory,
- A code-speed calculator,
- A CW reader that displays what you send, and
- A random-character code-practice generator that displays what it sends.

Code Reader

To me, the most unique feature of Uncle Albert's Unique Keyer is its ability to copy and display the code you send, whether you do it with the Keyer's on-board iambic keyer or an external keying device (keyer, bug or straight key). The CW-reader subroutine samples the code you send, translates it into characters by means of a look-up table, and displays it on the LCD. Then you can see what your code sounds like to the operators you work. You can see right away if you're running letters together or sending characters incorrectly—quite an eye-opening experience if you think your code is already perfect! (It helps my sending another way: I'm a poor speller, but I can do a better job if I can see the words as I spell them.) The letters scroll across the top line of the keyer's display, from right to left. The keyer displays common Amateur Radio procedural signals (*prosigns*) as shown in Table 1.

Table 1
How Uncle Albert's Unique Keyer Displays Procedural Signals

Meaning	*Character*	*Display*
End of Work ($\overline{\text{SK}}$)	···–·–	#
End of Message ($\overline{\text{AR}}$)	·–·–·	+
Double Dash ($\overline{\text{BT}}$)	–···–	=
Wait ($\overline{\text{AS}}$)	·–···	>
Go Only ($\overline{\text{KN}}$)	–·––·	(
Uncopiable Character		&

Speed Calculator

In calculating your sending speed, Uncle Albert's Unique Keyer keeps track of the lengths of your dots and dashes, and how many spaces you insert between characters and words. (The spaces are important, because extra space between code characters sent rapidly reduces your average sending speed.) The keyer calculates your speed every 40 characters and displays it on the LCD's lower line.

Iambic Memory Keyer

The iambic-keyer subroutine lets you set the element speed with your paddles and displays the speed as you adjust it. The keyer remembers the speed you've set, so when you later power the keyer back up or change modes by pressing **RESET**, the keyer comes up ready to run at that speed.

You can load the keyer's programmable memory via your paddles or external key,

---

[1] The keyer uses an 87C51 microcontroller chip, which must be programmed before use. *The 87C51 source code for the Uncle Albert's Unique Keyer program is unavailable.* A basic package, which includes a predrilled PC board, programmed 87C51, construction information and operating instructions, is available from the author for $38. A convenience package, which includes a predrilled PC board, programmed 87C51 microcontroller, all parts (except an enclosure, jacks, mounting hardware and power source) and construction and operating instructions, is available for $60 with a 16 × 2 display or $67 with a 40 × 2 display. Prices subject to change; Florida residents, add 6% sales tax. Order from Sam Ulbing, N4UAU, 5200 NW 43rd ST, Suite 102-177, Gainesville, FL 32606.

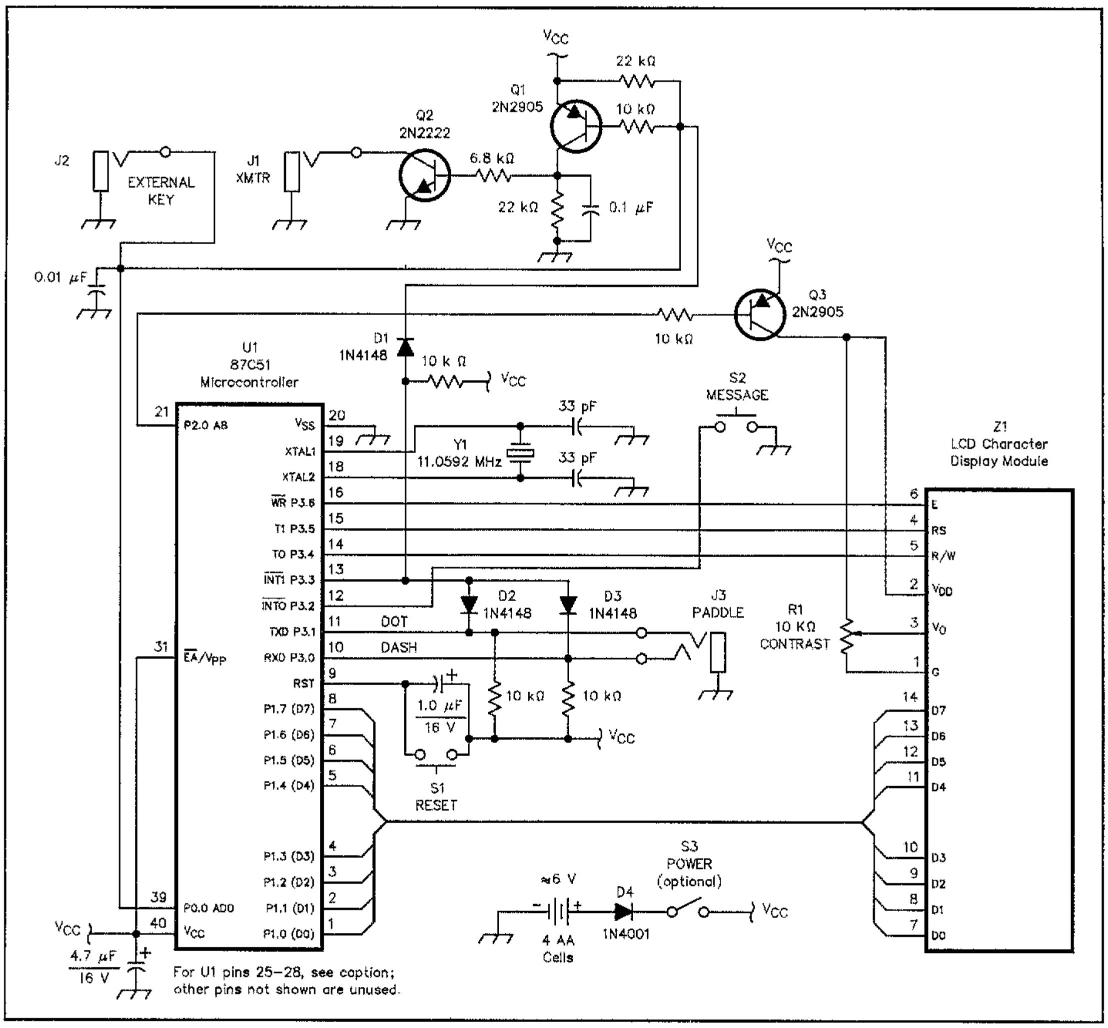


Fig 1—An 87C51 microcontroller and dot-matrix LCD character-display module (Z1) contain most of the Uncle Albert's Unique Keyer circuitry. The 87C51 must be programmed before use and is available, along with most of the other keyer parts, from the author as described in Note 1. The polarized capacitors in this circuit are electrolytics; the other fixed-value capacitors are disc ceramics. Aside from R1, which is a PC-board trimmer, the circuit's resistors are ¼-watt, carbon-film units. The text and Table 2 suggest modules suitable for Z1. Use the jacks appropriate for your station setup for J1, J2 and J3. IC pins not shown are unused, except for pins 25-28, connections to which are display-dependent, as described in the construction information available per Note 1.

and send the message at the keyer's set speed by pushing the **MESSAGE** button. The memory remains intact as long as you don't turn off the power. You can rewrite the message whenever you want to.

For transmitter and antenna testing, Uncle Albert's Unique keyer lets you use your dash paddle as a straight key. (The keyer can't display or calculate the speed of any code you send in this mode, however.)

Random-Code Generator

The random-code-generator subroutine can send as slowly as 5 WPM and features Farnsworth spacing below 18 WPM. That is, for keyer speeds below 18 WPM, the random-code generator sends characters at 18 WPM and spaces them to achieve the set sending speed. You can tell the program to send code in random five-character (*cipher*) groups or just a character at a time. In the cipher-group mode, the program sends until the display fills and stops to let you check your copying accuracy. Pressing a paddle starts the next session.

In the single-character mode, the program stops after each character. Pressing the dot paddle sends the next character. If you want to listen to the same character again, press the dash paddle.

Uncle Albert's Unique Keyer accomplishes all this with just two push buttons and your dot and dash paddles. A power switch is unnecessary because the 87C51 goes into the Power Down mode after a period of nonuse, drawing only 5 µA from its battery in this condition. (If you're concerned about this Power Down drain, you can add a switch to entirely disconnect the keyer from its battery.) During normal operation, the keyer's current demand averages around 6 mA. Four AA batteries should last a long time in this service.

The Circuit

Fig 1 shows the Uncle Albert's Unique Keyer schematic. The 87C51 microcontroller's clock uses an 11.0592-MHz crystal (Y1), a standard computer part. The **PADDLE** inputs connect to the 87C51's pins 11 and 10. Q1 and Q2 buffer the keyer's output (pin 39) for connection to a transmitter (or a practice oscillator). The keyed device should have a negative ground. The **EXTERNAL KEY** jack connects directly to Q1's drive line. The keyer accepts commands from the first input (**EXTERNAL KEY** or **PADDLES**) you activate after pressing **RESET**.

All the capacitors except C1 suppress transients—that is, reduce the keyer's sensitivity to noise spikes and interference from transmitted RF. C1 is necessary to reset the computer at power up. It allows the 87C51's **RST** pin to remain at V_{CC} for at least 2 ms after the power is applied—a condition necessary for smooth start-up. The three 1N4148 diodes (D1, D2 and D3) allow key/paddle closures at the **EXTERNAL KEY** and **PADDLE** inputs to "wake up" the 87C51 when it's in the Power Down mode.

Eleven lines run from the 87C51 to the dot-matrix LCD module, Z1. Eight of these lines provide data; the other three control circuitry in the display. R1 controls the display's contrast. Q3 controls the power to the LCD. When the 87C51 enters the Power-Down mode, it turns off Q3 to interrupt the LCD's supply. S1, **RESET**, starts or restarts the 87C51. S2, **MESSAGE**, stops and starts the memory message. Both switches work in conjunction with the paddles or straight key to communicate instructions to the 87C51.

I power my keyer with four AA cells through a 1N4001 diode (D4). Although the diode protects the circuit from reverse polarity, its main purpose is to keep V_{CC} below the 87C51's absolute maximum supply voltage (6.5 V). Four fully charged alkaline batteries can provide nearly 6.4 V, a level I consider to be too close to the 87C51's absolute maximum. Using only three batteries and no diode would power the 87C51 at the lower edge of its supply range, a situation that might lead to glitches as battery voltage falls. So, the 1N4001 is a compromise: Its voltage drop (about 0.6 V) limits the 87C51's maximum supply to around 5.8 V when four fully charged alkalines are used.

Display Choices

To permit builders freedom in choosing a display, the keyer has been programmed to drive 16 × 2, 20 × 2, 24 × 2, 32 × 2 and 40 × 2 displays (a jumper wire sets the IC to operate with a particular display configuration). Uncle Albert's Unique Keyer will work with many different LCD modules—for instance, the Hitachi LM052L, a 16 × 2 display that's quite small. On the other hand, the Optrex PWB40218 40 × 2 display (used in the keyer shown in the photographs) can show 40 characters at a time, permitting you to see more of what you're sending. (Visitors who cannot copy CW by ear find it more impressive, too!) The more characters a module can display, and the larger the characters, the more it will cost. Supertwist LCD modules, which afford better contrast over wider viewing angles than standard types, are also more expensive than their standard counterparts.

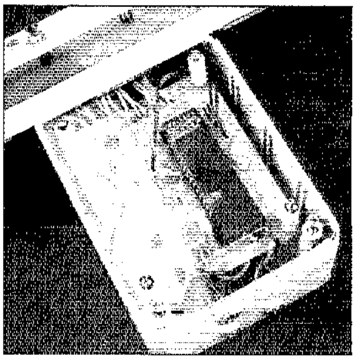
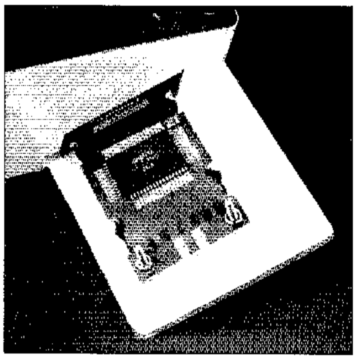

A PC board, available from the author, holds most of the components in this version of the keyer.

Table 2
A Common LCD-Module Pinout

Pin	*Function*
1	Ground (or V_{SS})
2	V_{CC} (or V_{DD})
3	V_O (Contrast)
4	RS (Register Select)
5	R/W (Read/Write Select)
6	E (Enable/Disable)
7	D0 or DB0 (Data Bit 0)
8	D1 or DB1 (Data Bit 1)
and so on to	
14	D7 or DB7 (Data Bit 7)

This listing shows the pinouts of LCDs carried by Digi-Key Electronics (Varitronix Limited character-display modules) and Mouser Electronics (Rohm and Epson character-display modules). Modules with this pinout and successfully used by the author include the Hitachi LM018L and LM052L, Philips LTN211-R, Optrex PWB40218, Varitronix MDL16265, an untraceable bought-used part marked *SLM22403BSA (242-HI-OAL)* and an Optrex PWB16230-CEM.


Surgery on a Radio Shack blue-plastic project box—here painted eggshell white—allows the box to hold a large LCD. A single formed piece of Lexan acts as the box top, display mask and keyer control panel.

Getting the Parts

Getting the parts for Uncle Albert's Unique Keyer is easy. The only Uncle Albert's Unique Keyer parts not available from Radio Shack are the 87C51 microcontroller, PC board, crystal and the LCD display. As detailed in Note 1, the programmed 87C51, PC board, LCD and other components are available from me if you wish to avoid shopping around. So far, I've built all of my keyers into plastic boxes and had no trouble with RF interference to the keyer. You may need to use a metal box and add bypass capacitors to your keyer's input and output lines if you run into difficulty with RF-induced keyer glitches.

If you decide to shop for an LCD module yourself, you'll find that many manufacturers make LCD modules. The modules I've seen appear to be identical in operation and pinout. (Of the six different displays I've used, all have worked well, although one [a Philips LTN211-R] needed a 1-µF capacitor across its power line to reduce the electrical noise it generates at start-up.) As Table 2 shows, the pinout of LCD character-display modules seems to be standard—at least for the modules available through two suppliers well-known to hams (Digi-Key and Mouser). Between these and other firms, and ham flea markets, you should have no problem buying a display for your keyer.

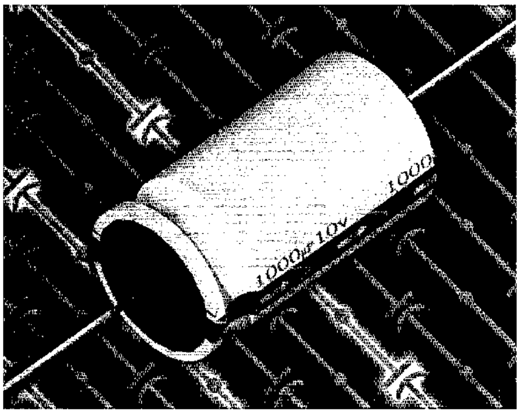
Construction

With so few parts, this project is quite simple to build. The LCD modules include LSI ICs that handle all the manipulations involved with driving the 5 × 7 dot matrix displays—all you do is hook them up and add a few other parts. Then all you need to do is power up the keyer and adjust the LCD "brightness" with R1—a setting that will depend on your viewing angle. Then you're ready to press the **RESET** button, set your keying speed and get on the air with Uncle Albert's Unique keyer! QST

SOUNDS LIKE OL' STEVE IS HOLDING HIS MICROPHONE UPSIDE DOWN, AGAIN

Under the Hood III: *Capacitors*

In this, the third in a series of articles aimed at providing a glimpse of the many components that go into our communications gear, we examine capacitors —how their construction and characteristics influence selection, use and circuit design.



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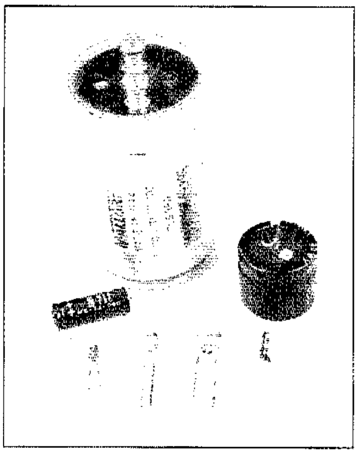
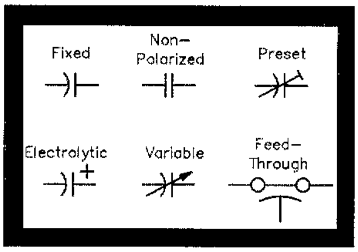
Capacitors are common in electrical and electronic equipment because what they do is so essential: They temporarily store energy electrostatically and release it in a variety of ways, depending on the applied voltage, frequency and associated circuitry. Capacitors resonate with inductors to form tuned circuits, filters and impedance-matching networks; interact with resistors to filter signals and set time constants; allow ac energy to pass while blocking dc; divide voltages and currents; and assist in turning ac into dc. Capacitors do their job across an enormous range of voltages, currents and frequencies—at levels ranging from that of radio impulses received from distant stars to the metal-melting might of interstate power grids.



Electrolytic capacitors span the large (a Mallory computer-grade part) to the tiny (a dipped tantalum, lower left). *(photos by Kirk Kleinschmidt, NT0Z)*



Like resistors, capacitors can be fixed or variable in value. In this Under the Hood, I'll cover some of the fixed and variable capacitor types used in Amateur Radio communication.

Capacitor Basics

In simplest form, a capacitor consists of two conductors, often generically called *plates*, separated by an insulator generically known as a *dielectric*. The capacitor stores energy in an electrostatic field between the plates. The plates' size and proximity, and the characteristics of the dielectric, determine the capacitor's *capacitance* —its ability to store energy—the basic unit of which is the *farad* (F), named after Michael Faraday. Less plate-to-plate spacing means more capacitance, as does greater mutual plate surface area. Connecting two or more capacitors in parallel increases the effective surface area and number of plates, and therefore directly sums their capacitance.

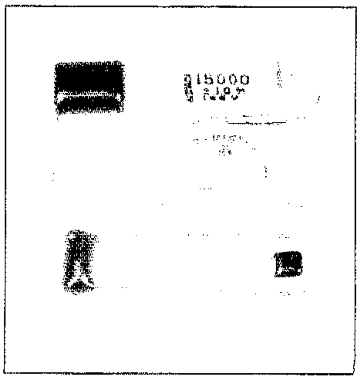
For a given plate spacing and size, a capacitor dielectric determines monolithic capacitance according to its *dielectric constant*—a number that corresponds to how much the dielectric increases capacitance compared to an equal thickness of air. For example, an air-dielectric capacitor that exhibits 100 pF at 60 MHz would increase in value to 260 pF if its dielectric were entirely replaced by polystyrene (dielectric constant = 2.6 at 60 MHz).

Capacitor Classification

Look at a parts list for almost any circuit that uses capacitors, and you'll see capacitors grouped according to their dielectric types and/or plate materials—air variable, ceramic, electrolytic, film, mica, oil, paper and more. Whatever they're made of, capacitors also vary widely in physical form, as the photos show.

Air

If you don't intentionally evacuate the space between a capacitor's plates, or fill it



Film capacitors: two polystyrenes (silver), a Mylar/paper (lower left) and two metallized-films.

with a gas, liquid or solid, that capacitor will be an air-dielectric type. Most air-dielectric capacitors are variable, although fixed air capacitors are sometimes used in high-power transmitting applications, and can be easily built by experimenters.

Ceramic

Ceramic capacitors have long been used for RF tuning and bypass applications, in part because of their low series inductance. Popular ceramic-capacitor types include disc, monolithic and multilayer. Disc-ceramic capacitors, easily recognized as the ubiquitous thin, tan, coin-shaped components common on many PC boards, are formed by two thin sheets of silver, nickel, or other metal that are separated by titanium dioxide or another ceramic dielectric. The resulting sandwich is either encased in a thin shell of ceramic or dipped in flame-retardant epoxy for protection against humidity.

Multilayer ceramic capacitors and surface-mount ceramic chip capacitors are formed by sandwiching layers of metallic-ink-on-ceramic electrodes. Monolithic ceramics have the highest capacitance per unit volume, in part because of their thin dielectric (on the order of 13 µm thick for a 16-V capacitor). Ceramic capacitors are available with voltage ratings of several kilovolts.

Electrolytic

Aluminum electrolytic capacitors are so named because (a) they are made of two aluminum conductors separated by an acid or salt electrolyte or (b) they use an electrolytically deposited dielectric film. In either case, the dielectric consists of a thin layer of solid aluminum oxide on the sheet destined to be the anode. Each foil sheet is electrochemically roughened for increased surface area which equates to greater capacitance. Aluminum electrolytic capacitors are common in power-supply filtering, interstage coupling and bypassing applications.

Because their aluminum-oxide dielectric insulates only against current flow in one direction, standard aluminum electrolytics are said to be *polarized*, and their leads are marked + and − accordingly. Although applying voltage with the wrong polarity can cause violent and permanent damage to an aluminum electrolytic, most have the ability to heal themselves following minor damage. Special designs, referred to as *computer-grade* and *high-ripple* electrolytics, are available for applications in which high alternating currents flow through the capacitor. *Bipolar* aluminum electrolytics are designed for applications in which polarity reversal is possible, or circuits in which short-duration ac voltages appear across the capacitor.

Tantalum electrolytic capacitors use a tantalum anode, a manganese-oxide dielectric and graphite-and-silver-paste cathode. The popular teardrop-shaped solid tantalum electrolytics are encapsulated in a flame-retardant epoxy coating that provides uniform lead spacing and protection against mechanical damage and moisture.

Like aluminum electrolytic capacitors, tantalum electrolytics are polarized and are designed to work in relatively low-voltage circuits. Although tantalum capacitors tend to be expensive, their compactness, low dc leakage and high capacitance have made them the most popular of surface-mount capacitors.

Film

Film capacitors use a plastic-film (commonly, polypropylene, polyester, polycarbonate, polystyrene, and polysulfone) dielectric. Film capacitors are commonly made by winding aluminum, tin, or other metal together with a plastic ribbon into cylindrical form. Alternatively, a metal film can be vacuum-deposited onto the plastic. Metallized polypropylene capacitors, useful in timing applications because of their stability and low leakage, are constructed in this way. Film capacitors are relatively inexpensive and provide capacitances up to a few microfarads at voltage ratings up to several hundred volts.

Mica

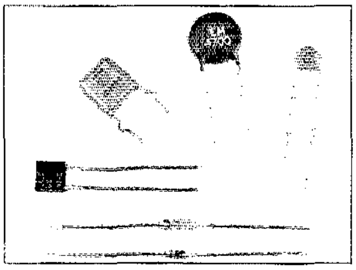
The popular *silver-mica* capacitors, used extensively in RF output circuits, are composed of a slab of mica sandwiched between two conducting plates. Mica's high dielectric strength and low dielectric loss makes such capacitors suitable for high-voltage and RF applications. *Postage-stamp* micas, so named because of the size and shape of their plastic packages, were common in RF tuning, coupling and bypassing applications before the appearance of disc ceramics.

More Types Available, More on the Way

The capacitor types I've mentioned so far are those most common in Amateur Radio equipment. Given the ongoing evolution of capacitor technology, it's perhaps best to think of those types only as rough guides of what a given capacitor type can do. Hybrids of, and expansions on, the major capacitor types are common, as new materials and new ways of applying established materials appear. For instance, Fig 1 shows how four capacitor types have been adapted for surface mounting.

Other capacitor types you may hear about include

- *Oil-dielectric capacitors*, still used in high-voltage power supplies to filter dc;
- *Paper-dielectric capacitors*, usually based on oil-impregnated paper, were common in audio coupling and RF/IF-amplifier-bypassing applications in pre-World-War II ham and broadcast receivers, but are now used mainly in dc and line-frequency applications because of their relatively high inductance;
- *Memory backup capacitors*, available in values up to *several farads* as alterna-



Ceramic capacitors: glass-encapsulated (the cylindrical pair), two discs and two monolithics. Ceramics are also available in axial packages that resemble film resistors.

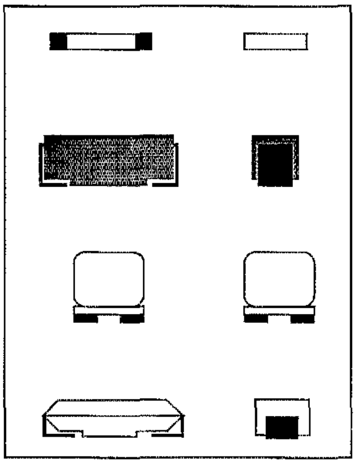
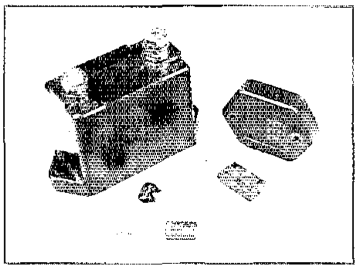


Fig 1—Virtually all capacitor types are available in surface-mount configurations. The surface-mount capacitor types shown here are, from top to bottom, ceramic chip, tantalum, aluminum electrolytic, and metallized-polyester film. Typical package dimensions for ceramic chip and tantalum-electrolytic surface mount capacitors are on the order of 2 or 3 mm on end.



Mica capacitors: Two large transmitting units (one of which carries a rating for current as well as voltage), two "postage-stamps" and a dipped silver-mica (directly below the *6032*).

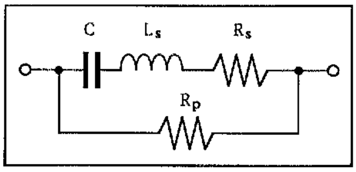


Fig 2—The equivalent circuit of a real capacitor reveals the presence of parasitic series inductance (L_S); series resistance due to wire leads, contact terminations, and electrodes (R_S); parallel resistance due to the finite resistivity of the dielectric and case materials (R_P); and the capacitance of the part (C). R_P is high in all but electrolytics. Dissipation due to R_S determines the maximum current that a capacitor can safely handle. At the frequency where L_S and C resonate, a capacitor presents a pure resistance; above this frequency, it behaves more like an inductor. A tantalum electrolytic of a few microfarads has a resonant frequency somewhere near 100 kHz.

tives to memory-backup batteries;

- *Feedthrough capacitors*, used to convey power or control lines through bulkheads, panels and chassis while bypassing the lines to ground for RF and noise;
- *Vacuum capacitors*, fixed and variable, in which a vacuum serves as the dielectric to provide high-voltage, low-loss performance in relatively small packages;
- *Doorknob* capacitors—large, cylindrical ceramics used in high voltage/current coupling, filtering and tuning applications.

A good long-term tactic in understanding, using and replacing capacitors is to learn to appreciate the significance of the functional characteristics used to describe all capacitors. Not only will your junkbox become more valuable to you, but you might be able to save a few dollars at your local parts store by substituting a cheaper, but equally capable, capacitor design for another.

Capacitor Characteristics

Aside from their physical properties, capacitors are functionally characterized in terms of their electrostatic capacity, dielectric loss, rated voltage, insulation resistance, operating frequency range, stability, reliability, and precision. Some of these characteristics can be better understood by referring to the functional equivalent of a capacitor (Fig 2). There is considerable overlap in the functional characteristics of most capacitor types.

Capacitance

The basic unit of capacitance is the farad (F), but a farad is so large compared to the capacitances common in radio and electronics that smaller and more practical microfarad ($\mu F = 10^{-6}$ F) and picofarad ($pF = 10^{-12}$ F) values are standard. Temperature, applied voltage and (for ac) frequency all affect a capacitor's actual value to varying degree, and its detailed specifications may reflect this. Like resistors, fixed capacitors are manufactured in standard values, for example, 1 µF, 2.2 µF, 4.7 µF, and 6.8 µF.

Tolerance

Depending on manufacturing tolerances, temperature, applied voltage and other factors, a capacitor's actual capacitance can vary greatly from its marked value. For the most part, tolerance is limited by the capacitor type. For example, disc-ceramic and metallized-polycarbonate-film capacitors are routinely available with 5% tolerances; mica capacitors, to within 0.5% (1% and 2% are typical). Electrolytic capacitors and general-purpose ceramics often carry wide tolerances—–10% to +80%, for example. One possible variation is the label *GMV* (guaranteed minimum value)—a specification, equal to a range of –0 to +100%, useful in bypassing and decoupling applications.

Rated Voltage

The rated voltage is the highest voltage that can be continually applied to a capacitor within a specified operating temperature range. Capacitors may be rated for ac, dc (usually as *working voltage dc* [WVDC]), or both. Contrary to ham lore, capacitors not rated for ac-line use—high-voltage dc types, for instance—should *not* be used in across-the-line or line-to-ground service because ac subjects capacitors to internal stresses that do not occur in dc service.

Equivalent Series Resistance (ESR)

Often specified for electrolytic capacitors, ESR sums a capacitor's ac and resistive losses as the value of a single equivalent resistor.

Dissipation Factor

Often specified for ceramic, film and mica capacitors, dissipation factor reflects ac loss as the ratio (expressed as a percentage) of ESR to the capacitor's reactance ($X_c = 1 \div 2\pi fC$). Examples of typical dissipation factors include 0.1% at 1 kHz for a polypropylene capacitor and 1% at 1 kHz for a metallized-film capacitor. Dissipation factor is sometimes expressed as its inverse, *Q* (quality factor), especially for capacitors specified for use in tuned circuits. The lower this number, the lower the capacitor's ac loss.

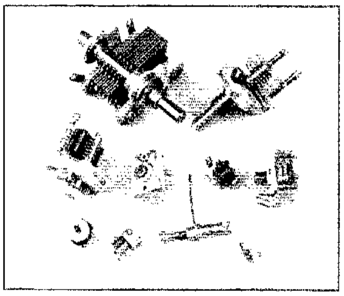
Tangent of Loss Angle (tanδ)

Commonly specified for electrolytic capacitors, tanδ, another measure of a capacitor's energy loss, is a capacitor's power loss divided by the reactive power

Variable Capacitors

Making a capacitor variable involves moving parts—moving the plates connected to one capacitor electrode relative to the plates connected to the other capacitor electrode. Historically, this motion has generally been achieved with rotating shafts, with one set of plates (the *rotor*, attached the shaft), moving, and the other set of plates (the *stator*) fixed, with air, ceramic, oil, plastic or a vacuum serving as the dielectric. The plates may be parallel planes that mesh laterally, or concentric tubes that mesh axially. Whatever the design, the more fully the plates mesh, the greater the capacitance.

Air-dielectric variable capacitors—"air variables"—were once common in receiving and transmitting applications. With the standardization of RF interconnections at 50 ohms and the introduction of digital frequency synthesis and broad-band, solid-state RF amplifiers, the need for air variables has significantly decreased. Even so, many types of mechanically variable capacitors appear in up-to-date RF designs. You'll find modern variable capacitors tuning transmitter-output, receiver-input and RF-matching and phase-shift networks, resonating filters, and adjusting oscillator



circuits to frequency. High-voltage air and vacuum variables continue to find use in RF matching networks—*antenna tuners*—and medium- and high-power RF amplifier circuitry, especially where vacuum tubes are used. Nonsynthesized broadcast receivers, and many ham and experimenter projects, continue to use low-voltage variable capacitors in many forms.

Arguably, though, *trimmer* capacitors—variables intended for set-and-forget service—are the most common variable capacitors today. Available in panel-mount, through-hole-PC-board and surface-mount form, they may use air, ceramic, film or mica dielectrics.—*NU1N* present when the capacitor is subjected to a sinusoidal voltage at a specific frequency. The lower this number, the lower the capacitor's ac loss.

Frequency Ratings

Although the useful frequency ranges of various capacitor types considerably overlap, and capacitor packaging becomes increasingly important as frequency rises, you may find a few wide generalizations useful:

- Electrolytics are generally used at audio frequencies down to dc.
- Depending on packaging and value, ceramic capacitors are used from dc to the low VHF range (parts with wire leads) or into the UHF range and beyond (surface-mount and chip capacitors).
- Mica capacitors (*not* postage-stamp types, though) are used from dc up into the lower VHF range.
- Depending on packaging and construction, film capacitors are useful from dc to the gigahertz range.

Operating Temperature

Capacitors are generally rated at a specific operating temperature. In addition, an operating-temperature *range*, typically -40 °C to +85 °C for electrolytics and -55 °C to +125 °C for ceramics and silver micas, may be specified. As is generally true of electronic components of all types, higher operating temperatures generally mean shorter capacitor life.

Expected Lifetime

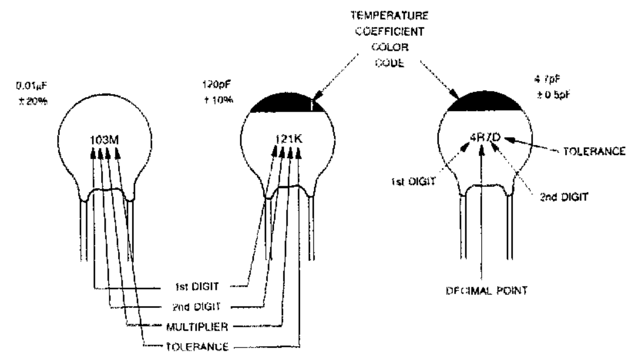
The service life of capacitors is generally specified in terms of temperature and voltage, for example, 100,000 hours at 85 °C and 16 V. Lower temperatures and working voltages can significantly lengthen a capacitor's expected lifetime.

Leakage Current

The executive summary is that capacitors "block dc," but the long answer is that some capacitors, especially electrolytics, leak *some* dc. This characteristic is particularly important in energy-storage and timing applications, and in battery-powered equipment. Capacitor specifications sometimes list a related characteristic, *insulation resistance*.

Temperature Coefficient

Every capacitor's capacitance varies somewhat with temperature. The change can be positive (increasing temperature translates to more capacitance) or negative (increasing temperature translates to less capacitance). A capacitor's temperature coefficient expresses how much (usually in parts per million) and in which direction (+ or –) its value changes per unit of temperature change (usually in terms of degrees Celsius). Capacitor temperature coefficient is particular important in oscillator and timing circuits, as Wes Hayward described in December 1993 *QST*'s "Measuring and Compensating Oscillator Frequency Drift." Capacitors are available in a wide range of temperature coefficients, including temperature-stable types that change value very little across wide temperature swings.



MULTIPLIER

Number	*Multiply by*
0	NONE
1	10
2	100
3	1000
4	10 000

Tolerance

Letter	*≤ 10 pF*	*> 10 pF*
B	±0.1 pF	—
C	±0.25 pF	—
D	±0.5 pF	—
E		±.5%
F	±1 pF	±1%
G	—	±2%
H	—	±2.5%
J	—	±5%
K	—	±10%
M	—	±20%
P	—	-0 +100%
S	—	-20 +50%
W	—	-0 +200%
X	—	-20 +40%
Z	—	-20 +80%

Color	*Temperature Coefficient*
Black	NP0
Brown	N030/N033
Red	N075/N080
Orange	N 150
Yellow	N 220
Green	N 330
Blue	N 470
Violet	N 750
White	P 100
Red and violet	P 100

Fig 3—The disc-ceramic capacitors in your radio likely carry value markings applied according to this or a similar system. Plastic-film and tantalum electrolytic capacitors are often similarly marked, but without the temperature-coefficient indicator. Aluminum electrolytics, and some tantalums, often carry "plain text" markings indicating their value, operating-temperature range and polarity.

Power-Handling Capability

Power-handling capability, usually stated in volt-amperes, is rarely an issue for small capacitors intended for signal and bypassing/decoupling use, but may be a critical in components that operate in high-power circuits. Capacitors specifically intended for use in high-RF-current service may be labeled to indicate their current-handling ability versus frequency. Capacitors that mysteriously fail in high-power RF circuits—in transmission-line filters or when used as dc-blocking/output-coupling capacitors in vacuum-tube amplifiers, for instance—may do so because of current-dependent overheating even when operated well within their *dc* ratings.

Capacitor Labeling

A capacitor's capacitance, voltage rating, tolerance, temperature characteristics and (if applicable) polarity can be indicated in a variety of ways. Surface-mount capacitors use an alphanumeric code printed on the capacitor body. Disc ceramics are usually marked with alphanumeric indicators of capacitance, voltage rating, and temperature characteristics. (Fig 3 shows one system you're likely to encounter in some form.) Alphanumeric or color labels may convey the characteristics of many smaller tantalums, ceramics, and micas. A dark strip or band may mark the negative electrode of aluminum electrolytics, the positive electrode of tantalum electrolytics, and the outer-foil lead of film capacitors.

Summary

Now that you've glimpsed the world of capacitors, a peek inside your radio can show how much their contribute to a radio's interior look: Their vibrant colors, sleek coatings and great variability in packaging bring even the most conservative PC board layout to life.

So ends Under the Hood's coverage of capacitors—the parts that store energy electrostatically. Next in Under the Hood, we'll look at components that store energy *magnetically*. See you then.

Mother Nature's Radio

Despite our modern understanding of radio phenomena, the study of the Earth's natural VLF radio emissions is hardly out of its infancy. The best part is—you can get involved!

By David Schneider, AD4CC
Northern Kentucky University
Dept of Physics and Geology
Highland Heights, KY 41076

If Mother Nature had made our ears capable of hearing electromagnetic radiation instead of audible sounds—in our normal auditory hearing range of 20 Hz-20 kHz—*what could we hear?* As *it turns* out, a lot! We'd hear "chorus," "hiss," "tweeks," "whistlers" and other exotic sounds. See the sidebar, "The Sounds of Natural VLF Radio."

What are these sounds and where do they come from? They're natural sounds *generated in the audio-frequency range of* the electromagnetic spectrum. The audio-frequency range of the spectrum covers the same range of frequencies as the audio-frequency (sound) portion, but electromagnetic waves travel at the speed of light, not at the speed of sound. In other words, our ears can hear a 10-kHz audio sound, but we can't hear a 10-kHz electromagnetic (radio) wave unless it's electronically converted to an audible sound (hence the need *for radio receivers!*).

These sounds are associated with disturbances in the Earth's atmosphere. They're initiated by lightning and by the charging of the atmosphere by the Sun.[1] The frequencies that produce these unusual sounds reside, for the most part, between a *few hertz and several hundred kilohertz.*

This broad range of frequencies is broken down into three main segments (see Table 1). Most of the naturally occurring radio effects from lightning have a maximum frequency of about 5 kHz, so these sounds usually fall into the Very Low Frequency (VLF) part of the electromagnetic spectrum.

HF-and-higher frequencies refract through the ionosphere, allowing communication over great distances. The lower ionosphere, the D layer, is about 80 km above the surface of the Earth. This means that about 4000 full waves at 20 meters (14 MHz) can fit between the Earth and the *lowest layer of the ionosphere.*

At VLF, however, only a few waves (or a fraction of a wave) can fit in this space. For example, a 5-kHz VLF signal has a wavelength of 60 km—it's barely able to fit in the D-layer waveguide! (The D-layer waveguide, or the Earth-ionosphere waveguide, are terms that describe the 80-km space between the surface of the Earth and the bottom of the D-layer.)

Armed with this knowledge, you'd probably think that naturally occurring VLF radio waves travel only short distances, at least by conventional propagation.

Well, by more conventional propaga-

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[1] Notes appear on page 51.

Table 1
Wavelength and Frequency

Description	*Abbreviation*	*Frequency*	*Wavelength*
Extremely Low Frequency	ELF	3 Hz-3 kHz	100,000-100 km
Very Low Frequency	VLF	3-30 kHz	100-10 km
Low Frequency	LF	30-300 kHz	10-1 km

The Sounds of Natural VLF Radio

As the Earth races on its complicated path through the heavens, powerful and mysterious planetary forces work their magic. The result: a cacophony of "natural radio" sounds:

☐ *Chorus:* Chorus sounds like a flock of chirping birds! It occurs most frequently in the morning hours, hence its nickname "dawn chorus." Increasing tones between 1 and 5 kHz seem to be most common. Chorus is usually accompanied by other VLF phenomenon such as hiss or whistlers.

☐ *Hiss:* Hiss sounds just like its name. A continuous band of frequencies denotes VLF hiss. Sonograms of hiss signals have shown cutoff frequencies ranging from 2 to 30 kHz. Hiss has also been associated with enhancements in auroral activity.

☐ *Tweeks:* Tweeks are believed to result from "spherics" that echo back and forth in the Earth-ionosphere waveguide. They usually sound pure in tone, much like a note from a musical instrument. The tone sometimes resembles a "ping" sound.

☐ *Whistlers:* Whistlers tend to be most common at night or just before dawn. They are also more frequent at mid-latitudes, peaking between 40-55 degrees geomagnetic latitude. See the text for more details.

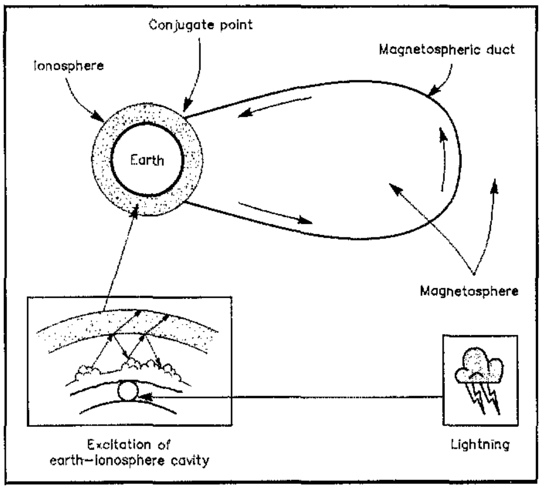


Fig 1—Schematic showing the generation and propagation of ducted whistlers.

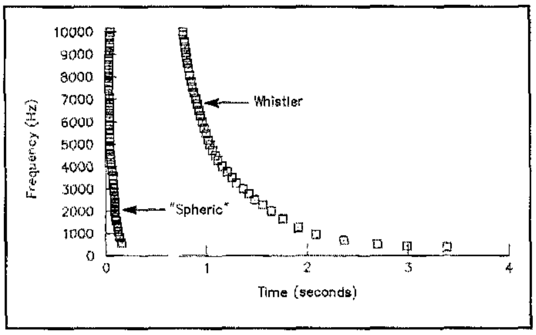


Fig 2—A lightning "spheric" followed by a "ducted" whistler.

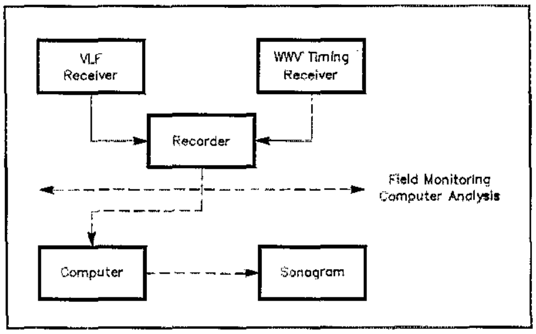


Fig 3—A typical whistler recording and analysis setup.

tion modes, VLF signals *do* travel relatively short distances, but even early experimenters proved that VLF waves can traverse large distances.[1] In fact, VLF waves can travel from one hemisphere to the other and back again!

Although HF radio waves sometimes travel thousands of miles from transmitting antenna to receiving antenna, VLF radio waves can travel more than 100,000 miles from a lightning stroke to a receiving antenna. The mechanism that propagates these VLF waves isn't the more-familiar ionosphere, it's the Earth's *magnetosphere*.

Planetary magnetospheres resemble magnetic dipole fields, but their currents flow in the plasmas that extend from the outer ionosphere to thousands of miles in space. These currents, resembling the familiar horseshoe-shaped magnetic flux lines, result in part from solar winds that reach the Earth and from the rotation of the Earth about its axis.

A wealth of information can be gained from the study of natural VLF waves. Propagation information, electron densities in the ionosphere and magnetosphere and the location of the lightning strikes that cause particular VLF waves are a few of the studies that are under way.

Man-made VLF signals are exploited by several groups, especially the US Navy. The Omega signals broadcast at various places around the Earth are used as maritime and aviation navigation aids. Man-made VLF generally falls in the 10-40 kHz range. These predictable signals can be used in conjunction with naturally occurring VLF to help us better understand their propagation characteristics.

Whistler VLF Theory

Whistlers are interesting and informative natural VLF radio waves (see Fig 1). A whistler is a VLF wave that originates with a lightning strike. Generally, there are at least two wave events associated with the lightning pulse. The first event is a short-lived wave that propagates from the lightning strike through the Earth-ionosphere waveguide to the receiver. This event is called the *atmospheric*, or simply *spheric*.

The second event is the whistler itself. The whistler wave travels up to and through the ionosphere, enters the magnetosphere and follows the Earth's magnetic field through a relatively narrow magnetic "duct" to the opposite hemisphere. After arriving in the opposite hemisphere, it again traverses the ionosphere, exits and propagates in the Earth-ionosphere waveguide to the receiver.

This is called a "one-hop" whistler; you can see the reason for its delay time relative to the spheric. A "two-hop" whistler is one in which the hemisphere where the lightning stroke originated also receives the whistler wave after it's ducted twice in the magnetosphere.

Under the right atmospheric conditions, this "hopping" between hemispheres can happen many times, sometimes several hundred! The number of hops determines the delay between the lightning strike that produces the original spheric and the whistler event.

Ducts that direct the whistler wave to the opposite hemisphere end up at a conjugate point on the Earth's magnetic field. A whistler that originates in the Northern Hemisphere will be ducted to the Southern Hemisphere to a point in the Earth's magnetic field equal in magnitude to the magnetic field where it originated.

How to Get Involved in Whistler Research

Hams took part in receiving natural VLF signals across North America this past year, and you can participate in this monitoring program and directly contribute to scientific endeavors of great interest to NASA, the Navy and the scientific community.

Further analysis of whistlers and their sonograms is ongoing. Northern Kentucky University and VLF monitoring teams across North America will make additional recordings in March 1994.

Our goal is to expand our efforts to other parts of the globe and incorporate VLF recordings from as many different geomagnetic locations as possible. If you'd like to take part in this scientific team effort, please write me at the address shown at the beginning of this article. We'd like to hear from you!

The physical consequence of traveling thousands of miles through magnetic ducts is that higher frequencies arrive at the receiver before the lower frequencies (see Fig 2). The more hops the whistler takes, the greater the time between the higher- and lower-frequency arrivals (that is, the dispersion increases).

An accurate measurement of the dispersion can lead to an understanding of the whistler's propagation path, as well as the whistler's spectral shape. In addition to being fun to listen to, a deeper understanding of the composition of the ionosphere and the magnetosphere (ion density, electron density, electron temperature, and so on) can be realized.

Whistler VLF Experiment

One way to study whistler phenomena is to simultaneously record whistlers at sites hundreds or even thousands of miles apart. This technique can provide information on whistler propagation by directly comparing the spectra of the same whistler recorded simultaneously at distant sites.

For example, a one-hop whistler must traverse the ionosphere at least twice and enter at one place into the ionosphere from the magnetosphere. By comparing the spectra of the same whistler at receiving stations thousands of miles apart, we can look for changes in the wave form.

This may manifest itself in one spectra as a missing or attenuated segment that may be caused by absorption as the whistler wave travels from its exit point in the ionosphere to a distant receiver.

Noting which frequency components have been attenuated can lead to information about which part of the ionosphere—the D, E or F region—caused the effect. As you may guess, receiving a whistler that has propagated through the gray line—the line separating day from night—could give

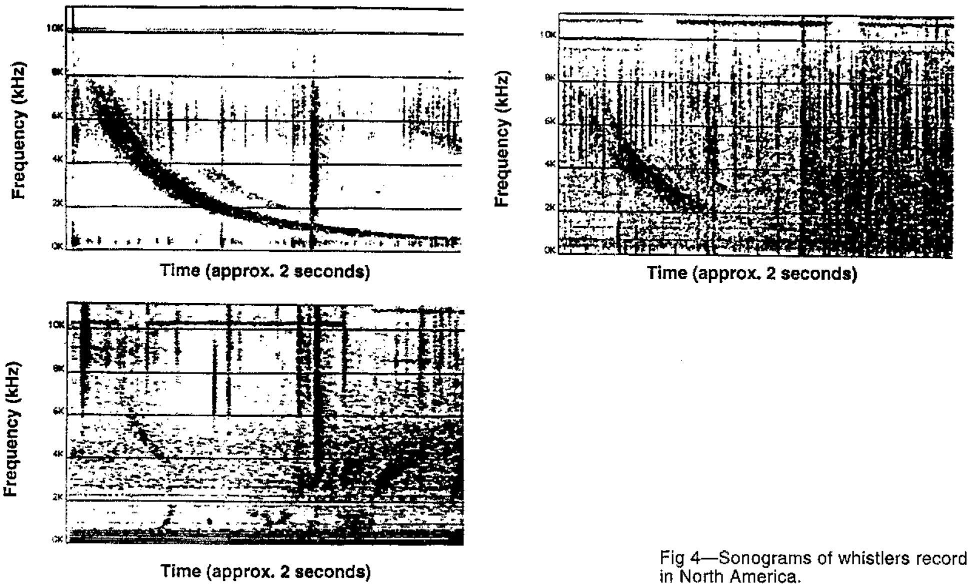


Fig 4—Sonograms of whistlers recorded in North America.

details about the density of ions and electrons in the D region, which is most pronounced on the daylight side of the Earth.

Fig 3 shows a schematic for a typical VLF monitoring and analysis station. The basic configuration consists of the VLF receiver and antenna, a timing device (usually WWV's 5, 10, 15 or 20-MHz time signal broadcast from Ft Collins, Colorado) and a stereo cassette or VHS recorder.

One channel of the stereo recorder will record WWV, and the other channel will record the natural VLF phenomenon. Using a single timing source makes it possible to accurately compare recordings made simultaneously at several locations.

VLF receivers can be purchased from Conversion Research for about $70 (complete with an antenna), or you can make one from inexpensive parts.[2] Computer analysis can then be accomplished with commercial software and hardware, such as *Sound Edit* and *Mac Recorder* on an Apple Macintosh computer. Display and analysis of sonograms can also be done with counterpart software on an IBM-compatible PC.

Whistler sonograms recorded March 22, 1993, are shown in Fig 4. These sonograms, recorded at several points in North America, are of the same whistler. The horizontal lines near 10 kHz are man-made Omega broadcasts. The vertical lines are spherics from local lightning discharges. The variation in cutoff frequencies for this whistler at various locations provides clues to its propagation characteristics in the Earth-ionosphere waveguide.

Acknowledgments

The success of this VLF program is possible because of dedicated participants at Northern Kentucky University and across the US. Contributors include Dr Mike McPherson, Dr Bill Wagner, Dan Spence, Toxanne Barnes, Steve Phelps and Justin Rains. Special thanks goes to Dr Dennis Gallagher of the Magnetospheric Physics Branch of NASA's Marshall Space Flight Center.

Funding for the VLF project is provided by the Kentucky Space Grant Consortium (KSGC) and NASA.

David Schneider has been licensed since his sophomore year in high school. His previous call signs were KA4LHP and N4RAM. He holds a BS in physics and mathematics from Northern Kentucky University, an MS in atomic physics from the University of Connecticut and a PhD in chemical physics from the University of Cincinnati.

He's assistant professor of physics at NKU and consults for the specialty chemicals industry. This past summer, he spent 10 weeks at NASA's Marshall Space Flight Center in Huntsville, Alabama, where he started a three-year project to study VLF.

David says his early involvement in Amateur Radio was a major influence in his choosing a career in the sciences.

Notes:

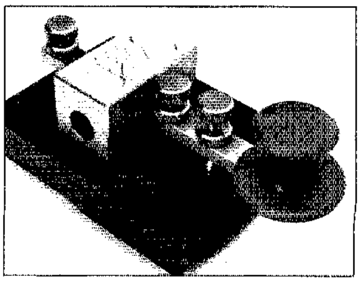
[1] See, for example, R. A. Helliwell, *Whistlers and Related Ionospheric Phenomena*, Stanford University Press, Stanford, California (1965).

[2] An excellent description of whistlers and other VLF phenomenon, plus instructions on how to build a VLF receiver, can be found in the July 1992 issue of *Science PROBE!* The article is titled "Listening to Nature's Radio," by Michael Mideke.

New Products

JONES STRAIGHT KEY

◊ A new hand key, produced by Peter Jones Engineering of England, shares many features with the Jones dual-paddle key, including a heavy steel base with a brilliant red finish, rotary ball bearings and tension spring enclosed in a solid machined-brass block, instrument-knurled heads on all adjustment screws, serial number engraved on the brass block and heavy duty construction throughout. The Jones hand key comes with a standard "Navy" knob. The pure copper electrical contacts under the base will key any transmitter, and a solid touch is provided by the key's 3$\frac{1}{4}$ pound weight. Retail price $118 plus $4 s/h. Palomar Engineers, PO Box 462222, Escondido, CA 92046; tel 619-747-3343, fax 619-747-3346.



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OUR COVER

This isn't exactly rush hour, but "Land Rovering" along mysterious jungle trails does have its inconveniences! This 'Rover is crossing a washed-out trail in Malaysia as part of the 1993 Camel Trophy, a 1000-mile, 16-team adventure expedition.

Richard Diamond, G4CVI, and his South Midlands Communications Team provided technical and communications support for the expedition—and managed to field a successful ham radio DXpedition to desirable 9M6-land. Read about their adventure in "Camel Trophy '93" on page 29. *(cover photos by Lee Farrant)*

CONTENTS

February 1994
Volume LXXVIII Number 2

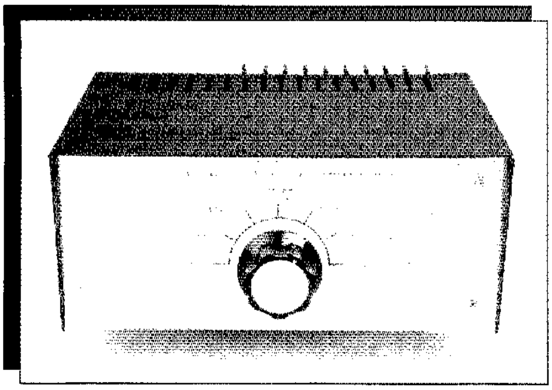
TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS



An Easy-to-Build 25-Watt MF/HF Amplifier

Do you need a medium-power linear amplifier for SSB or CW? Congratulations— you just found it!

By Gary Breed, K9AY
7318 S Birch St
Littleton, CO 80122

Here's a 25-W, 1.8- through 30-MHz, class-A linear power amplifier that's simplicity itself. What makes it simple is the use of a self-biased transistor module requiring few external components. To control harmonic output, a set of five-section low-pass filters is included. Power-supply requirements are +28 V at 2.5 A and –5 V at 200 mA.[1] With a gain of about 13 dB, a 1- to 1.4-W driving signal is all that's needed to deliver 25 W output. Gain is flat within ±0.75 dB across the covered frequency range.

If 25 W isn't enough for you, it's easy to directly apply the design information to build a 50-W amplifier—all you do is use a larger transistor module! Another step toward project simplicity is the availability of kits. Each kit contains all the major components for either a 25- or 50-W version.[2]

Amplifier Design

When designing a power amplifier, the first step is to select the right transistor(s). Excellent bipolar-junction transistors (BJTs) and field-effect transistors (FETs) are available from well-known companies such as Motorola, M/A-COM PHI, SGS-Thomson, Philips, Mitsubishi and others. A number of smaller companies also make power transistors, usually for more-specialized applications. MicroWave Technology, Polyfet RF Devices, and Directed Energy may be company names unfamiliar to you, but they all make power transistors for MF and HF applications.

In this amplifier, I use the SLAM-0111 from MicroWave Technology.[3] I didn't choose it because of its gain, its efficiency, or even its price; I selected it because it's *very easy to use*. The device consists of two power JFETs (the particular specialty of MicroWave Technology), operating in push-pull. Since JFETs behave similarly to triode vacuum tubes, the company dubbed them *Solid State Triodes*. SLAM (Solid-state-triode Linear Amplifier Module) devices include thick-film bias resistors in the package with the transistors. These resistors set the gate bias for class-A operation, and establish a 50-Ω input impedance. At the rated power and supply voltage, the push-pull output impedance is also 50 Ω!

With such convenient input and output impedances, matching the devices to a 50-Ω system merely requires 1:1 balun transformers at the input and output. Because the bias voltage is internally generated, the only other external circuitry required is a suitably bypassed and isolated 28-V power supply.

Circuit Description

The amplifier schematic is shown in Fig 1. The balun driving the gates of the push-pull transistors is a conventional transformer. The primary and secondary windings are each three turns of #28 wire, wound on a two-hole ferrite balun core of 73 material ($\mu_i = 2500$). These transformers are broadband enough to provide 1.8- to 30-MHz operation and offer dc isolation with no additional components. The input-transformer primary is center-tapped and bypassed to provide access to the gates for external dc bias (more on this later).

The output transformer is constructed in the same manner as the input transformer—it's just larger. Two ferrite beads of 77 material ($\mu_i = 2000$) make a two-hole core, with primary and secondary windings of three turns each, using #24 hookup wire. The primary (transistor side) is center-tapped and bypassed to provide dc voltage to the drains. Feeding dc through a center-tapped transformer eliminates the need for the usual bifilar RF choke seen in push-pull amplifiers—another reduction in the component count. Multiple bypass capacitor values (0.01, 0.1 and 10 µF) are used to cover the MF/HF range. That's the basic amplifier block: two transformers, a SLAM device, and a few bypass capacitors!

Class-A Operation Notes

By definition, transistors operating in class A conduct over the entire 360 degrees of the signal (that's all the time, of course). This operational mode assures that the transistor is always operating in the linear region of its input-to-output transfer characteristic. To do this, the device must be biased to handle the maximum signal at all times.

Obviously, this class of operation is pretty inefficient, since full current is drawn whenever the amplifier is on. A "perfect" transistor operating class A can only be 50-percent efficient, and real transistors do no better than about 40 percent. This amplifier draws 2.5 A from a 28-V power supply for an input power of 70 W. When it is providing 25 W, it's 36-percent efficient. (When there is no input, it's 0-percent efficient!)

To help reduce the heat generated by an amplifier that requires 70 W, a negative bias can be applied to the gates when not transmitting. A bias of –5 V results in a 0.25-A standby drain current instead of the full 2.5 A. The internal bias resistors are about 50 Ω on each gate, and dissipate a maximum of 1 W. Under these biasing conditions, the resistors each dissipate 0.5 W. *Don't* try to cut off the transistors completely with greater bias voltage! You'll risk burning out the resistors.

Some may ask, "If class A is this power hungry, why use it?" In a word: *linearity*. If you want excellent linearity (which means minimum distortion caused by harmonics or intermodulation), class A is *the* way to go. For example, all small-signal amplifiers for receivers and low-level transmitter stages operate class A because they must handle signals without distortion. However, they operate at very low power, so power dissipation is rarely an issue. This power amplifier further minimizes distortion by using push-pull operation, which cancels even-order distortion products in the output and makes the next part of the design easier than usual.

[1] Notes appear on page 34.

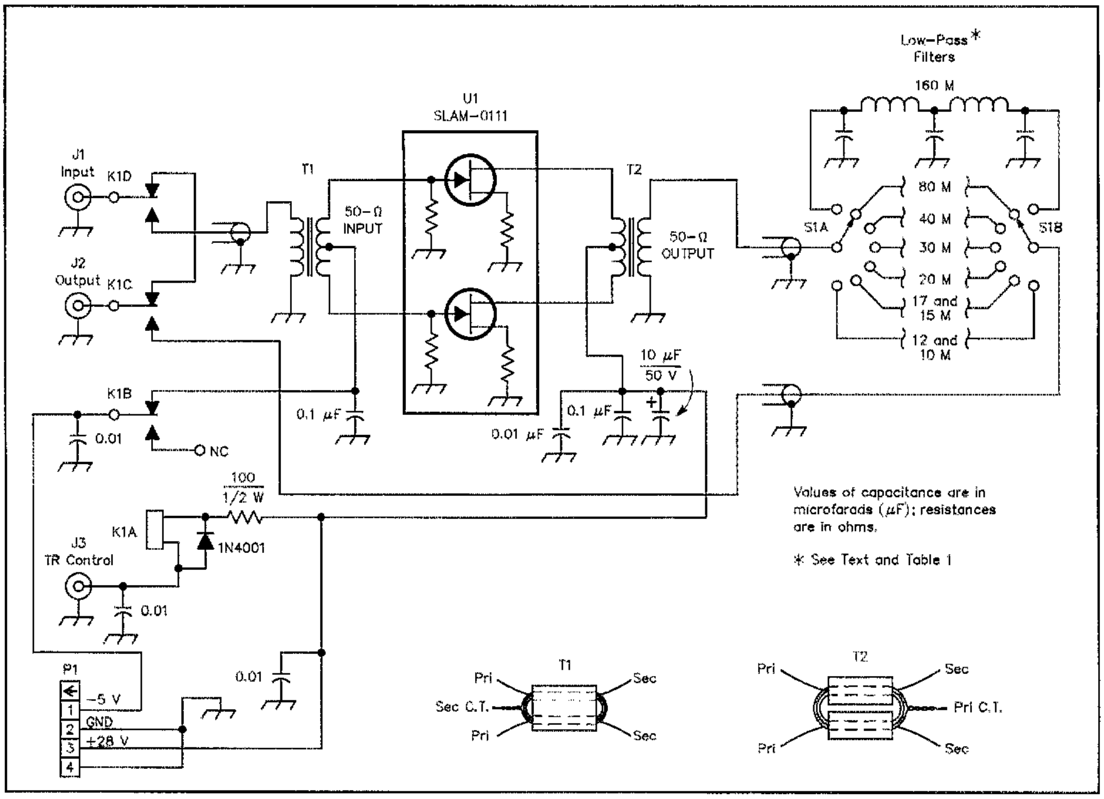


Fig 1—Schematic diagram of the 25-W class-A amplifier. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. Equivalent parts can be substituted.

J1—Panel-mount BNC socket.
J2—SO-239 connector.
J3—Phono jack.
P1—4-pin male Jones plug.
K1—3PDT relay with a 24-V dc coil. A surplus Potter & Brumfield KHP series 4-pole relay is shown in Fig 6; one pole is unused. (All Electronics catalog number 4PRLY-24N [$4] or Ocean State Electronics R12-17D3-24 [$10.90] are suitable. See the Part Suppliers List on pp 35-40 of *The 1994 ARRL Handbook* for addresses and telephone numbers.—*Ed.*)
S1—2-pole, 7-position ceramic rotary switch. My switch is made from two surplus CRL 11-position switch wafers and an indexing assembly providing selectable stops. The wafers are spaced about 1½ inches apart. CRL PA-200 series switch wafers and PA-300 series shaft and indexing assemblies are suitable (switches are available from Newark Electronics; tel 312-784-5100, fax 312-784-5100, ext 3107, to locate your nearest Newark distributor).
T1—Primary: 3 turns #28 AWG; secondary, 3 turns #28 AWG, center-tapped. Core: Fair-Rite #2873002402 balun (Amidon BN 73-2402).
T2—Primary, 3 turns #24, center-tapped; secondary, 3 turns #24. Core: two Fair-Rite #2677006301 beads (Amidon FB 77-6301).
U1—SLAM-0111 ultralinear 25-W, class-A, self-biased power FET module or SLAM-0122, 50-W version (Microwave Technology, 4268 Solar Way, Fremont, CA 94538, tel 510-651-6700, fax 510-651-2208).
Misc: RG-174 coax, enclosure (3½ × 7⅞ × 5¼ inches [HWD]), heat sink (3 × 4⅛ × 1⅜ inches [HWD]), PC-board material, knob, mounting hardware.

Harmonic Filter Design

As mentioned previously, the amplifier uses several low-pass filters to cover the nine MF/HF amateur bands. Each filter was initially designed for a cutoff frequency 20 percent higher than the upper end of their respective 160, 80, 40, 30, 20, 15 and 10-meter ham bands. The 15-meter filter is also used for 17 meters, and the 10-meter filter for 12 meters.

With no filtering, even-order harmonics (2nd, 4th, etc) are more than 40 dB below the carrier, the result of good push-pull balance using factory-matched transistors. The 3rd and 5th harmonics are more than 15 dB down. To reduce the 3rd harmonic to at least 50 dB below the carrier, a five-section Chebyshev filter with low passband ripple is an appropriate choice. This type of filter has a good SWR in the passband, and a smooth roll-off characteristic. The design process began by creating ideal designs using a public-domain filter design program.[4]

Ideal designs rarely correspond to standard capacitor or inductance values that can be realized with a discrete number of turns on common toroid cores. Using a circuit analysis program,[5] the ideal designs were analyzed to see the effects of such real-world limitations on harmonic rejection and SWR performance.

First, the ideal component values were entered into the program, and varied ±20 percent to see which ones had the greatest effect on performance. C1 and C5 (see Fig 2 and Table 1) were found to be least sensitive to variations, L2 and L4 were moderately sensitive; varying C3 had the greatest effect on both passband and stopband performance. The ideal capacitor values were then replaced with standard capacitor values or—in some cases—parallel combinations of two common capacitor values. Inductors were given the nearest value available for coils wound on either T-50-2 or T-50-6 toroid cores. The final filter designs are the result of trade-

Table 1

Filter Circuit and Comparison of Ideal and Final Component Values.

	Ideal Filter Values			*Actual Filter Values*		
Cutoff Freq. (MHz)	*C1, C5 (pF)*	*C3 (pF)*	*L2, L4 (µH)*	*C1, C5 (pF)*	*C3 (pF)*	*L2, L4 (µH)*
2.40	1521	2620	4.55	1470 (1000 + 470)	2880 (2200 + 680)	4.41 (30 t on T50-2)
4.80	761	1310	2.27	830 (560 + 270)	1430 (1000 + 430)	2.37 (22 t on T-50-2)
8.76	417	718	1.25	430	820	1.25 (16 t on T-50-2)
12.18	300	516	0.900	300	560	0.960 (14 t on T-50-2)
17.22	212	365	0.634	220	370 (270 + 100)	0.706 (12 t on T-50-6)
25.74	142	244	0.424	150	240	0.460 (10 t on T-50-6)
35.64	102	176	0.306	100	180	0.314 (8 t on T-50-6)

In some cases it is necessary to parallel two smaller-value capacitors to obtain the proper values of capacitance for C1, C2 and C5

The inductors are wound on T-50-2 or T-50-6 cores. Inductors for the 160- and 80-meter filters are wound with #26 AWG wire in order to fit all turns on the cores; the other inductors are wound with #22 wire.

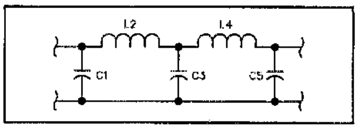


Fig 2—Schematic of the filter used for each band.

offs between inductance, capacitance and filter performance. Table 1 shows the filter topology, along with a comparison of the original ideal filter component values and the values selected for the finished unit.

Construction

I built my amplifier and low-pass filter modules on single-sided PC boards, using pads to mount the components. No holes are drilled (except for mounting screws) and all leads are attached by soldering them to the pads. The PC-board patterns for the amplifier and filters are available (see Note 2).

Fig 3 shows the amplifier-assembly parts. This assembly is mounted to a heat sink (see Figs 4 and 5) capable of dissipating more than 40 watts without excessive temperature rise. (This assumes a worst case of 50-percent transmitting time, and 7-watts dissipation in standby.) A cutout in the middle of the amplifier board allows placement of the SLAM device. The PC board leaves a conducting path around the ends of the SLAM to maintain a ground potential across the entire board. Four mechanical components make up the amplifier assembly. The first is a 0.1875-inch-thick aluminum base plate to which the SLAM is mounted. Next are two aluminum 0.1-inch-thick spacers, which are placed between the base plate and the circuit board. These spacers set the proper distance from the base plate to SLAM leads. The SLAM is installed through the top of the PC board, and its leads are soldered to the traces on top of the board.

Construction is easiest if the transformer connections to the SLAM are not soldered until after the SLAM is installed. This eliminates the possibility that the transformer connections will get in the way when you try to solder the SLAM into place. As with any power device, place a thin coating of thermal compound between the SLAM and the base plate, and between the base plate and the heat sink. Solder bypass capacitors directly to the transformer center tap and to the ground plane, with the minimum possible lead lengths.

The low-pass filter board is constructed one filter at a time. First, install the capacitor at the center (C3), then the inductors L2, L4), and finally the end capacitors (C1, C5). All inductors are wound with even spacing over three-quarters of the core circumference. Simply solder the capacitors to the pads and ground plane. Silver-mica capacitors were used in the prototype because they were on hand. Ceramic-disc capacitors with 200- to 500-V ratings will work equally well.

If the band switch is located close to the filter board (see Fig 6), short lengths of hookup wire can connect the filters to the switch wafers.

A spacious box houses the filter and amplifier assemblies, along with a TR relay that also switches the standby bias. Power and relay control leads are bypassed where they enter the enclosure.

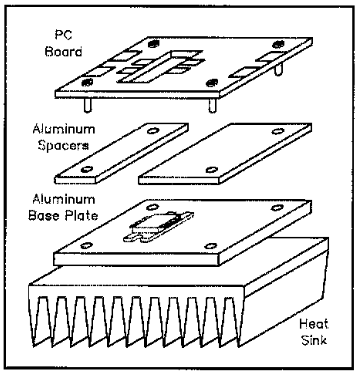


Fig 3—Mechanical assembly of the amplifier-module PC board, aluminum spacers and heat sink.

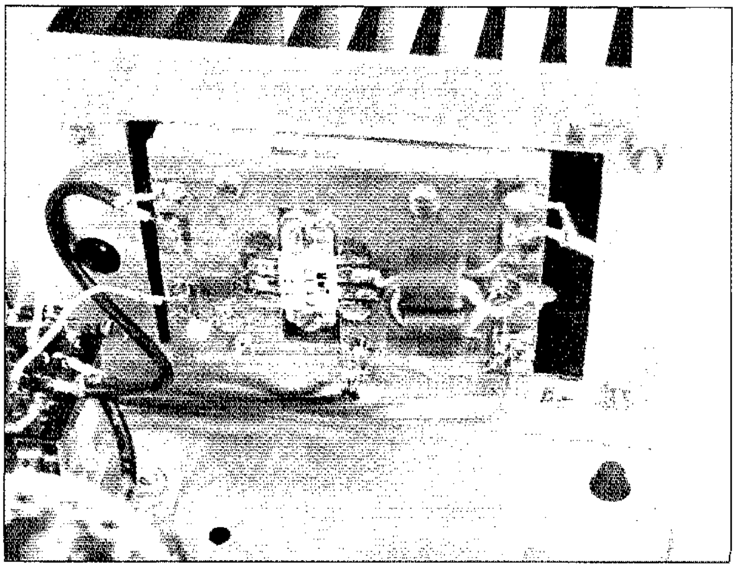


Fig 4—The assembled amplifier-module PC board in position and secured to the heat sink (see Fig 5, next page).

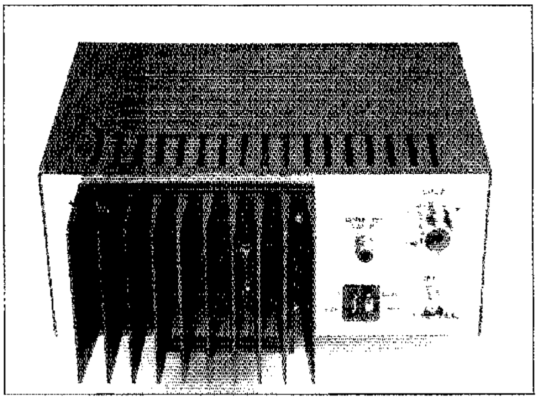


Fig 5—Rear view of the completed amplifier showing the hefty heat sink.

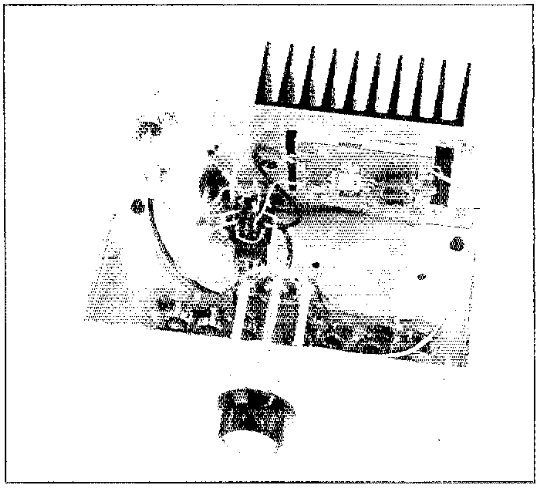


Fig 6—This interior view of the amplifier shows its simple and clean layout. The band switch is centered on the front panel. Immediately beneath the band switch is the filter assembly. Behind the switch and to the left is the TR relay, K1. A four-pin Jones plug power connector is mounted on the rear panel behind and to the left of the relay. On the bottom, near the outside lip of the rear panel, is J1. Above it is J2, with J3 to its right. Most of the rear panel—from its middle to the right lip—is occupied by the SLAM IC PC board and the aluminum spacers secured to the heat sink mounted on the rear panel's exterior. Rubber feet on the cabinet bottom help prevent scratching the supporting surface beneath and keep the amplifier from sliding. The band-switch knob center section is 1⅛ inches in diameter; the skirt flares to a diameter of 1½ inches.

Before final assembly, I gave the panels of the case a brushed look using a sanding block with oiled sandpaper. Band markings for the switch (see the title-page photo) are drawn on a large, adhesive-backed label attached to the front panel.

Performance

Amplifier gain ranges from 12.5 to 14 dB between 1.8 and 30 MHz. The gain flatness is basically a function of the input and output transformers. (It's possible to make the amplifier gain flat within 1 dB from 1 MHz to 100 MHz using transmission-line transformers and frequency compensation.) The required drive power for 25 watts output is 1.0 to 1.4 watts.

On-the-air performance is excellent. Besides low distortion in the SSB mode, a small advantage of linear amplification is a complete absence of rise and fall distortion of a CW waveform, which sometimes occurs in class-C amplifiers.

Summary

This project shows how new RF products can make home construction of amateur equipment very easy. Home-brewers can benefit from a growing trend in RF product engineering: reducing development time by using "super components" that require few external components and little engineering time to design them into a product.

A secondary purpose of this project is to show how even simple software tools can be used to speed up design. The programs used to design the amplifier's low-pass filters are inexpensive, and accurate at frequencies in the MF/HF bands. In this case, they made it possible to examine trade-offs among standard-value components for seven different filters, without having to build, measure and tweak each one.

The result is a linear power amplifier with good gain and performance. Its uncomplicated design leaves little room for error, and no fancy test equipment is needed to successfully build it. Projects this easy can make an old-timer forget about the "simpler" days of vacuum tubes!

Notes

[1] Power supplies are available from Marlin P. Jones & Assoc, Inc, PO Box 12685, Lake Park, FL 33403-0685, tel 407-848-8236; fax 1-800-432-9937.

[2] Parts kits for this project are available from Crestone Engineering, PO Box 3702, Littleton, CO 80161, tel 303-770-4709. Each kit includes all electronic and mechanical components for the amplifier module and low-pass filter assembly, including circuit boards, heat sink and rotary band switch. Kits do *not* include an enclosure, connectors, or TR relay. A 25-W kit using the SLAM-0111 is $115; a 50-W kit using the larger SLAM-0122 is $190. Add $6 per kit for shipping. Payment may be made by check, money order, VISA, MasterCard or American Express.

PC board patterns for the amplifier and filters are available free from the ARRL. Send your request to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. With your request for the BREED AMPLIFIER PC-BOARD TEMPLATE, enclose a business-size envelope with one First-Class stamp.

[3] MicroWave Technology, 4268 Solar Way, Fremont, CA 94538. Their products are distributed by Richardson Electronics, 40W267 Keslinger Road, LaFox, IL 60147, tel 708-208-2200.

[4] Mike Ellis, "A Comprehensive Filter Design Program," *RF Design,* July 1991. The program is available from the RF Design Software Service, PO Box 3702, Littleton, CO 80161-3702, tel 303-770-4709 (part #RFD-0791, $15 postpaid).

[5] *NOVA,* a shareware program by Robert Stanton, also available from the RF Design Software Service (part #RFD-0391, $15 ppd).

Gary Breed, K9AY, has been licensed since 1961. The design and construction of equipment and antennas is his favorite part of our ham radio hobby, but some serious contesting and DXing also emanates from his station. Gary is Editor of RF Design *magazine, a technical journal for design engineers. He is a member of the ARRL, RSGB and IEEE, as well as the Mile Hi DX Association and Arapahoe Radio Club.* QST

New Products

THE EASY WAY TO BUILD CLASSIC TRANSMITTERS

◊ If you've ever wanted to build simple, one- or two-tube transmitters like hams did in "the good old days," *Junkbox Projects: Circuits You Can Build With Hamfest Parts,* a new book by Robert Null, N4QR, is just what you've been waiting for.

The 20-page booklet details 20 simple rigs you can assemble from scavenged flea market parts (most use tubes, but there are a couple of transistor projects). You'll find everything from crystal sets to three-tube transmitters.

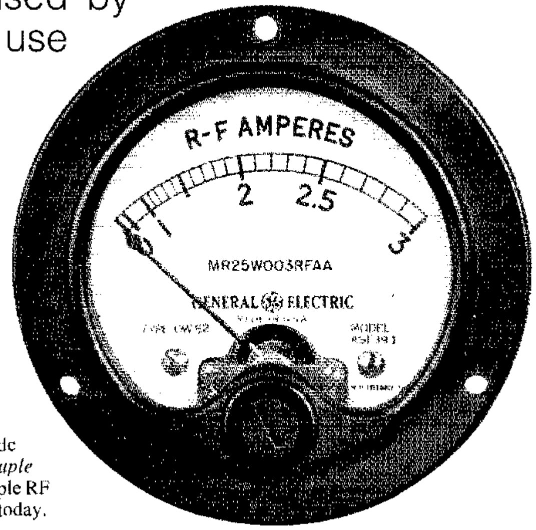
Old-timers: Dig out those 6L6 tubes; there's a lot of history in these 20 pages!

Prices: $4 postpaid in the US; $5 in Canada and Mexico; $6 elsewhere. Robert Null, N4QR, 501 N 1st Ave, Maiden, NC 28650-1105. QST

Revisiting the RF Ammeter

The antenna instrument most used by hams of yesteryear now finds use mainly in specialized nonamateur applications. Here's what RF ammeters do, and how you can use them.

By John Stanley, K4ERO
8495 Hwy 157
Rising Fawn, GA 30738



Many ham stations include at least one *directional wattmeter*, also commonly called *RF wattmeter*. We use RF wattmeters to measure forward power (the rate at which radio energy moves from our transmitters or transceivers to our antennas) and reflected power (the rate at which antenna-feed-line mismatch, if any, reflects RF power back to our transmitters or transceivers).[1] The directional wattmeter as we know it[2] is a relative latecomer, however. For much of Amateur Radio's history, hams depended on *RF ammeters* to indicate proper antenna-system performance.

Since increasing the power in a circuit increases the circuit current (power = current squared × resistance [$P = I^2R$]), an ammeter can indirectly indicate power. The $P=I^2R$ relationship also holds for a given antenna geometry (wire length and orientation). Hams of yesteryear cared little for SWR or forward and reflected power; they just tuned for maximum current in the antenna. Even after SWR became a concern, the RF ammeter was the instrument of choice for measuring it.

Measuring Current at Radio Frequencies

Dc ammeters are relatively common, but you can't use one of these meters to measure RF. (You might get no reading or, more likely, you might get smoke!)

Ac ammeters made for 60-Hz applications won't work for RF either: They include coils that act as inductors, impeding the flow of RF current. To measure RF, an ammeter must be usefully sensitive and have a low impedance at the measurement frequency. This requires a *thermal ammeter*—a special type of meter that uses heat generated by the passing current to activate an indicator of some type.

A thermal ammeter may indicate current flow by means of wire expansion caused by heating (*hot-wire* ammeters do this) or by using the voltage generated by an RF-current-heated thermocouple to drive a dc meter (as in *thermocouple* ammeters). Thermocouple RF ammeters are standard today.

Hot-Wire Ammeters

Back in the days of spark transmission, the preferred method of indicating proper antenna loading was to put a piece of iron wire in series with the antenna. RF-current heating expanded and lengthened the wire, and mechanical amplification transferred the wire's lengthening to the movement of a dial pointer (Fig 1). This approach did not allow the reading of small currents, however, and as hams switched to CW transmission (which, at first, often meant operating at powers considerably lower than those commonly used for spark), the hot-wire ammeter was not as useful as it had been.

The Thermocouple Ammeter

Heating the junction between two dissimilar metals generates a feeble dc voltage across the junction. This is the basis of the *thermocouple*. Hook a sensitive dc meter to an RF-heated thermocouple, and you have a thermocouple ammeter. Fig 2 shows an early thermocouple ammeter hams built for use between 150 and 200 meters.

RF-Ammeter Scale Nonlinearity

RF-ammeter scales are inherently nonlinear—spread out at the high end and "scrunched up" at the lower end as shown in the title picture. Sometimes you'll run into RF ammeters that have been linearized somewhat, but the nonlinear characteristic is most common. For this reason, an RF ammeter should have a full-scale value only a bit more than the expected reading so its pointer falls into an easily readable portion of its scale. If you need to measure a wide range of RF-current levels, you may therefore need several meters to do the job accurately.

The Importance of Meter Sensitivity

A thermocouple ammeter requires much less series resistance than a hot-wire ammeter to produce a given current indication. To put it another way, the thermocouple ammeter is more sensitive—its pointer moves farther for a given value of RF current—for a given meter *internal resistance*. This is important because the power dissipated by an RF ammeter's internal resistance is power that doesn't get to the antenna.

A current meter's internal resistance is also important because it causes measurement error. Used in series with a 10-ohm load, for instance, a meter with an internal resistance of 1 ohm causes a 10% error. Practical ammeters have a very low resistance, so that the decrease in current caused by putting the ammeter in series with the circuit is small. As a general rule, a really good 1-ampere meter should have less than 0.1 ohm of internal resistance.

Carefully calibrated and used, a good RF ammeter can give results to 1 or 2% accuracy, at least up to 30 MHz. This is more accurate than some expensive commercial wattmeters, and is much better than the typical home-brew power meters that many hams use with great confidence!

Signal Purity and Thermal Ammeters

Since the RF ammeter uses heat to produce meter deflection, it is inherently a "true RMS" meter. I won't go into detail as to what this means, except to say that the shape of the ac waveform greatly affects the ratio of power it produces for a given peak voltage. The exact power in a circuit can be calcu-

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[1] Notes appear on page 36.

lated only if true RMS (*root-mean-square*) current (or voltage) can be measured. Most meters actually measure either average or peak voltage or current and then multiply by the appropriate factor to find RMS values. If the measured wave is not a pure sine wave, the "appropriate factor" will vary, and you must choose the right one. With a thermal ammeter, this is automatic.[3,4]

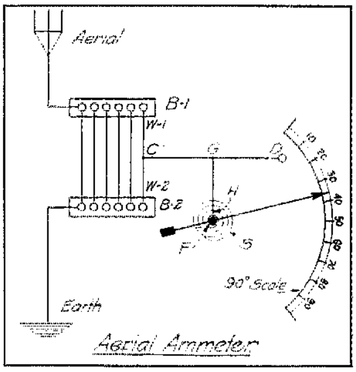

Fig 1—"An *ammeter* which any wireless man who is clever with his tools may construct at small cost.... *CD* is a piece of silk fibre about 3" long. *GH* is another piece about 2" long and is attached to the center of *CD* by a bit of beeswax. The lower end of *GH* is wound about the shaft which carries the spring *S* and the *pointer*. The spring tends to pull the pointer toward the maximum scale position, but it is resisted by the fibre thread and the wire *W-1, W-2*. When no current flows through *W-1, W-2*, the pointer rests at zero, but when the wire is heated by the passage of radio frequency currents the tension on *CD* is reduced, relieving the tension opposing the spring *S*, which, in turn, pulls the pointer across the scale.... *B-1* and *B-2* are *heavy copper lugs. W-1, W-2* is a piece of No. 40 Therlo wire (0.003" in diameter) about 4" long. Its resistance is about 9.5 ohms. Using only one wire, the instrument will measure at its maximum scale, 0.1 ampere. With several wires in parallel the range may be increased to any desired value." The station transmitter connects between *B-2* and Earth. (*from E. Bucher,* The Wireless Experimenter's Manual, *1920, page 125*)

RF Ammeter Calibration

RF ammeters made before 1939 or so have a rising response at high radio frequencies, and a correction curve must be applied.[5] Post-WWII meters are much better. A modern thermocouple ammeter has a flat response from below 50 Hz to about 60 MHz, and can be easily calibrated by connecting it in series with an accurate digital ammeter and putting 50- or 60-Hz current through both into a load that draws a current near the high end of the ammeter's scale. Alternatively, a known voltage can be applied to a thermocouple meter through a known resistance. A digital multimeter can serve as a calibration standard.

To cross-check a meter's ac calibration, test its calibration again using dc. Be aware, however, that many RF ammeters read slightly high when connected according to one polarity and low by the same amount for the opposite polarity. This is caused by a slight voltage drop across the heating element getting into the thermocouple. If the thermocouple were mounted perfectly symmetrically on the heating wire this would not occur, but it commonly does. Take readings both ways and use their average.

How you mount an RF ammeter can affect its accuracy. If the meter has a metal case, mounting it directly on a metal panel can affect its accuracy by shunting RF to ground through the meter's internal capacity. As shown in the current *ARRL Antenna Book*, mounting the meter on a nonconductive subpanel and spacing it from the metal panel can solve this problem.

Even RF ammeters in nonconductive cases must be mounted with care. A meter calibrated for steel-panel mounting will read high in a nonferrous panel; a meter calibrated for nonferrous-panel mounting will read low in a ferrous panel.[6] The best approach is to calibrate the meter after you've mounted it where you intend to use it.

Using RF Ammeters

One use of an RF ammeter is a dummy antenna with a built-in power meter. This would consist of, for example, a good 50-ohm dummy load with a 5- or 6-ampere meter in series. The 5-ampere calibration could also be marked **1250 W**, the 4-ampere calibration, **800 W**, and so on (remember: $P = I^2R$). Carefully calibrated, this arrangement can indicate your transmitter's output with more accuracy than even an expensive power meter. Of course, the dummy antenna need not be 50 ohms; it can be any value you need. But whatever power calibration you apply will be accurate only with the load resistance you used during calibration.

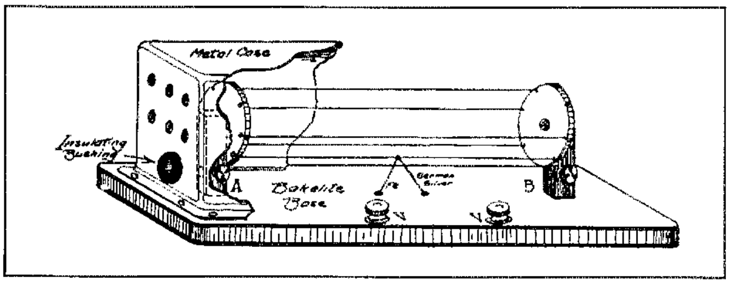

Fig 2—"A thermo-couple suitable for radio frequency measurements... The thermo-couple consists of a piece of No. 36 iron wire and another piece of No. 36 German silver wire, each 3" long. The ends of the two wires are carefully twisted together and soldered to one of the No. 40 heating wires..." Two metal discs hold six slender resistance wires that shunt most of the antenna current around the thermocouple. One of the wires heats the thermocouple. The indicating voltmeter connects at terminals *V-V.* (*from Bucher*)

This brings us to a critical fact about RF-ammeter measurements. RF current can flow into resistance, reactance or a combination of both (impedance), but only RF current flowing through a resistance does useful work. The drawback is that an RF ammeter can't distinguish reactive and resistive current.

So see Fig 3. You can assume that increasing an RF ammeter's reading means more radiated power only if your adjustments don't change anything connected to the meter output terminal. (For all you know, you might make your antenna more capacitive and less resistive. The ammeter would read higher, but the increase would go into charging the capacitance instead of radiating more signal.) Assuming that whatever you connect to the meter's output terminal stays the same, though, just about any adjustment (transceiver tuning/loading controls, antenna-tuner settings, whatever) done on the *input* side of the meter to increase the meter indication means a stronger radiated signal.

RF Ammeters are Still Useful

The directional wattmeter has largely superseded the RF ammeter for a number of sound reasons. For instance, a directional wattmeter can accurately indicate output power even in the presence of standing waves. But RF ammeters are still quite useful, especially in situations where you don't need to know absolute values of power or current. Every time you make a station improvement with the help of an RF ammeter, you can think about how we sometimes abandon technology not so much because something better comes along, but because what we've been using simply goes out of style.

Notes

[1] Once we know the ratio between a system's forward and reflected powers, we can calculate the equivalent *standing-wave ratio* (SWR), a factor traditionally of high interest to radio amateurs, as shown on page 24-8 of the current *ARRL Antenna Book*. The wattmeters commonly used by hams often include reflected-power scales calibrated directly in SWR.—*Ed.*

[2] W. Bruene, "An Inside Picture of Directional Wattmeters," *QST*, Apr 1959, pp 24-28.

[3] You don't have to worry about how your transmitted signal's waveshape may affect ammeter accuracy for another reason: If your transmitter meets FCC purity-of-emissions rules as it should, its output can be taken as sinusoidal for all practical RF-power-measurement purposes.

[4] Part 97 of the FCC Rules require that Amateur Radio power measurements be made in terms

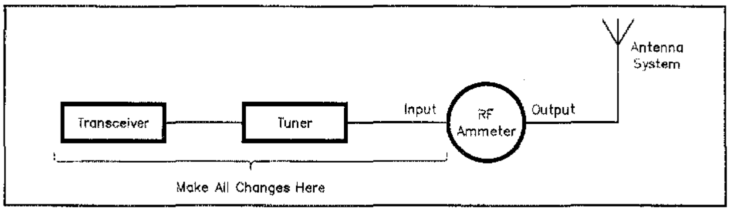

Fig 3—An RF ammeter can unambiguously indicate antenna-system improvement only if you make all of your system adjustments on its input side. Higher indications caused by changes in the antenna components connected to its output side may result from reactive current that does not increase your system's radiated power.

Sources for RF Ammeters

When it comes to getting your own RF ammeter, there's good news and bad news. First, the bad news. New RF ammeters are expensive—about $70 to $200. AM radio stations are the main users of these today. The FCC defines the output power of AM stations based on the RF current in the antenna, so new RF ammeters are made mainly for that market. They are quite accurate, and their prices reflect that.

The good news is that used RF ammeters are often available. *QST* advertiser Fair Radio Sales of Lima, Ohio, has been a consistent RF-ammeter source, for instance. Ham flea markets are also worth trying. Some grubbing around in your nearest surplus store or some older ham's junk box may provide just the RF ammeter you need.

RF Ammeter Substitutes

Don't despair if you can't find a used RF ammeter. It's possible to construct your own if you're a diehard experimenter. Both hot-wire and thermocouple types can be homemade. I'd be interested to learn of any designs you come up with!

Pilot lamps in series with antenna wires, or coupled to them in various ways, can indicate antenna current[*] or even forward and reflected power.[†]

Another approach is to use a small low-voltage lamp as the heat/light element and using a photodetector driving a meter as an indicator. (Of course, your eyes and judgment can serve as the indicating part of the instrument.) Checking feed-line balance could be as simple as a couple of lamps with the right current rating and as low a voltage as possible. You should be able to tell fairly well by eye which bulb is brighter or if they are about equal. You can calibrate a lamp-based RF ammeter with 60-Hz or dc power.

As another alternative, you can build an RF ammeter that uses a dc meter to indicate rectified RF from a current transformer that you clamp over a transmission line wire.[††]

Copper-Top Battery Testers as RF Ammeters

Finally, there are the *free* RF ammeters which come as the testers with Duracell batteries! These are actually 3- to 5-ohm resistors with built-in liquid-crystal displays. The resistor heats the liquid-crystal strip; the length of the "lighted" portion (heat turns the strip clear, exposing the fluorescent ink beneath) indicates the magnitude of the current.

Despite their + and – markings, these indicators are not polarized. Their resistance is low enough to have relatively little effect on a 50-ohm system. (For example, putting one in series with a 50-ohm dummy antenna would increase the system SWR from 1 to 1.1. The maximum current these testers can measure is in the 200- to 400-mA region. (You can achieve higher ranges by putting a carbon-composition or -film resistor in parallel with the strip.) Best of all, if you burn one out during your tests, you can replace it at any drugstore, hardware store or supermarket for a few dollars, with some batteries thrown in free!—*K4ERO*

[*] F. Sutter, "What, No Meters?," *QST*, Oct 1938, p 49.
[†] C. Wright. "The Twin-Lamp," *QST*, Oct 1947, pp 22-23, 110, 112.
[††] Z. Lau, "A Relative RF Ammeter for Open-Wire Lines," *QST*, Oct 1988, pp 15-17.

First licensed in 1956 as KN4ERO, John Stanley has since operated as VU2IE, TI2ERO, HC1JX, HC1XG/HC8, K4ERO/PJ4 and S79J.

John graduated from MIT in 1962 with a BSEE, and later took a master's degree in theology from Emory University. He has worked for Texas Instruments, UCLA Plasma Physics Lab, and radio station HCJB, Quito, Ecuador. John is an ARRL Technical Advisor. He and his wife, Ruth, WB8LUA, presently work as consultants, mainly with Christian radio stations outside the US. This work has taken them to over 40 countries. QST

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of *peak envelope power*, or PEP. Because an RF ammeter is an averaging instrument, it can accurately indicate the peak envelope power only of signals for which average and peak envelope power are the same (such as a key-down CW signal). An RF ammeter cannot correctly indicate PEP for an SSB voice signal.

[5] J. Miller, "Improved Thermo-Ammeter Construction to Increase Accuracy on Ultra-High Frequencies," *QST*, May 1938, p 44-45.

[6] G. Floyd, "Every Mil I Have is Yours, *or* Beware of Steel Panels!" *QST*, Nov 1951, pp 32, 104, 106.

New Books

RIDING THE AIRWAVES WITH ALPHA AND ZULU

By John Abbott, K6YB

artsci Inc, PO Box 1428, Burbank, CA 91507; tel 818-843-4080, fax 818-846-2298, 288 pp, 8½ × 11 inches, B&W artwork. $14.95.

Reviewed By Brian Battles, WS1O
QST *Features Editor*

A cartoon book license study manual—is this kid stuff? My children, Ian, age 9, and Jill, age 7, have had fun reading it since the night I brought it home. (These are little kids who have been exposed to a great deal of their dad's Amateur Radio operating and have had access to the ARRL's *Now You're Talking!* and other license manuals, but have yet to show serious interest in studying for exams.) With *Alpha and Zulu*, however, they actually got a kick out of the artwork and dialog, so much so that I began to think that this may be the thing that gets them started on preparing for their tickets. Next to the ARRL's *Archie's Ham Radio Adventure* comic book, this is the only Amateur Radio-oriented book my children appear to enjoy reading. I still don't know if this will be the "magic potion" that encourages them to work on getting licensed; *Alpha and Zulu* comes out and gets put away almost as often as their Ren & Stimpy comics and Barbie storybooks, so it's too soon to tell.

From an adult perspective, this looks like a good idea for younger prospective hams or anyone who prefers a "lite" approach to what's essentially a leisure-time pursuit. The book helps to introduce Morse code through its cast of lively little characters called "phoneticos." Each character is named after a letter in the standard ITU Phonetic Alphabet, and is drawn as a stick figure whose body and head are composed of large blocks (dahs) and small blocks (dits) that spell their initials in code. The reader meets short, stocky Alpha, medium-sized Charlie, Juliett looming over the others, and all the rest of the gang up to rather tall Zulu. Although it's sprinkled with some typographical errors (*it's* for *its*, etc) and *Alpha* doesn't conform to the standard ITU spelling, *Alfa*, perhaps these problems will be fixed in the next edition.

Each page features the phoneticos characters acting out a mini-skit that presents a short lesson on material covered by the Novice and Technician written examinations, and is followed by the phoneticos quizzing each other (and the reader) with a single-page self-test based on questions from the FCC license exam question pool. There are also remarks about operating, and randomly placed corny jokes and puns to keep the reading lively.

After reading this book and watching my kids flip through it, I'd say *Riding the Airwaves with Alpha and Zulu* is a fun introduction to the material needed to prepare a beginner for an Amateur Radio license examination, but I don't think it's strong enough medicine to work alone. It's a pleasant starting point. As a supplement, young students will benefit from the additional help of an adult instructor, and adult readers will benefit from the in-depth aid of an experienced Elmer or a conventional manual like *Now You're Talking!* QST

Stacking Tribanders: A Super Station—Sorta

Want to enhance the DX and contesting performance of your station, but you don't have acres and acres of land for multiple towers? How about stacking tribanders?

By R. Dean Straw, N6BV/1 and Fred Hopengarten, K1VR
32 Beacon Hill Road
Windham, NH 03087
6 Willarch Road, RFD 1
Lincoln, MA 01773

Stacked Yagis are all the rage at top contest and DXing stations, and have been for at least several decades. Many of the aptly named *Super Contest* stations employ multiple towers, festooned with stacked arrays of long-boom monoband Yagi antennas. Usually one tower is devoted to each HF band. The rest of us have neither the acreage, financial resources nor just plain ambition to construct (and maintain) stacked 5 over 5 over 5 over 5 element monobanders on 10, 15 and 20 meters, plus stacked 3 over 3-element 40-meter Yagis! Nonetheless, remarkable improvements in station performance can be achieved by stacking triband Yagis in far more modest installations. Just why do people stack Yagis?

Stacks and Wide Elevation Footprints

Detailed studies using sophisticated computer models of the ionosphere have revealed that coverage of a wide range of elevation angles is necessary to ensure consistent DX or contest coverage on the HF bands. These studies have been conducted over all phases of the 11-year solar cycle, and for numerous transmitting and receiving sites throughout the world. Table 1 is an example of such a

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[1] Notes appear on page 44.

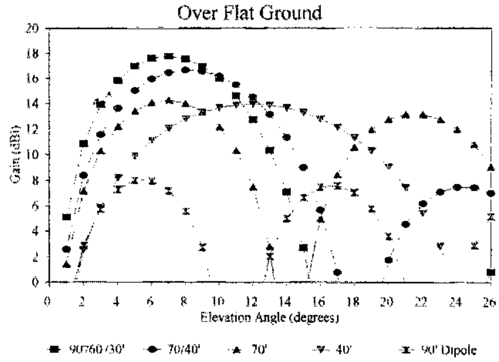

Fig 1—Comparison of 10-meter TH7DX tribander combinations. The 10-meter stack of three at 90, 60 and 30 feet has an elevation footprint between 3.5° to 11° at its half-power points, and a peak gain of 17.8 dBi. The stack at 70 and 40 feet has a peak gain of 16.7 dBi at 8°, with coverage from 4° to 12.5° at its half-power points. Note the use of dBi (isotropic antenna in free space) for gain rather than dBd (dipole). For example, a dipole at 90 feet might seem like a good reference for the stack of three, and at 7° elevation, the 17.8 dBi peak gain of the stack of three is almost 10 dB (10 dBd) greater than the gain of the 90-foot dipole. However, at 11°, where the dipole is in a null, the 14.6 dBi gain of the three-stack is 32 dB stronger than the dipole—this would be a gain of 32 dBd! Clearly, dBi is a more realistic measure of performance than dBd when stacks are used.

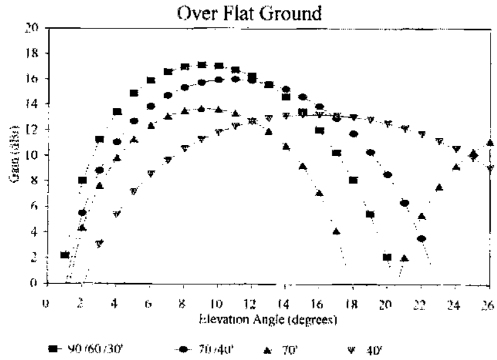

Fig 2—Comparison of 15-meter TH7DX tribander combinations. The stack at 90, 60 and 30 feet yields an excellent footprint over the range of 4° to 14° at its half-power points, with a peak gain of 17.1 dBi. The stack at 70 and 40 feet has a peak gain of 16.0 dBi at 11°, with coverage from 5° to 17° at its half-power points. Like the 10-meter stack of three, the stack of two TH7DXs is very close in overall performance, except for lower gain at very low angles, where the higher top antenna comes into play in the stack of three.

Table 1

Range of Elevation Angles from New England to Europe

Band	*Elevation angles for 99% of the time to Western Europe*	*Elevation angles for 99% of the time to Eastern Europe*
80 meters	5.2° - 33.0°	13.4° - 23.7°
40 meters	4.9° - 19.3°	3.0° - 17°
20 meters	3.3° - 17.0°	1.4° - 13.0°
15 meters	3.8° - 13.8°	1.0° - 11.7°
10 meters	4.6° - 14.0°	1.0° - 12.8°

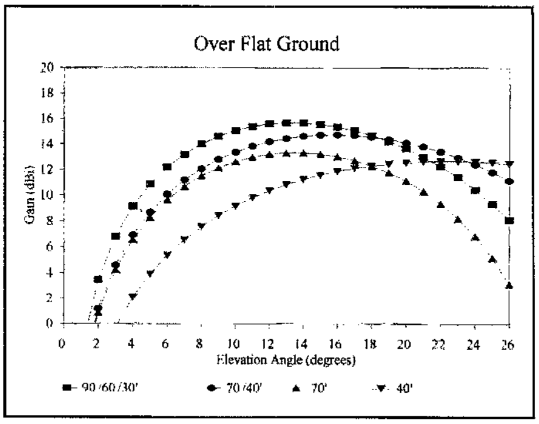

Fig 3—Comparison of 20-meter TH7DX tribander combinations. The peak gain for the 90, 60 and 30-foot stack is 15.7 dBi at 13° elevation. The –3 dB elevation coverage is from 6.5° to 21.5°. The peak gain for the stack of two at 70 and 40 feet is 14.7 dBi at 16°, and the –3 dB elevation coverage is from 7.5° to 25°. The stack of three has proven to be an extremely effective antenna.

study using a program called *IONCAP*[1] for the path from New England to both Western and Eastern Europe, listing the statistical range of elevation angles covering 99% of the time that signals arrive over the whole 11-year solar cycle. Different tables are required to describe paths from New England to other parts of the world, and to describe the paths from other transmitting sites to various parts of the world.

Fig 1 shows the computed elevation response for various combinations of Hy-Gain TH7DX triband Yagis on 10 meters, calculated using Brian Beezley's *MNC*[2] program. The highest curve is for a stack of three TH7DXs at heights of 90, 60 and 30 feet, placed on one tower above flat ground with an average conductivity and dielectric constant. Overlaid on the same graph are the elevation patterns for a single TH7DX at 70 feet, representing a fairly typical station setup. Also shown is the pattern for a single TH7DX at 40 feet, the pattern for a stack of two TH7DX tribanders at 70 feet and 40 feet on one tower and the pattern for a single 90-foot-high TH7DX.

Note that the elevation patterns show gain in *dBi*; that is, they are referenced to an isotropic antenna rather than to a dipole as many amateurs are more accustomed to seeing. It is difficult to pick a single height above ground for a reference dipole when comparing it to a stack of Yagis at different heights. The response for a 90-foot-high 10-meter dipole is shown on the graph to illustrate the difficulty in trying to compare a single reference dipole to a stack having a broad elevation pattern. This is most obvious at the angles where the reference dipole exhibits a null, such as the region between 9.5° to 12° for the 90-foot-high dipole, where the "gain over a dipole" would be artificially large. A reference of dBi, for an isotropic antenna in free-space, avoids this dilemma.

At 10 meters, the stack of three triband Yagis at 90, 60 and 30 feet has good coverage for low elevation angles, and good coverage out to about 11° elevation, where its pattern crosses that of the single 40-foot-high antenna. At an elevation of 2°, the stack of three has 8 dB more gain than the single 40-foot-high antenna, but only 2 dB of gain over the stack of two antennas at 70 and 40 feet. For the range of angles needed to cover Western and Eastern Europe, the race between the stack of three and the shorter stack of two is pretty close. A single TH7DX on 10 meters at 90 feet suffers dramatically whenever the elevation angles are higher than approximately 9°, as commonly occurs into Western Europe during the strongest part of the 10-meter opening from New England. Both of the stacks illustrated here give a wider *elevation footprint* than any single antenna, so that all the angles can be covered automatically without having to switch from higher to lower antennas manually. This is perhaps the major benefit of using stacks, but not the only one.

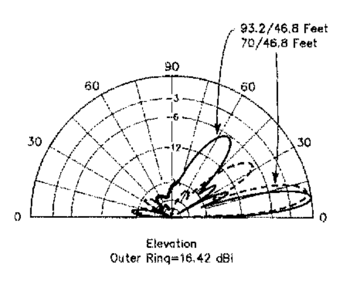

Fig 4—The effect of stacking distance for 15-meter TH7DXs. The stack at 93.2 and 46.8 feet (one-wavelength spacing) has a lower peak elevation angle (because of the top antenna's height) and just slightly more stacking gain than does the stack at 70 and 46.8 feet. The exact distance between practical HF Yagis is not critical to obtain the benefits of stacking. For a stack of tribanders at 90, 60 and 30 feet, the distance in wavelengths between individual antennas is 0.87 λ at 28.5 MHz, 0.65 λ at 21.2 MHz, and 0.43 λ at 14.2 MHz.

Fig 2 compares the 15-meter elevation responses for tribanders at the same heights as for 10 meters. Here, the best system is also the stack of three at 90, 60 and 30 feet, followed by the stack of two at 70 and 40 feet. For most of the time, the single Yagi at 70 feet is down from the stacks by at least 3 dB. The stack of three at an elevation of 8° has a gain of about 7 dB over the single tribander at 40 feet. Again, either 15-meter stack gives a wider elevation footprint than any single antenna does.

Fig 3 shows the 20-meter elevation response for the same triband antennas. The edge in favor of the bigger stack narrows somewhat compared to the other antennas, mainly because the 30-foot spacing (0.43 λ) between antennas in the stack is more of a compromise for gain on 20 meters than for the upper bands. However, the stack of three still gives a gain of 6 dB over the single 40-foot-high tribander at a 10° elevation angle, and has a wider elevation footprint than any single antenna.

Stacks and Compression of the Forward and Rearward Elevation Lobes

The basic principle of a stacked array is that it concentrates energy from higher angle lobes (which don't contribute much to communications anyway) into the main elevation lobe. The stack squeezes down the main elevation lobe, while maintaining the frontal lobe azimuth pattern of a single Yagi. This is the reason why many state-of-the-art contest stations are stacking arrays of relatively short-boom antennas, rather than stacking long-boom, higher-gain Yagis. A long-boom HF Yagi narrows both the azimuthal pattern and the elevation pattern, making pointing of the antenna more critical, and making it more difficult to spread a signal over a wide azimuthal area, such as all of Europe and Asiatic Russia at one time.

The compression of the higher-angle lobes has another desirable effect, beyond that of creating more gain. It reduces QRM from high-angle signals arriving from the direction in which the antenna is pointed,

Are You *Nuts*? Why *Tribanders?*

Without a doubt, the most common question we are asked when people hear about our antenna systems is: "Why did you pick *tribanders* for your stacks?" We chose triband antennas with full recognition that they are compromise antennas. Other enterprising amateurs have built stacked tribander arrays. Bob Mitchell, N5RM, is a prominent example, with his so-called *TH28DX* array[3] of four TH7DX tribanders on a 145-foot-high rotating tower. Mitchell employed a rather complex system of relay-selected tuned networks to choose either the upper stacked pair, the lower stacked pair or all four antennas in stack. Others in Texas have also had good results with their tribander stacks. Contester Danny Eskenazi, K7SS, has very successfully used a pair of stacked KT-34XA tribanders for years.

A major reason why we used tribanders is that over the years both authors have had good results using TH6DXX or TH7DX antennas. They are ruggedly built, mechanically and electrically. They are able to withstand New England winters without a whimper, and their 24-foot long booms are long enough to produce significant gain, despite trap-loss compromises. Amateurs speculating about trap losses in tribanders freely bandy about numbers between 0.5 and 2 dB. We are comfortable with the lower figure, as are the Hy-Gain engineers with whom we have discussed trap losses. Consider this: If 1500 watts of transmitter power is going into an antenna, a loss of 0.5 dB amounts to 163 watts. This would create a significant amount of heat in the six traps that are on average in use on a TH6DXX, amounting to 27 watts per trap. If the loss were as high as 1 dB, this would be 300 watts total, or 50 watts per trap. Common sense says that if the overall loss were greater than about 0.5 dB, the traps would act more like big *firecrackers* than resonant circuits!

A long-boom tribander like the TH6DXX or TH7DX also has enough space to employ elements dedicated to different bands, so that the compromises in element spacing usually found on short-boom 3- or 4-element tribanders can be avoided.

Another factor in the conscious choice of tribanders was first-hand frustration with the serious interaction that can result from stacking monoband antennas closely together on one mast in a Christmas Tree configuration. N6BV's worst experience was with the ambitious 10- through 40-meter Christmas Tree at W6OWQ in the early 1980s. This installation used a Tri-Ex SkyNeedle tubular crankup tower with a rotating 10-foot long heavy-wall mast. The antenna which suffered the greatest degradation was the 5-element 15-meter Yagi, sandwiched five feet below the 5-element 10-meter Yagi at the top of the mast, and five feet above the full-sized 3-element 40-meter Yagi, which also had five 20-meter elements interlaced on its 50-foot boom. The front-to-back ratio on 15 meters was at best about 12 dB, down from the 25+ dB measured with the bottom 40/20-meter Yagi removed. No amount of fiddling with element spacing, element tuning or even orientation of the 15-meter boom with respect to the other booms (at 90° or 180°, for example) improved its performance. Further, the 20-meter elements had to be lengthened by almost a foot *on each end of each element* in order to compensate for the effect of the interlaced 40-meter elements. It was a lucky thing that the tower was a motorized crankup, because it went up and down hundreds of times as various experiments were attempted!

Interaction due to close proximity to other antennas in a short Christmas Tree can definitely destroy carefully optimized patterns of individual Yagis. Nowadays, interaction can be modeled using a computer program such as *MN* or *MININEC* or *NEC*. A gain reduction of as much as 2 to 3 dB can easily result due to close vertical spacing of monobanders, compared to the gain of a single monoband antenna mounted in the clear. Curiously enough, at times such a reduction in gain can be found even when the front-to-back ratio is not drastically degraded, or when the front-to-back occasionally is actually *improved*. In his excellent book, Dave Leeson, W6QHS,[4] mentions that the 10-meter Yagi in his closely stacked Christmas Tree (15 meters at the top, 10 meters in the middle, and 20 meters at the bottom of the rotating mast) loses "substantial gain" because of serious interaction with the 20-meter antenna. (We calculated that the free-space gain in the W6QHS stack drops to 5 dBi, compared to about 9 dBi with no surrounding antennas.) Monobanders are *definitely not* universally superior to tribanders in multiband installations! In private conversations, W6QHS has indicated that he would not repeat this kind of short Christmas Tree installation again. He also has indicated that his 37-foot-high TH7DX is a great antenna for working Japan, although it does help to be located on a commanding mountaintop.

Last, in the N6BV/1 installation, triband antennas were chosen because the system is meant to be as simple as possible, given a certain desired level of performance, of course. Triband antennas make for less mechanical complexity than do an equivalent number of monobanders. There are five Yagis on the N6BV/1 tower, yielding gain from 40 to 10 meters, as opposed to using 12 or 13 monobanders on the tower!

The verdict? Both the N6BV/1 and K1VR systems play well. We can each be beaten by some local stations with monoband stacks. Whether this happens because they use monoband antennas compared to our triband antennas, or whether we get beaten because they command superior hilltop locations is open to debate. We both have been fortunate enough to place in the *Top Ten* in our respective competition categories in a number of major DX Contests.

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and from high-angle signals coming from other directions, such as local QRM. A stack also squeezes down the elevation response of the rearward lobe, just like the forward lobe. On the negative side, however, the front-to-back ratio of a stack is often degraded compared to that of a single, optimized Yagi, although this is not usually a severe problem.

By definition, a stack of triband Yagis has a constant vertical spacing between antennas in terms of feet or meters, *not* in terms of wavelength. There is a great deal of folklore and superstition among amateurs about stacking distances. *There is nothing magical about stacking distances for practical HF Yagis.* The gain gradually increases as spacing in terms of wavelength is increased between individual Yagis in a stack, and then decreases slowly once the spacing is greater than about 1.0 λ. The difference in gain between spacings of 0.5 λ and 1.0 λ for a TH7DX Yagi amounts to only a fraction of a decibel. Fig 4 shows the elevation patterns for two 15-meter TH7DXs stacked at 70 and 46.8 feet (half-wavelength spacing), and at 93.2 and 46.8 feet (one-wavelength spacing). The elevation footprint for the higher stack has slightly more gain at lower angles, as expected, and the peak gain is just slightly higher, but the stack with the smaller spacing still has a good gain and a desirable pattern. The situation is different on VHF, where truly long-boom, high-gain designs are practical and desirable, and where stack spacing is correspondingly more critical because of complex mutual coupling and interaction between the antennas.

Stacks and Fading

We have solicited a number of reports from stations, mainly in Europe, to compare various combinations of antennas in stacks and as single antennas. The peak gain of the stack is usually just a little bit higher than that for the best of the single antennas, which is not surprising. Even a large stack has no more than about 6 dB of gain over a single Yagi at a height favoring the prevailing elevation angle. Fading on the European path can easily be 20 dB or more, so it is very confusing to try to make definitive comparisons. We have noticed over many tests that the stacks are much less susceptible to fading compared to single Yagis. Even within the confines of a typical SSB bandwidth, frequency-selective fading occasionally causes the tonal quality of a voice to change on both receive and transmit, often dramatically becoming fuller on the stacks, and tinnier on the single antennas. This doesn't happen all the time, but is often seen. We have also observed that the depth of a fading is less, and the period of fading is longer, on the stacks compared to single antennas.

Exactly *why* stacks exhibit less fading is a fascinating subject, for which there exist a number of speculative ideas, but little hard evidence. Some maintain that stacks outperform single antennas because they can afford *space diversity* effects, where by virtue of the difference in physical placement one antenna will randomly pick up signals that another one in another physical location might not hear. This space diversity argument is difficult to argue with, and equally difficult to prove scientifically. A more plausible explanation about why stacked Yagis exhibit superior fading performance is that their narrower frontal elevation lobes can discriminate against undesired propagation modes. Even when band conditions favor, for example, a very low 3° elevation angle on 10 or 15 meters from New England to Western Europe, there are signals, albeit weaker ones, that arrive at higher elevation angles. These higher-angle signals have traveled longer distances on their journey through the ionosphere, and thus their signal levels and their phase angles are different from the signals traversing the primary propagation mode. When combined with the dominant mode, the net effect is that there is destructive and constructive fading. If the elevation response of a stacked antenna can discriminate against signals arriving at higher elevation angles, then in theory the fading will be reduced.

Stacks and Precipitation Static

John Kenny, W1RR, has pointed out that the top antenna in a stack is often much more affected by rain or snow precipitation static than is the lower antenna. We have also observed this phenomenon, where signals on the lower antenna by itself are perfectly readable, while S9+ rain static is rendering reception impossible on the higher antenna or on the stack. This means that the ability to select individual antennas in a stack can sometimes be very important. We recommend that you don't permanently hard-wire antennas in a stack together.

Stacks and Azimuthal Diversity

Azimuthal diversity is a term coined to describe the situation where one of the antennas in a stack is purposely pointed in a direction different from the main direction of the stack. During most of the time in a DX contest from the East Coast, the lower antennas in a stack are pointed into Europe, while the top antenna is often rotated toward the Caribbean or Japan. In a stack of three identical Yagis, the first-order effect of pointing one antenna in a different direction is that one-third of the transmitter power is diverted from the main target area. This means that the peak gain is reduced by 1.8 dB, not a very large amount if signals are 10 to 20 dB over S9 anyway.

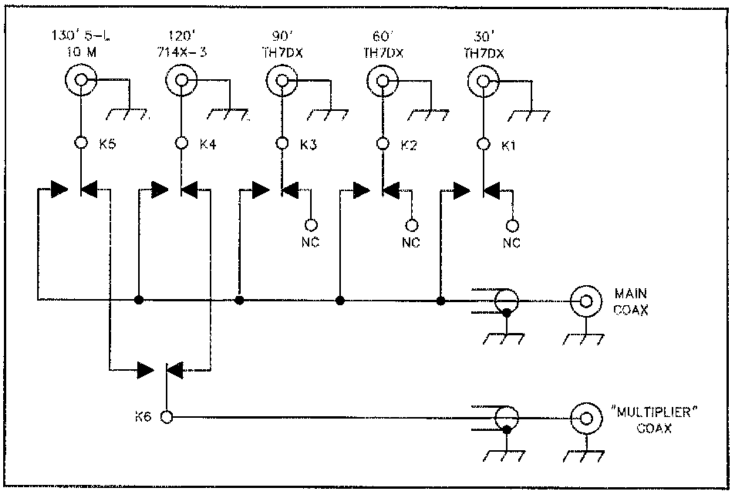


Fig 5—N6BV/1 switch box system. This uses a modified DX Engineering remote switch box, with relay K6 added to allow selection of either of the two top antennas (5-element 10-meter Yagi or 40/20/15-meter triband 714X-3) as a "multiplier" antenna. Nota bene: There is no special provision for SWR equalization when any or all of the Yagis are connected in parallel as a stack fed by the main coaxial cable. Each of the five Yagis is fed with equal lengths of flexible Belden 9913 coax, so that phasing can be maintained on any band. The main and "multiplier" coaxes going to the shack are 0.75-inch-OD Hardline cables.

The N6BV/1 Antenna System

The N6BV/1 system in Windham, New Hampshire, is located on the crest of a small hill about 40 miles from Boston, and could be characterized as a good, but not dominant, contesting station. There is a single 120-foot-high Rohn 45 tower, guyed at 30-foot intervals, with a 100-foot horizontal spread from tower base to each guy point so that there is sufficient room for rotation of individual Yagis on the tower. Each set of guy wires employs heavy-duty insulators at 57-foot intervals, to avoid resonances in the 80- through 10-meter amateur bands. There are five Yagis on the tower. A heavy-duty 12-foot-long steel mast with 0.25-inch walls is at the top of the tower, turned by an Orion 2800 rotator. Two thrust bearings are used above the rotator, one at the top plate of the tower itself, and the other about two feet down in the tower on a modified rotator shelf plate. The two thrust bearings allow the rotator to be removed for service.

At the top of the mast, 130 feet high, is a 5-element, computer-optimized 10-meter Yagi, which is a modified Create design on a 24-foot boom. The element tuning has been modified from the stock antenna in order to achieve higher gain and a better pattern over the band. At the top of the tower (120-foot level) is mounted a Create 714X-3 triband Yagi. This is an interesting tribander, with a 32-foot boom and five elements. Three elements are active on 40 meters; four are active on 20 meters, and four are active on 15 meters. The 40-meter elements are loaded with coils, traps and capacity hats, and are approximately 46 feet long. A triband 20/15/10-meter Hy-Gain TH7DX tribander is fixed into Europe at the 90-foot level on the tower, just above the third set of guys.

At the 60-foot level on the tower, just above the second set of guys, there is a "swinging-gate" side-mount bracket, made by DX Engineering of Oregon. A Hy-Gain Tailtwister rotator turns a TH7DX on this side mount. (Note that both the side mount and the element spacings of the TH7DX itself prevent full rotation around the tower; about 280° of rotation is achieved with this system.) At the 30-foot level, just above the first set of guys, is located the third TH7DX, also fixed on Europe. All five Yagis are fed with equal lengths of Belden 9913 low-loss coaxial cable, each measured with a noise bridge to ensure equal electrical characteristics. At each feed point a ferrite-bead choke balun (using 7 large beads) is placed on the coax. All five coaxial cables go to a relay switch box mounted at the 85-foot level on the tower. Fig 5 shows the schematic for the switch box, which is fed with 250 feet of 75-Ω, 0.75-inch-OD Hardline coaxial cable. The stock DX Engineering remote switch box has been modified by adding relay K6, so that either the 130-foot or the 120-foot rotating antenna can be selected through a second length of 0.75-inch Hardline going to

The K1VR Array: A More Elegant Approach to Matching

Stacks! Just imagine. "The antenna system here is 4 over 4 on 10, 3 over 3 on 15, and 3 over 3 on 20. The upper and lower Yagis are both rotatable." K1VR has that system, and it works. It was put together on a single tower in his backyard; with careful planning, many radio amateurs could do something similar.

The K1VR stacked array is on a 100-foot-high Rohn 25 tower with sets of guy wires at 30, 60 and 90 feet, made of nonconducting Phillystran. Phillystran is a nonmetallic Kevlar rope which is covered by black polyethylene to protect against the harmful effects of the sun's ultraviolet rays. A caution about Phillystran: Don't allow tree branches to rub against it. It is designed to work in tension, but, unlike steel guy wire, it does not tolerate abrasion well.

Both antennas are Hy-Gain TH6DXX tribanders, with the top one at 97 feet and the bottom one at 61 feet. The lower antenna is rotated by a Telex Ham-M rotator on a homemade "swinging-gate" side mount, which allows it to be rotated 300° around the tower without hitting any guy wires or having an element swing into the tower. At the 90-foot point on the tower, a 2-element 40-meter Cushcraft Yagi has been mounted on a TIC General RingRotor so that it can be rotated 360° around the tower.

After several fruitless attempts trying to match the TH6DXX antennas so that either could be used by itself or together in a stack, K1VR settled on using a relay-selected broadband toroidal matching transformer. When both triband antennas are fed together in parallel as a stack, it transforms the resulting 25-Ω impedance to 50 Ω. The transformer is wound on a T-200-A powdered-iron core, available from Amidon, Palomar Engineering, Ocean State Electronics or RadioKit. Two lengths of twin RG-59 coax (sometimes called Siamese or WangNet), four turns each, are wound on the core. Two separate RG-59 cables could be used, but the Siamese-twin cable makes the assembly look much more tidy. The shields of the RG-59 cables are connected in series, and the center conductors are connected in parallel. See Fig A for details.

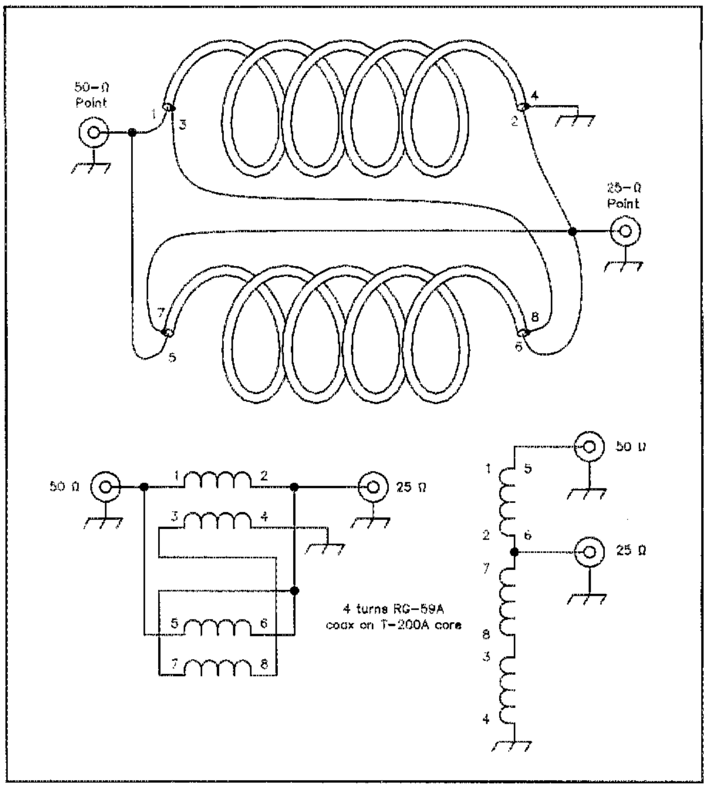


Fig A—Diagram for matching transformer for the K1VR stacked tribander system. The core is a powdered-iron core T-200A, with four turns of two RG-59A or "Siamese" coax cables. Center conductors are connected in parallel and shields are connected in series to yield a 0.667:1 turns ratio, close to the desired 25-Ω to 50-Ω transformation.

Fig B shows the schematic of the switch box, which is located in the shack. Equal electrical lengths of 50-Ω Hardline are brought from the antennas into the shack and then to the switch box. Inside the box, the relay contacts were soldered directly to the SO-239 chassis connectors to keep the wire lengths down to the absolute minimum. K1VR used a metal box that was larger than might appear necessary because he wanted to mount the toroidal transformer with plenty of clearance between it and the box walls. The toroid is held in place with a piece of insulation foam board.

Before placing the switch box in service, the system was tested using two 50-Ω dummy loads, with equal lengths of cable connected in parallel to yield 25 Ω. The maximum SWR measured was 1.25 at 14 MHz, 1.3 at 21 MHz and 1.15 at 28 MHz, and the core remained cold with 80 watts of continuous output power.

One key to the system performance is that K1VR made the electrical lengths of the two Hardlines the same (within 1 inch) by using a borrowed TDR (time domain reflectometer). Almost as good as Hardline, K1VR points out, would be to cut exactly the same length of cable from the same 500-foot roll of RG-213. This eliminates manufacturing tolerances between different rolls of cable.

K1VR's experience over the last 10 years has been that at the beginning of the 10- or 15-meter morning opening to Europe the upper antenna is better. Once the band is wide open, both antennas are fed in phase to cast a bigger shadow, or footprint, on Europe. By

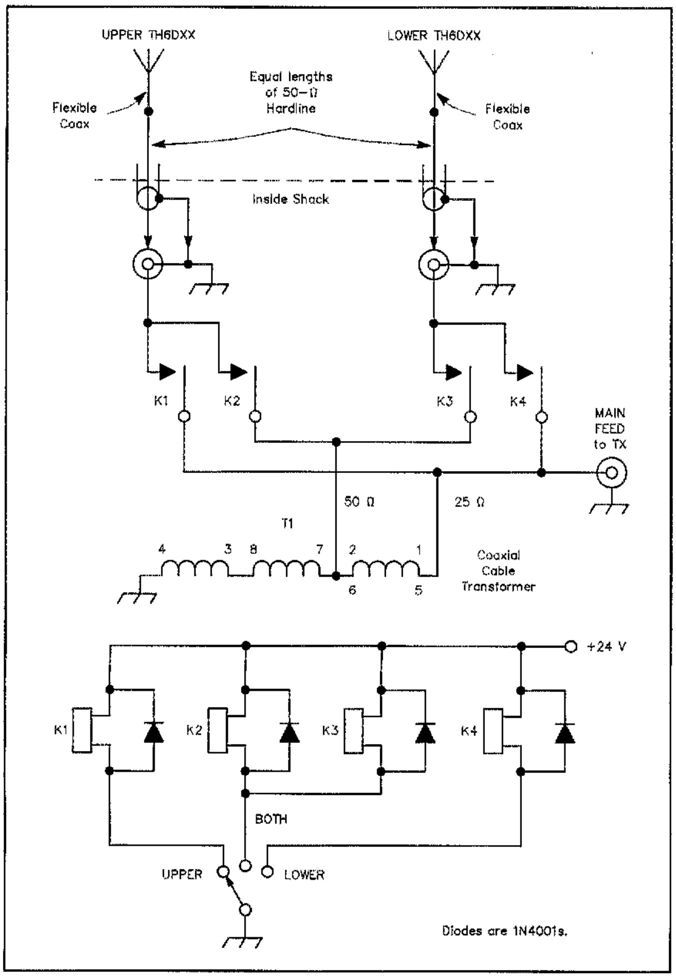


Fig B—Relay switch box for the K1VR stacked tribander system. Equal lengths of 50-Ω Hardline (with equal lengths of flexible 50-Ω cable at each antenna to allow rotation) go to the switch box in the shack. The SWR on all three bands for UPPER, LOWER or BOTH switch positions is very close to constant.

mid-morning, the lower antenna is better for most Europeans, although he continues to use the stack in case someone is hearing him over a really long-distance path through Europe. He reports that it is always very pleasant to be called by a 4S7 or HS0 or VU2 when he is working Europeans at a fast clip!

The next step in the evolution of the K1VR array is to put a third TH6DXX on the tower, at the 31-foot level. This should improve materially his 10- and 15-meter footprints into Europe. However, he will find himself quite challenged to come up with simple matching networks if he still wants to work with a constant SWR, no matter which triband antenna is switched in or out of the array.

K1VR also points out that the tribanders need not be TH6DXXs. A pair of bigger KLM KT34XA tribanders or smaller Cushcraft A4S tribanders could be used. Fred notes a pair of A4Ss with 40-meter add-on kits could be used, to provide stacked dipoles on 40 meters. The spacing on 40 meters would be quite close (about 0.25 λ), but it would be fun to try.

the shack. This creates a *multiplier* antenna, independent of the *main* antennas. A second band can be monitored in this fashion while calling CQ using the main antennas on another band. Band-pass filters are required at the multiplier receiver to prevent overload from the main transmitter.

The 0.75-inch Hardline has very low losses, even when presented with a significant amount of SWR at the switch-box end. This is important because, unlike K1VR's system (see the sidebar "The K1VR Array: A More Elegant Approach to Matching"), no attempt is made to maintain a constant SWR when relays K1 through K5 are switched in or out. This seemingly cavalier attitude comes about because of several factors. First, there are many different combinations of antennas that can be used together in this system. Each relay coil is independently controlled by a toggle switch in the shack. N6BV could not manage to devise a matching system that did not become incredibly complex because of the numerous impedance combinations used over all the three bands. Second, the worst-case additional transmission-line loss due to a 4:1 SWR mismatch when four antennas are connected in parallel on 10 meters is only 0.5 dB. It is true that a linear amplifier must be retuned slightly when switching combinations of antennas in and out, but this is a small penalty to pay for the reduced complexity of the switching and matching networks. The 90/60/30-foot stack is used for about 95% of the time during DX contests, so the small amount of amplifier retuning for other antenna combinations is considered only a minor irritation.

Some Suggestions for Stacking Tribanders

We don't imagine that many amateurs will try to duplicate exactly our contest setups. However, many hams already have a tribander on top of a moderately tall tower, typically at a height of about 70 feet. It is not terribly difficult to add another, identical tribander at about the 40-foot level on such a tower. The second tribander can be pointed in a fixed direction of particular interest (such as Europe or Japan), or it can be rotated around the tower on a side mount or a Ring-Rotor (made by TIC General). If guy wires get in the way of rotation, the antenna can usually be arranged so that it is fixed in a single direction. Insulate the guy wires at intervals to ensure that they don't shroud the lower antenna electrically. A simple feed system consists of equal-length runs of surplus 0.5-inch, 75-Ω Hardline (or more expensive 50-Ω Hardline, if you are really obsessed by SWR) from the shack up the tower to each antenna. Each tribander is connected to its respective Hardline feeder by means of an equal length of flexible coaxial cable, with a ferrite choke balun, so that the antenna can be rotated.

Down in the shack, the two Hardlines can simply be switched in and out of parallel to

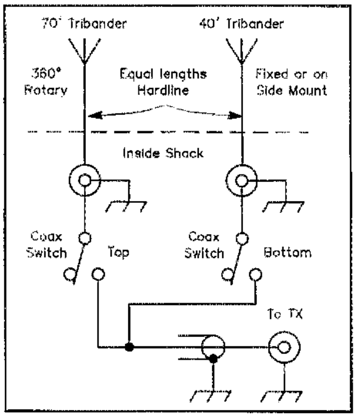


Fig 6—Suggested simple feed system for 70/40-foot stack of tribanders. Each tribander is fed with equal lengths of 0.5-in., 75-Ω Hardline cables (with equal lengths of flexible coax at the antennas to allow rotation), and can be selected singly or in parallel at the operator's position in the shack. Again, no special provision is made in this system to equalize SWR for any of the combinations.

select the upper antenna only, the lower antenna only, or the two antennas as a stack. See Fig 6. Any impedance differences can be handled as stated previously, simply by retuning the linear amplifier, or by means of the internal antenna tuner (included in most modern transceivers) when the transceiver is run barefoot. The extra performance experienced in such a system will be far greater than the extra decibel or two that modeling calculates, believe us!

Conclusion

The vagaries of the ionosphere are exceedingly, incredibly complex. The ionosphere is the great equalizer in HF radio. Ken Wolff, K1EA, the owner of a famous contesting station with large stacked arrays high on a hill, stated recently that as a big gun he is really very proud to be so useful to so many little pistol stations. It seems that just about *everybody* has a story of how he managed to beat out K1EA in at least one pileup. The ionosphere blesses just about everyone sometime or another. But the well-stacked stations are the ones who get blessed consistently.

Notes

[1] *IONCAP* is a computer program used by a number of governmental agencies to model the ionosphere; the propagation graphs in the *How's DX* column in *QST* are generated using *IONCAP*. The program requires a lot of computing resources, plus a tremendous amount of patience to use, because it is notoriously "unfriendly" to users.

[2] *MN* or *MNC*, by Brian Beezley, K6STI, 507½ Taylor, Vista, CA 92084.

[3] R. Mitchell, "The TH28DX, How To Stack Four TH7DX Antennas On A 145 Foot Self-Rotating Tower," *CQ*, Aug 1986, pp 11-21.

[4] D. Leeson, W6QHS, *Physical Design of Yagi Antennas*, (Newington: ARRL, 1992), pp 7-13.

W1AW schedule

Pacific	Mtn	Cent	East	Sun	Mon	Tue	Wed	Thu	Fri	Sat
6 am	7 am	8 am	9 am			Fast Code	Slow Code	Fast Code	Slow Code	
7 am	8 am	9 am	10 am			Code Bulletin				
8 am	9 am	10 am	11 am			Teleprinter Bulletin				
9 am	10 am	11 am	noon							
10 am	11 am	noon	1 pm		Visiting Operator Time					
11 am	noon	1 pm	2 pm							
noon	1 pm	2 pm	3 pm							
1 pm	2 pm	3 pm	4 pm	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code
2 pm	3 pm	4 pm	5 pm	Code Bulletin						
3 pm	4 pm	5 pm	6 pm	Teleprinter Bulletin						
4 pm	5 pm	6 pm	7 pm	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code
5 pm	6 pm	7 pm	8 pm	Code Bulletin						
6 pm	7 pm	8 pm	9 pm	Teleprinter Bulletin						
6^{45} pm	7^{45} pm	8^{45} pm	9^{45} pm	Voice Bulletin						
7 pm	8 pm	9 pm	10 pm	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code
8 pm	9 pm	10 pm	11 pm	Code Bulletin						
9 pm	10 pm	11 pm	Mdnte	Teleprinter Bulletin						
9^{45} pm	10^{45} pm	11^{45} pm	12^{45} am	Voice Bulletin						

❑ ***Morse code transmissions:***

Frequencies are 1.818, 3.5815, 7.0475, 14.0475, 18.0975, 21.0675, 28.0675 and 147.555 MHz.

Slow Code = practice sent at 5, 7½, 10, 13 and 15 WPM.

Fast Code = practice sent at 35, 30, 25, 20, 15, 13 and 10 WPM.

Code practice text is from the pages of *QST*. The source is given at the beginning of each practice session and alternate speeds within each session. For example, "Text is from July 1992 *QST*, pages 9 and 81," indicates that the plain text is from the article on page 9 and mixed number/letter groups are from page 81.

Code bulletins are sent at 18 WPM.

❑ ***Teleprinter transmissions:***

Frequencies are 3.625, 7.095, 14.095, 18.1025, 21.095, 28.095 and 147.555 MHz.

Bulletins are send at 45.45-baud Baudot and 100-baud AMTOR, FEC Mode B. 110-baud ASCII will be sent only as time allows.

On Tuesdays and Saturdays at 6:30 PM Eastern time, Keplerian elements for many amateur satellites are sent on the regular teleprinter frequencies.

❑ ***Voice transmissions:***

Frequencies are 1.855, 3.99, 7.29, 14.29, 18.16, 21.39, 28.59 and 147.555 MHz.

❑ ***Miscellanea:***

A DX bulletin replaces or is added to the regular bulletins between 8 PM Eastern time Thursdays and 8 PM Eastern time Fridays.

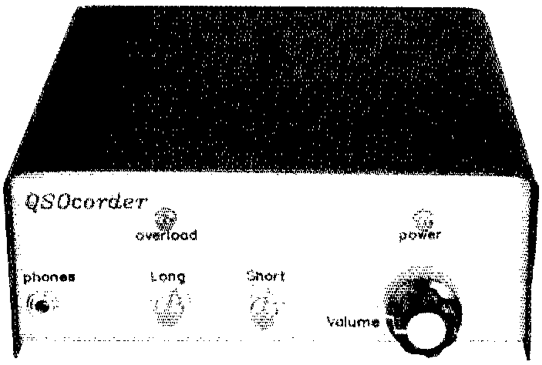
W1AW is open to visitors during normal operating hours: from 1 PM until 1 AM on Mondays, 9 AM until 1 AM Tuesday through Friday, from 1 PM to 1 AM on Saturday, and from 3:30 PM to 1 AM on Sundays. FCC licensed amateurs may operate the station from 1-4 PM Monday through Saturday. Be sure to bring your current FCC amateur license or a photocopy.

In a communications emergency, monitor W1AW for special bulletins as follows: voice on the hour, teleprinter at 15 minutes past the hour, and CW on the half hour.

Headquarters and W1AW are closed on New Year's Day, President's Day, Good Friday, Memorial Day, Independence Day, Labor Day, Thanksgiving and the following Friday, and Christmas Day. On the first Thursday of September, Headquarters and W1AW will be closed during the afternoon.

The QSOcorder

You'll never have to say *"Say again?"* again. The QSOcorder can provide you with an instant replay of the audio you missed.



By Steven E. Reyer, WA9VNJ

PO Box 17821
Milwaukee, WI 53217

When I'm operating, especially HF SSB or CW, it seems I invariably have to ask the other station for a repeat. In a ragchew, the repeat may be needed because of an unfamiliar name or QTH, or a sudden distraction. In a contest, everything happens so fast that the exchange is gone in a second. On CW, a momentary memory lapse, and *poof!* A little interference, some missed words or a sentence, and the QSO screeches to a halt.

Wouldn't it be great if you could easily replay the previous few seconds of audio from the receiver? Sort of like saying "Say again," but without saying it? Now you can. The QSOcorder is a digital signal processing (DSP) system that lets you run your QSOs more easily and efficiently. You simply ask the QSOcorder for a repeat of up to 10 seconds of previously received audio, by pushing a button. This is all accomplished digitally (including the button push...) with no tapes or moving parts to wear out. Best of all, you can build the QSOcorder yourself! A PC board and complete kit of parts is available to make the job easy and enjoyable.[1]

How It Works

You simply insert the QSOcorder between your receiver and speaker. While you're listening to the receiver, the QSOcorder continually records the incoming audio. In the Automatic mode, it operates like an endless-loop tape recorder, but without the tape. At any time, you can press the **SHORT** or **LONG** button to replay the previous 5 or 10 seconds of audio, respectively. As long as you press the button, the QSOcorder replays the audio. When you hear the information you want, you release the button and are instantly back to real-time audio. The QSOcorder works with any audio delivered by the receiver to the speaker.

In the Manual mode, the QSOcorder waits for you to trigger it to record. You may want to capture a particular portion of incoming audio. The QSOcorder records that segment and holds it for as many replays as you need. As long as power is applied to the unit, the sound segment stays in memory.

I always keep my QSOcorder in-line and running in the Automatic mode. I've found it handy when trying to work DX in a pileup. It's also been a good learning tool for working CW, or when I'm trying to follow a CW net. Sometimes I just need a little extra help when working a new station. Listening to an unfamiliar voice with a new name and QTH—especially under conditions of heavy QRM or QRN—can be a real challenge. Here's where the QSOcorder can help you to be a sharp operator.

The QSOcorder Circuit

Fig 1 is a block diagram of the QSOcorder; the schematic is shown in

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[1] See page 48.

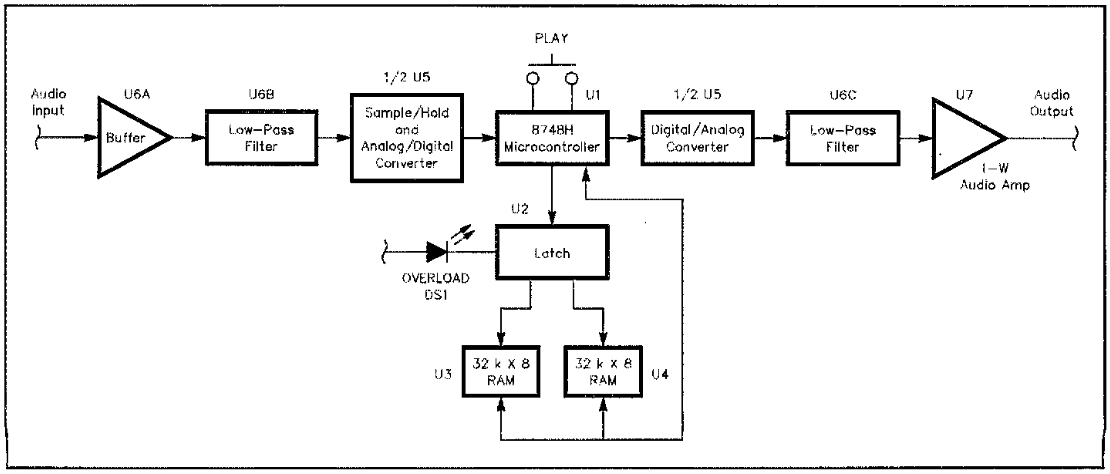
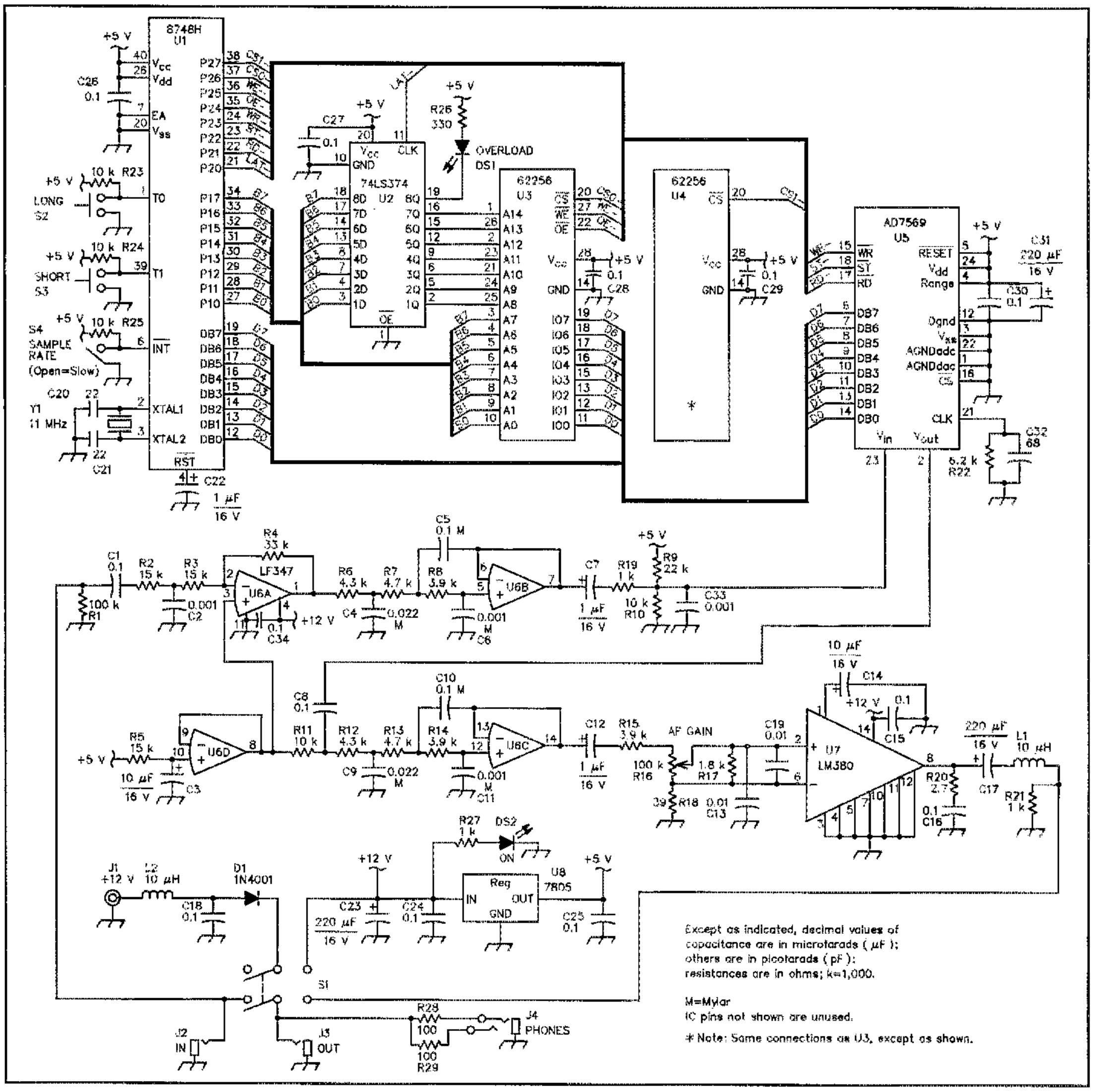


Fig 1—QSOcorder block diagram.



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

M=Mylar

IC pins not shown are unused.

* Note: Same connections as U3, except as shown.

Fig 2—Schematic of the QSOcorder circuit. Unless otherwise specified, resistors are ¼-W, 5%-tolerance, carbon-composition or film types. Equivalent components can be used. The input anti-alias and output reconstruction filters are identical Sallen-Key active R-C Chebyshev low-pass filters with 3-kHz cutoffs and a passband ripple of about 0.5 dB.

C20, C21—22-pF, 5%-tolerance ceramic or silver mica.
C32—68-pF, 5%-tolerance ceramic or silver mica.
DS1—Red LED.
DS2—Green LED.
J1—Coaxial dc power jack.
J2, J3—Single-circuit phone jack.
J4—Stereo phone jack.
L1, L2—10-µH choke.
R16—100-kΩ linear-taper potentiometer.
S1—DPDT toggle switch.
S2, S3—SPST normally open, momentary-contact switch.
S4—SPST toggle switch.
U1—8748H microcontroller (*must be programmed with QSOcorder software*; see Note 1). *Do not use an 8748 without the H suffix.*
U2—74LS374 octal flip-flop.
U3, U4—62256 32-k × 8 (256-kbit) static RAM (any speed).
U5—Analog interface subsystem (Analog Devices AD7569; available from Newark Electronics. Contact Newark at 312-784-5100 to locate your nearest distributor).
U6—LF347, LM348 or TL074 quad op amp.
U7—LM380 audio power amplifier.
U8—LM340T-5 (or 7805) 5-V, 1-A voltage regulator, TO-220 case.
Y1—11-MHz fundamental crystal, 20-pF load (see text).
Misc: Enclosure (Radio Shack 270-253A), PC board, IC sockets (1-40 pin, 2-28 pin, 1-24 pin narrow, 1-20 pin, 2-14 pin), TO-220 heat sink and #4 hardware.

Fig 2. You can drive the QSOcorder from the 8-ohm speaker output jack of most rigs. The unit exhibits a high impedance (20 kΩ) load to the receiver and requires a maximum signal of 2.5 V P-P. Audio from the **IN** connector passes through a third-order, 0.5-dB ripple, Chebyshev low-pass filter (U6A and U6B) with a cutoff frequency of 3 kHz.

Following the filter, the signal is presented to U5, the AD7569 audio I/O port, which provides sample-hold and analog-to-digital (A/D) conversion functions. The 8-bit samples flow into U1 where decimation and storage into the two 32-kbyte RAMs (U3, U4) takes place. (See the sidebar "Digital Audio Storage Methods.") When data is recovered during any play mode, it's interpolated by U1 and sent to U5 again, this time using U5's digital-to-analog (D/A) converter. This analog voltage passes through another Chebyshev low-pass filter U6C, which smooths this staircase voltage and sends it to U7, the LM380 power amplifier, to drive a speaker.

U1, the heart of the system, contains 1 kbyte of program EPROM, 64 bytes of RAM, and 27 I/O lines. Only a crystal (Y1) and capacitor (C22) are needed to get it running. Although the 8748H lacks some features of newer microcontrollers, none of those features are needed. Also, the IC is readily available from many parts and surplus sources. (Be careful: *Do not substitute* the slower 8748 for the 8748H! It will *not* run at 11 MHz!) The 8748H is programmed (see Note 1) in assembly language for optimal speed, precise timing, and most efficient EPROM use. The QSOcorder program is about 600 lines long.

Address latch U2, a 74LS374, allows the software to multiplex information at the P1 port. An extra bit on the latch provides drive for the **OVERLOAD** LED (DS1).

Power requirements are nominally 200 mA at 12 V dc. (The voltage source can range from 8 to 16 V. You can use your transceiver's 12-V power supply, for instance.) If you use a dc supply of the type that plugs into an ac outlet, be sure its current rating is sufficient. Note, too, that some supplies of that type have *poor* filtering. You may have to get a better-quality supply if hum is a problem. DS2 acts as a power-on indicator. The 5-V power supply is a conventional linear design (U8), with a reverse-polarity-protection diode (D1).

Construction

An inside view of the prototype is shown in Fig 3. Of course, the easiest way to build the QSOcorder is to use the parts kit, which includes a PC board (see Note 1). But if you decide to wire-wrap or "dead bug" the circuit, here are a few pointers to keep in mind.

Physically separate the analog and digital portions of the circuit. Lay out the IC sockets to minimize the wire lengths used. That'll ensure the chips are optimally placed so the signals travel the minimum distances, lessening the likelihood of signal radiation. Be sure to include the bypass capacitor at each IC's power connection. Don't "daisy-chain" the power and ground connections from chip to chip. Instead, connect the individual power and ground leads to the heavy power bus where it enters the board. This helps to minimize ground loops and power-supply noise. Keep the panel-component leads short.

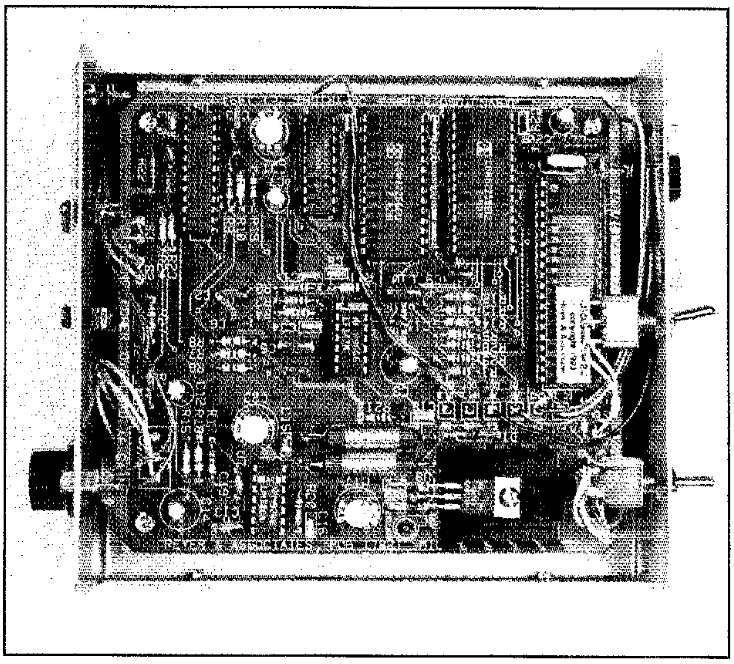


Fig 3—The 4.5 × 5-inch PC board fits neatly inside a Radio Shack 270-253A metal utility cabinet. The cover and bottom shell have been chopped approximately an inch to achieve a lower profile (see the title-page photo).

Wire-wrapping is a workable technique for this circuit. Unfortunately, there are many passive components that aren't easy to include in a wire-wrap design. In my prototype, I mounted such components on headers and wire-wrapped to them. The headers were mounted close to the associated chips. The 0.1-μF bypass capacitors were soldered directly to the IC socket bottoms.

The resistor and capacitor values associated with U6B and U6C are relatively critical. Use Mylar capacitors where shown. The components at U5 pin 21 should ideally have the tolerances shown. If the crystal frequency isn't *exactly* 11 MHz, the software will still work fine. Although 11 MHz is the stated upper limit of the 8748H processor, I've successfully used commonly available 11.0592-MHz microprocessor crystals.

EMI (Electromagnetic Interference) Considerations

L1 and L2 provide RF isolation for the low-impedance power-input and audio-output connections. The high-impedance audio-input point is RF protected by the R2-C2 network. The processor circuitry, running at 11 MHz, is capable of generating RFI of its own. As is true with any ham gear, the QSOcorder should be mounted in a grounded metal enclosure. My PC-board version is enclosed in a Radio Shack 270-253A cabinet with the top and bottom shell sheared down an inch to improve its appearance. Bypassing signal lines running into and out of the cabinet helps minimize EMI.

Operating the QSOcorder

Connect the QSOcorder between your rig's speaker output and an external speaker. Turn on the rig and QSOcorder. DS1 will light for about one second while the QSOcorder clears its RAM.

Advance the QSOcorder **AUDIO GAIN** control (R16) about one-third open. Then, increase the receiver audio gain. You should hear the receiver audio emanating from the speaker. With a signal present, advance the receiver audio gain control until **OVERLOAD** LED just starts to blink. Decrease the receiver audio gain slightly until the LED stops blinking, or does so only occasionally. This is the optimum setting for the receiver audio gain control. (Once it's adjusted, *don't touch it!*) From now on, use the *QSOcorder's* **AUDIO GAIN** control to adjust the audio level. This ensures that the input signal optimally uses the A/D converter range.

Most modern receivers have adequate automatic gain control circuits (AGC) to

Digital Audio Storage Methods

There are numerous ways of storing audio, including means of compressing it to fit in a small storage space. In the QSOcorder, I took the simple approach of sampling the audio in 8-bit samples at a rate between 13 and 18 thousand samples per second, then *decimating* the data to reduce the effective rate so as to fit the sampled audio into the available RAM space. A discussion of these techniques is beyond the scope of this article, but the following references provide more information:

K. Balmforth, "The Electronic Parrot," *QST*, Dec. 1988, pp 14-23.

J. Jarrett, "ChipTalker," *QST*, Dec. 1991, pp 17-22.

D. Hershberger, "Low-Cost Digital Signal Processing for the Radio Amateur," *QST*, Sept. 1992, pp 43-51.

S. Tretter, *Introduction to Discrete-Time Signal Processing*, (New York: Wiley, 1976), pp 21-22.

A. Oppenheim, ed, *Applications of Digital Signal Processing*, (Englewood Cliffs, NJ: Prentice-Hall, 1978).

keep the audio level roughly constant. If the input level is too high, you'll notice distortion. If it's too low, the audio sounds hissy. The QSOcorder is more forgiving than many DSP systems to incorrect audio adjustment, but it's best to obey the **OVERLOAD** LED.

Automatic Mode

When first turned on, the QSOcorder is running in the Automatic mode. The audio you hear is being recorded in RAM, with the most recent 10 seconds always available for replay. Press and hold the **LONG** button (S2) to start replaying from 10 seconds ago. When you release it, you'll hear real-time audio, then the QSOcorder resumes recording. Using the **SHORT** button (S3) starts playback from 5 seconds ago. In the Automatic mode, the QSOcorder replays only while either button is pressed.

SAMPLE RATE switch (S4) controls the rate at which samples are taken and thus has an influence on the playback sound quality and the recording time. In the **FAST** position, audio played back from the memory sounds most natural, but the recording time is reduced by about 25% (to 7.5 and 3.7 seconds) compared to that obtained with the **SLOW** setting (10 and 5 seconds).

Thus, in the **SLOW** position, the playback times are longer, but the audio may sound slightly imperfect. Some listeners don't notice the difference—experiment. I generally use the **FAST** position. Of course, both recording and playback should be at the same speed. Switching in between causes unusual sound effects (they're funny—try it!).

The normal listening audio sounds excellent regardless of the **SAMPLE RATE** switch setting. The only possible noticeable effects occur during playback, when either the **LONG** or **SHORT** button is pressed. (The actual sample rates are 13.41 and 17.55 kHz.)

Manual Mode

In the Manual mode, you can have the QSOcorder wait to record a particular audio segment. Then, you trigger it to record 10 seconds of audio. After that, the QSOcorder stops recording and holds the segment for replay. You can replay the segment as often as you like as long as power remains applied to the QSOcorder.

To enter the Manual mode, keep the **LONG** button pressed while you power up the QSOcorder, then release the button. DS1 stays lit, indicating the QSOcorder is in the Manual mode and ready to be triggered. When the desired audio arrives, momentarily press the **LONG** button: DS1 extinguishes and recording starts. When it's complete, DS1 lights again. At this point, the sound segment is in memory. To replay it, momentarily press the **LONG** button.

In Manual mode, the QSOcorder *always* records and plays the full 10-second duration, even though only a momentary press of the **LONG** button is made. You can replay the stored audio segment as often as you like. To exit Manual mode and return to Automatic mode, momentarily press the **SHORT** button (or turn the QSOcorder off and back on).

In the Manual mode, the QSOcorder can serve as a transmit buffer. To do this, change the connections, with the input source being an amplified microphone, and the output feeding the mike circuit of your transmitter. I'll leave this experimentation up to you, as every shack is slightly different.

Summary

The QSOcorder's advantages can best be appreciated after actually using it. It allows you to instantly replay any received audio. This gives you a second chance at understanding an unfamiliar name or phrase, or to compensate for in-shack or radio noises. During casual operating, the QSOcorder makes the QSO more comfortable and enjoyable. In a more stress-filled situation, such as net operation or an emergency, it helps to lighten the load.

Acknowledgments

Many thanks to Dave, W9GR, for his analog filter design, and for other invaluable contributions and excellent suggestions. And a special thanks to my wife, Linda, for her unfailing help, understanding, and encouragement.

Steve Reyer was licensed in 1967 at the age of 17. He holds a PhD in electrical engineering from Marquette University and currently is a Professor of Electrical Engineering at the Milwaukee School of Engineering. Steve has published dozens of articles in electronics hobby, Amateur Radio and engineering research journals. Steve is also an industrial consultant in the areas of digital signal processing, communications, and microprocessor systems.

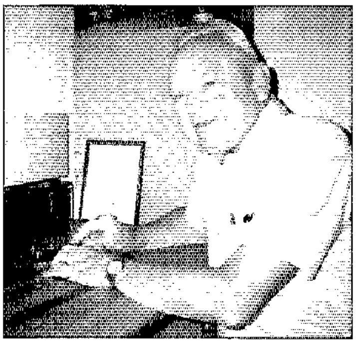
---

[1] A parts kit, including the PC board, all PC board and front panel components, including the *preprogrammed* 8748H microcontroller, is available. Not included in the parts kit are the cabinet, dc power supply, cables and connectors. The parts kit price is $95, plus $7 shipping and handling to the US and Canada. Foreign orders please include $20 for postage and handling. All prices quoted are for US funds drawn on a US bank and are subject to change without notice. Wisconsin residents must add 5.5% sales tax. Send orders to: Reyer & Associates, PO Box 17821, Milwaukee, WI 53217.

Software for this project is available for personal, noncommercial use. It is available as 8748H source code and as Intel format hexadecimal files that are used when programming the 8748H EPROM. The cost is $25 postpaid within the US and Canada. Please specify 5.25- or 3.5-inch PC/MS-DOS floppy disk format. The software is also available on the ARRL BBS (203-666-0578) as *QSOCORD.ZIP*, and also by FTP on the Internet site world.std.com (in the /pub/hamradio/arrl files area).

A PC-board template package is available free of charge from the ARRL. This is a *double-sided* PC board. To obtain a template package, address your request for the REYER QSOCORDER PC BOARD TEMPLATE PACKAGE to the ARRL Technical Department Secretary, 225 Main St, Newington, CT 06111. Please include a business-size SASE.

Strays



Proving that it's never too late or too difficult to achieve something you want, Elmer Kretzschmar, KC5BVE, of San Antonio, Texas, just earned his Novice license at age 80. His favorite mode is 40-meter CW. *(photo courtesy of James Kretzschmar, N4HCJ)*

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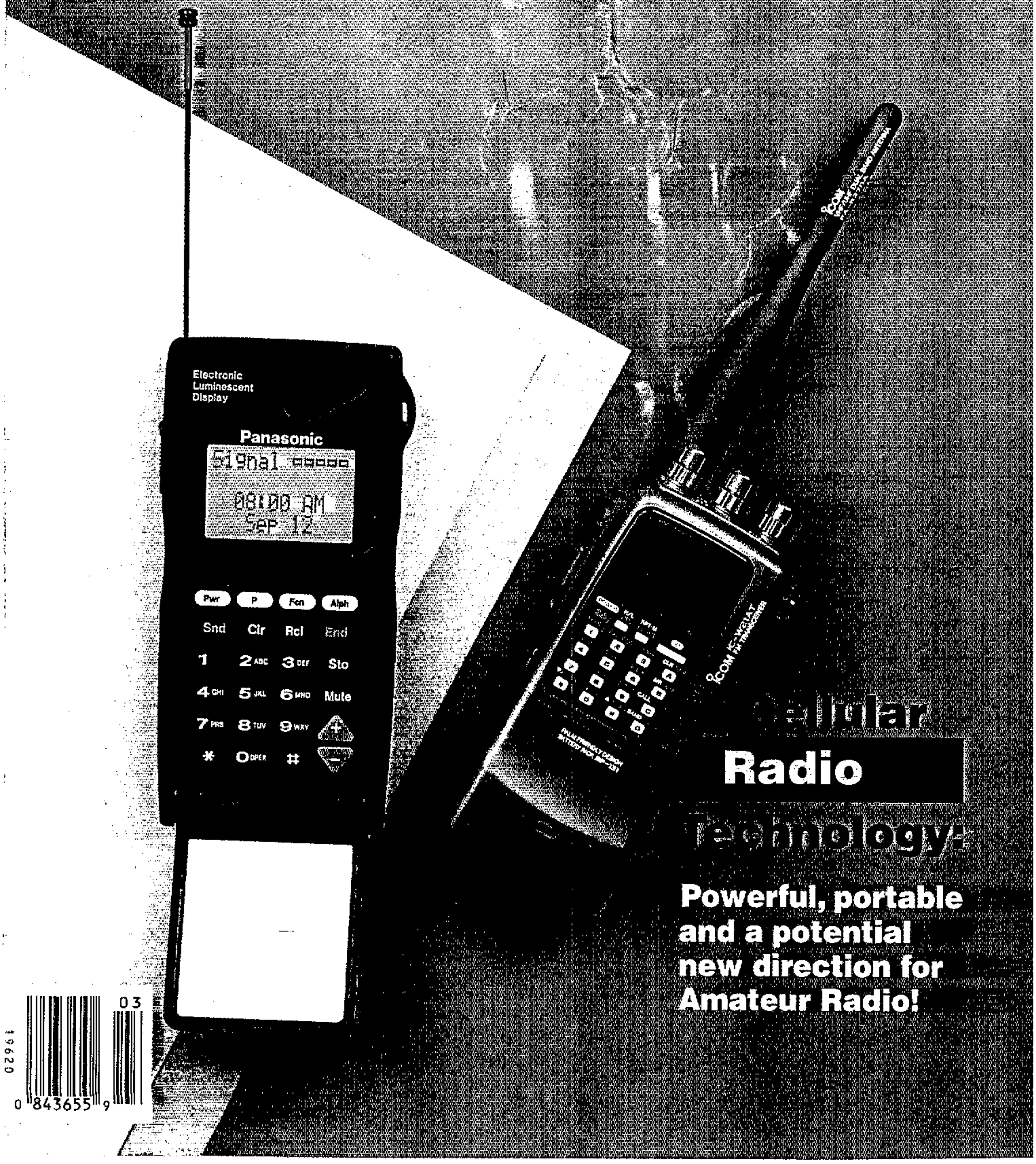

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OUR COVER

They're both hand-held radio transceivers, but how closely are they related? What influence might cellular radio technology have on Amateur Radio? Communications consultant Norman Stone, WG1C, takes a look at these RF "cousins" in the article on page 50. *(photo by Kirk Kleinschmidt, NT0Z)*

CONTENTS

March 1994
Volume LXXVIII Number 3

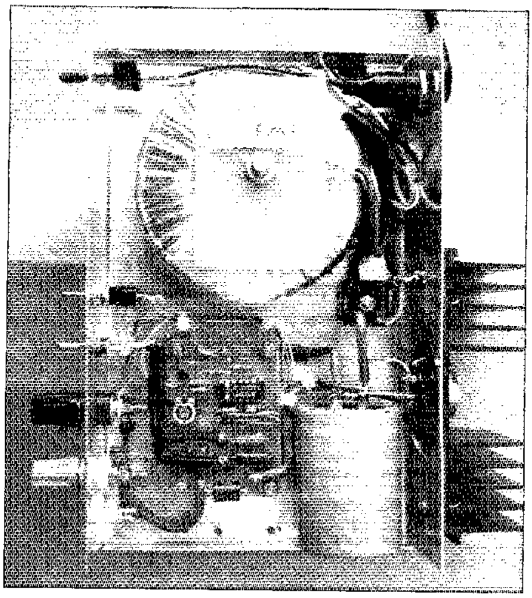
TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS



A Lead-Acid Battery Charger

This battery charger is designed specifically for use with sealed lead-acid batteries, such as those found in burglar alarms, emergency lighting units and older portable telephones.

By Ben C. Spencer, G4YNM
33 New King St
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England

This project should be of interest to all amateurs who, at any time, operate their stations from batteries, and of special interest to ARES/RACES and RAYNET[1] members. There's nothing rare or hard-to-find in this design—even the batteries can be salvaged from various sources. Most of the circuit components can be found in your (or your buddy's) junk box, or bought cheaply at ham fests. Construction and testing are straightforward and a PC board is available.[2]

A Bit About Lead-Acid Batteries

Typical batteries are shown in Figure 1. The center 12-V battery was liberated from a burglar alarm and the outer 6-V batteries liberated from a defective portable cellular telephone; separate 2-V cells are also readily available.

Two 6-V packs, or six 2-V cells, can be connected in series to provide 12 V, and so on. Although *obtaining* sealed lead-acid batteries is easy, *charging* them presents something of a problem.

A sealed lead-acid battery has the following charging requirements:

- The battery charging voltage must not exceed 2.45 V per cell.
- The maximum current charge rate must not exceed 15% of the battery's ampere-hour (Ah) rating.
- The battery must not be overcharged.

Irrespective of the discharged battery's initial state, a properly operating battery charger should charge the battery, give you an indication of the charging status and switch itself off when the battery is fully charged. This one (which I affectionately call The Buzzard) does all this, and it can also be used for charging sulfuric-acid-based car batteries.

Circuit Description

Refer to Figure 2. The charger is protected by F1 and turned on and off by S1. T1 steps down the line voltage to 18 V RMS, which is then rectified by U3 and filtered by C1; the raw dc voltage is approximately 26.

The right-hand side of the circuit comprises the voltage regulator. The regulator's output voltage is set by R9, R10 and R11:

$$V_{OUT} = 1.25\ [1 + (R10 + R11) \div R9] + I_{ADJ}(R10 + R11) \qquad \text{(Eq 1)}$$

where $I_{ADJ} = 50\ \mu A$. This provides an output of approximately 14.75 V, of which 0.6 V is dropped across diodes D3 and D4, presenting 14.15 V at the battery charging terminals (+B and –B).

U2 can deliver up to 5 A on a continuous basis. However, sealed lead-acid batteries *must not* be charged at a rate above 15% of their ampere-hour capacity, so a current-limiting circuit is required for any battery with a 33-Ah capacity or less. Maximum current is drawn when a battery in a low state of charge is connected to the battery-charging terminals. The current-limiting circuit is composed of R6, current-sensing resistor R8, and Q1, allowing a maximum current flow of 3 A.

Voltage developed across R8 is proportional to the current through it. When that voltage reaches 0.6 V, the current turns on Q1, pulling Q1's collector and the **ADJ** pin of U2 to ground, causing the regulator output voltage to fall. Any attempt to draw more current results in a greater decrease in U2's voltage. The current limit is set purely by the value of R8; placing S3 and R7 into the circuit allows for two charge rates: **STANDARD** (R7) and **RAPID** (R8). Selecting R7 provides a current limit of 1 A.

A control line can be used when required to drive Q1 into conduction to force the output of U2 to a minimum (1.25 V) whether or not a battery is connected, thus offering total control over U2's output.

Off-Charge State

When the unit is first powered up, the voltage at U1A pin 2 is significantly lower than the voltage at U1A pin 3, so the output of U1A is approximately 26.5 V. Via R3 and D2, current fed to Q1's base causes it to saturate, pulling the **ADJ** pin of U2 to ground and setting U2's output to 1.2 V. An additional 0.6 V is dropped across D3 and D4, resulting in about 0.6 V output at the battery charging terminals (+B and –B); R13 acts as a light load for D3 and D4.

A battery can be connected to the charging terminals without any risk of sparks flying, or a partially discharged battery discharging into the charger (because D3 and D4 immediately become reverse-biased).

U1A's output also drives op amps U1B and U1C, which feed LEDs DS1 and DS2. Because U1A's output is high, the green LED, DS1, is lit to show that the charger is switched on. (Following a charging cycle, illumination of DS1 represents a completed charge cycle.)

On-Charge State

If a discharged battery is connected to the battery charging terminals, D3 and D4

---

[1] Notes appear on page 27.

are reverse-biased as U2 is initially in the off-charge mode. To start a charging cycle, it's necessary to either pull U1A pin 3 to *ground, or pull U1A pin 2 to the positive* rail. When S2 (**START**, a momentary contact DPST push-button switch) is operated, it ties U1A pin 2 to the positive rail via R1, and connects U1A pin 3 to ground. U1A's output state changes causing U1B and U1C to flop, turning off DS1 and lighting red *LED DS2 to indicate that the battery is* charging and hence, is ready for use.

When U1A's output changes state, it removes the drive to Q1, allowing U2 to raise its output voltage. The current flowing through the discharged battery also flows through R8, which may cause Q1 to conduct (depending on the state of the battery), thus forcing current limiting. In any event, the current through R8 is amplified by noninverting dc amplifier U1D, which has a low-frequency gain of about 180:

$$G = (R4 + R5) + R5 = 184 \qquad \text{(Eq 2)}$$

As S2 has now been released (and provided sufficient current flows through R8), U1D's output is above the reference on U1A pin 3. Then, the output of U1A is held low, Q1 is turned off and the charger remains in *the on-charge mode.*

When the current through R8 falls to about 10% of its maximum value, U1D's output voltage is equal to the reference voltage on the noninverting input (pin 3) of U1A. An additional slight decrease in current causes U1D's output to fall below the voltage at U1 pin 3, causing U1A to flip. This drives Q1 into saturation, shutting down the regulator to 1.2 V output and returning the charger to the off-charge state with DS1 again illuminated to indicate the *completion of charging.*

At this point, the battery is about 90% charged and, hence, at 14.7 V the battery is prevented from discharging into the charger by (the now) reverse-biased diodes D3 and D4, so the battery terminal voltage decays to its natural standby voltage of about 13.2 V.

It isn't worth trying to charge the battery

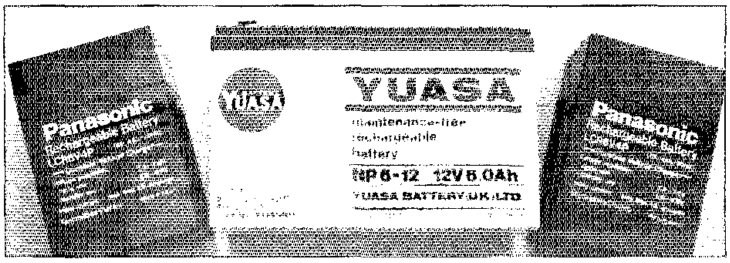
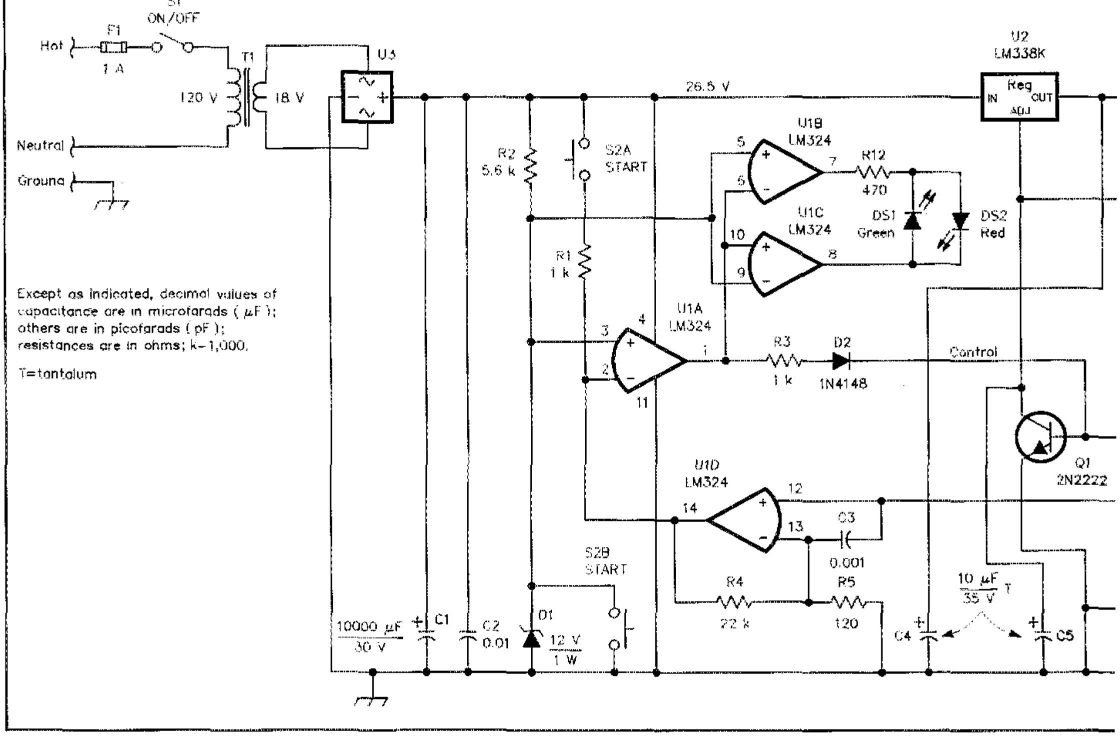


Figure 1—Typical gelled sealed lead-acid batteries.

Figure 2 (below)—Schematic of the battery charger circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

D1—12-V, 1-W Zener diode.
DS1—Green LED.
DS2—Red LED.
F1—1-A, slow-blow.
Q1—2N2222, BC237, BC107.
S1—SPST toggle.
S2—DPST momentary push-button or toggle switch.
S3—SPDT toggle.
T1—120-V primary, 18-V, 5-A secondary.
U1—LM324 quad op amp.
U2—LM338K 5-A voltage regulator.
U3—50-V, 6-A bridge rectifier.

Misc: Single-sided PC board (see Note 2); heat-shrink sleeving; 2°C/W heat sink; TO-3 mounting kit; heat-conductive grease; panel-mount fuse holder; ac line cord; chassis grommet; binding posts (one red, one black); mounting hardware, 2.5-mm cable, hookup wire, enclosure.



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

T=tantalum

to 100% capacity for two reasons: (1) As the battery terminal voltage increases, the current-charge rate decreases asymptotically to a very low level that never actually reaches zero. Waiting progressively longer and longer for less and less energy transfer doesn't make sense; (2) it's far easier to determine a cutoff point at the 10%-current charge rate than at, say, the 1%-current charge rate.

Construction

To simplify construction, a single-sided PC board is available (see Note 2), although you certainly can choose another construction method. If you use the PC board, install all the components except D2. Instead, solder two short pins (use flea clips, pins pulled from an IC socket, or short lengths of wire) to the PC-board holes in place of D2. This enables the board to be fitted to the case and partially tested, after which D2 can easily be soldered to the existing pins.

S1, S2, S3, the battery charging terminals +B and –B, and LEDs DS1 and DS2, are mounted on the front panel as shown in Figure 3. (In the prototype shown, I used a single bicolor LED instead of individual LEDs). The ac line cord, F1's holder, the heat sink and U2 are mounted on the case rear as shown in Figure 4. Note: U2 *must* be isolated from the heat sink by an insulator.

Route the ac mains wiring and dc wiring at opposite ends of the enclosure. Use heat-shrink tubing to cover all mains wiring to reduce the risk of electric shock. A prototype of the battery charger is shown in the title-page photo. Having ready access to them, I used a toroidal transformer.

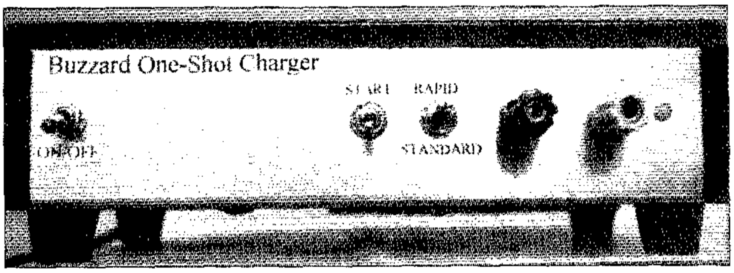


Figure 3—The simple front-panel arrangement of the prototype built in a low-profile cabinet.

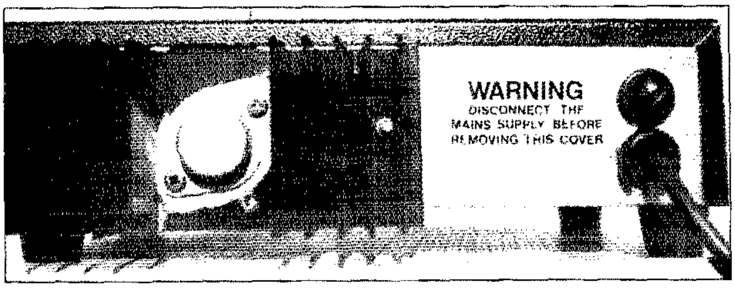
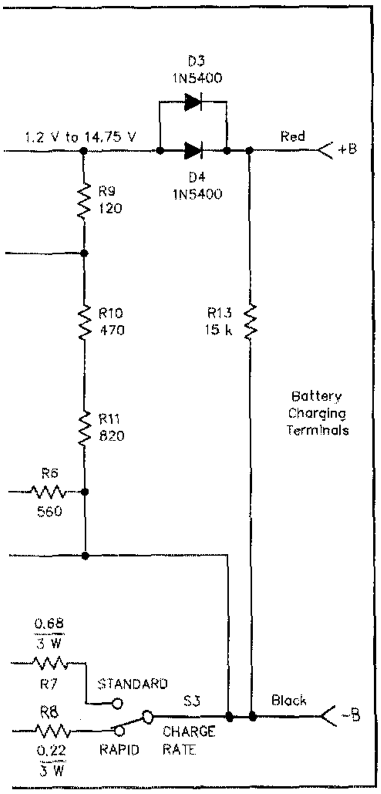


Figure 4—U2's heat sink occupies a good portion of the battery charger's rear panel. Don't forget to use heat-sink grease when mounting U2 to the heat sink.



Testing the Charger

Charger testing requires only a voltmeter and an ammeter. Apply line voltage and set S1 to the **ON** position. Check to see that the green LED (DS1) is lit and that the potential across the battery charging terminals is approximately 14.7 V. Remove the ac line cord and allow the output voltage to decay fully. Then solder D2 across the PC pins in the PC board.

Once again, apply line voltage. Check that DS1 is lit and that the potential across the battery charging terminals is approximately 0.6 V. Operate and hold the **START** switch (S2) and check that there's approximately 14.7 V across the battery charging terminals; DS1 will stay lit at this stage.

Note: *In the next two tests, don't short the battery charging terminals for longer than necessary!*

Set the **CHARGE RATE** switch (S3) to **STANDARD.** Connect an ammeter across the battery charging terminals. Check that the charger immediately goes into the on-charge mode, the red LED (DS2) should be lit and a current of about 1 A should flow. Then remove the ac line voltage and the ammeter.

Set S3 to **RAPID** charging. Connect an ammeter across the battery charging terminals. Check that the charger immediately goes into the **ON CHARGE** mode. DS2 should light and about 3 A should flow. Remove the line voltage and the ammeter.

Now the acid test. Apply the line voltage, select the appropriate charge rate and turn on the charger. Connect a partially discharged battery to the battery charging terminals. DS1 should still be lit. Operate the **START** switch and check that DS1 extinguishes and that DS2 lights.

After awhile, DS1 should light indicating that the charging cycle is complete and that the battery is now 90% charged and ready for use.

RF Compatibility—A Cautionary Note

I found my prototype battery charger to be susceptible to high localized RF fields, specifically when a battery was connected to the charger with two long leads (38-inches; about ¹⁄₂ λ at 2 m) and placed within a meter or so of a 25-W, 2-m transmitter's antenna. The behavior of the charger was unpredictable, flicking between the on-charge and off-charge states. The addition of C3 eliminated the problem.

Acknowledgments

Thanks are due to Andrew Clarke of DMS Technologies (Gates Energy Products manufacturers of the Cyclon sealed lead-acid batteries) for his help in the early stages of this project.

Notes

[1] RAYNET (Radio Amateurs emergencY NETwork) is the UK equivalent of RACES.

[2] PC boards are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price: $4.50, plus $1.50 shipping. A PC-board template package is available from the ARRL. Address your request for the SPENCER BATTERY CHARGER TEMPLATE to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Enclose a business-size SASE. QST

The Pfeiffer Quad Antenna System

Here's how you can shrink a standard quad and wind up with a no-compromise antenna. Build one and prove it to yourself!

By Andy Pfeiffer, K1KLO
132 Whippoorwill Rd
Old Lyme, CT 06371

Over the years, I built many Yagi antennas and, until recently, had three homemade monoband Yagis in service. I had avoided experimenting with quads after seeing the sad remains of one visited by a New England wind and ice storm. (Here in Old Lyme, at the mouth of the Connecticut River, we have our share of winter storms and hurricanes.) Although the quad is an excellent performer, to many hams it's a three-dimensional nightmare. It's physically difficult to maneuver and *prone to destruction when subjected to the* violent forces of nature. No more.

Design History

In April 1991, I dismantled my home-brew, 12-meter, three-element monoband Yagi and decided I'd attempt to shrink the quad so it would no longer be hurricane bait. The quads I'm about to describe are small and mechanically rugged. They'll weather anything the elements can hurl against them—and do so without any noticeable loss in performance. Therefore, I've permanently retired my Yagis!

My 12-meter Maltese Quad (so named for its resemblance to the Maltese cross) has been in constant service for over two years. A second Maltese Quad, built and erected about one year ago, replaced a *home-brew, three-element, 17-meter* monoband Yagi, and is as effective as its 12-meter forerunner. Being unable to let go of the tiger's tail, I designed and built a *third* Maltese Quad (this one for 20 meters) modified such that it *further* reduces the size of the antenna, compared to a standard quad frame.

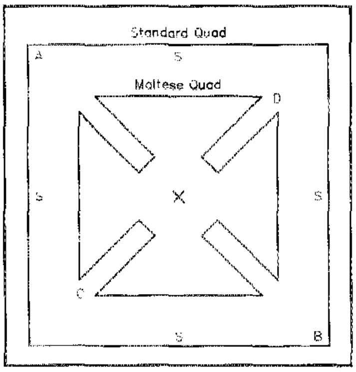


Figure 1—The outer square is the wire perimeter of a standard-size-quad driven element for the 12-meter band. Using the formula $250 \div f_{MHz}$ for one side of the square, and a center frequency of 24.940 MHz, each side is 10 feet long. This translates to a spreader diagonal *(A-B) of 14 feet 2 inches. The inner* configuration defines the wire perimeter of the Maltese Quad. It's drawn to the same scale, but has a spreader diagonal (C-D) of only 8 feet. This indicates the respectable size reduction obtained by the linear-loading approach. (The spreaders have been omitted from this drawing to maintain clarity.)

Design Approach

It's not difficult to design and build an antenna physically smaller than normal: Loading coils and traps can do the trick. Using them, however, introduces losses. To me, that approach is a waste of time and effort. I don't want a compromise antenna. In lieu of coils and capacitors, I chose to use linear loading.

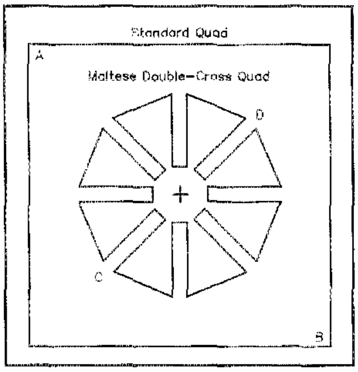


Figure 2—The perimeter wiring of a standard 20-meter quad driven element. The inner configuration (that of the Maltese Double-Cross Quad) is of a wheel with eight spokes (eight spreaders). Its spreader diagonal, C-D, is a mere 10 feet 4 inches compared to the 25-foot diagonal of the standard 20-meter quad!

Standard Maltese Quad

During the following discussion, refer to Figures 1 to 5. In Figure 1, the outer square represents the wire perimeter of the driven element of a 12-meter standard-size quad driven element. Using the formula $250 \div f_{MHz}$ for one side of the square, and a *center frequency of 24.940 MHz, each side* (S) is 10 feet long. This equates to a spreader diagonal (the distance from point A to point B) of 14 feet 2 inches. The Maltese Quad (inner drawing) has a spreader diagonal length (point C to point D) of only 8 feet! (The spreaders have been omitted from Figure 1 to maintain drawing clarity.)

For a standard 20-meter quad and a center frequency of 14.175 MHz, each side in Figure 1 is about 18 feet long. This translates to a spreader diagonal, A to B, of 25 feet 5 inches. By comparison, the Maltese Quad has a spreader diagonal of only 14 feet! This clearly shows the considerable size reduction obtained by using linear loading.

A Double-Cross

The perimeter of a standard 20-meter

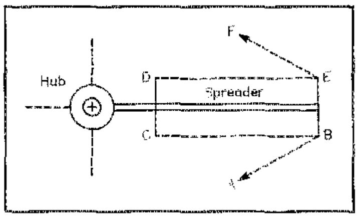


Figure 3—Here's a single spreader with the two yardarms in position. The dashed lines, A-F, indicate the course of the Maltese-Quad perimeter wire. E-B is the fiberglass rod; C-D is the aluminum rod. (The mechanical layout and the wiring of the reflector element is identical to that of the driven element.)

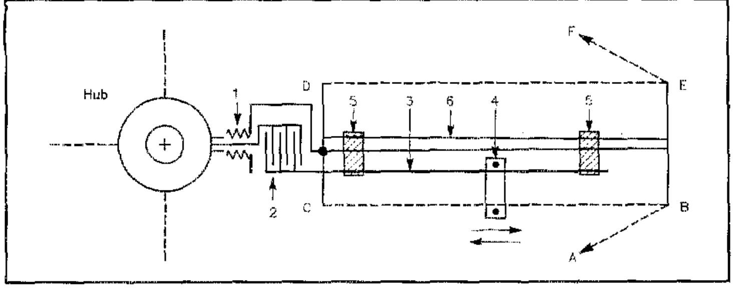


Figure 4—Drawing of the gamma match used with all of the Maltese Quads. Although some of the components used in the gamma match may seem esoteric, they are materials I have on hand. Other Maltese Quad builders have substituted other materials without detrimental effect. (1) is an SO-239 connector for attachment of the coaxial-cable feed line; (2) is an air-variable capacitor; (3) is the gamma rod, made of a length of 1/8-inch-diameter copper wire; (4) is the match's adjustable bar, also made of copper; (5) are supports for the gamma rod (made of Lexan polycarbonate); (6) is the fiberglass spreader arm of the quad. The gamma match is adjusted by tuning the variable capacitor and moving the adjusting bar (4), which slides over the gamma rod and bottom wire (C-B) of the linear-loading section. Refer to *The ARRL Antenna Book* or *The ARRL Handbook* for more information on the gamma match and its adjustment.

quad element is shown in Figure 2. The inner drawing (that of the Maltese Double-Cross Quad) is of a wheel with eight spokes (eight spreaders). Its spreader diagonal, C-D, is only 10 feet 4 inches compared to the standard quad's diagonal (A-B) of 25 feet. A quite respectable difference!

Construction

Because of my former occupation, I have a junk box of materials that most hams do not, although the materials can be purchased. I also have a machine shop. Consequently, this isn't a step-by-step construction project with a detailed parts list. Nevertheless, the Maltese Quads have been duplicated by hams in other parts of the world. In fact, I know that a few have been built in England. These hams used tools and materials at their disposal, so put your ham ingenuity to work.[1, 2]

Solid fiberglass fishing-rod blanks,[3] fiberglass tubing (or well-varnished bamboo, if you can find it) can be used for the spreaders. If you use fiberglass tubing, don't drill any holes along their length and thereby weaken them. Also, be sure to seal the tubing at both ends. The mechanical layout and the wiring of the reflector element is identical to that of the driven element, but the reflector is larger, as I'll explain later.

I'll refer to a single quad element as a "four-spoke wheel" (four spreaders) attached to a center hub (the spider). The rim of this four-spoke wheel is the wire perimeter.

The Linear-Loading Sections

Refer to Figures 3 through 5. Because linear loading is used (see Figure 3), the mechanical configuration of the Maltese Quad and its radiator wiring differ from that of the standard quad.

Clamped to the tip of each Maltese-Quad spreader is a short insulating arm (E-B), forming a T with the spreader. I refer to these short pieces as the "outer yardarms." About one foot from the hub end of each spreader is an "inner yardarm" (C-D). The inner and outer yardarms, and the wires running between them (E-D and B-C), form the linear-loading sections, with the inner conductive yardarm (C-D) being part of the radiating element.

Figure 3 shows how the lengths of the inner yardarms, C-D, and the outer yardarms, B-E, determines the spacing between the parallel wires, C-B and D-E. Changing this spacing alters the element's resonance. Changing the proximity of points C and D to the opposite points on the adjacent inner yardarm has the same effect.

The ends of the inner and outer yardarms have tie points to receive the perimeter wire and the wires of the loading sections. The dashed lines, A-F, indicate the course of the quad perimeter wire.

The outer yardarms on my quads are lengths of 3/16-inch-diameter solid fiberglass. At points B and E (Figure 3), the ends of the outer yardarms, stainless-steel, threaded, #8-32 studs receive the perimeter wire and the wires coming from the inner yardarms.

Each of the inner yardarms is a short length of 1/4-inch-diameter solid-aluminum rod equal in length to the outer yardarm. At points C and D, the inner yardarms have stainless-steel #6-32 machine screws, to which are attached the linear-loading wires coming from the outer yardarm.

Figure 3 shows a single spreader with its two yardarms in position. The lines A-F indicate the course of the Maltese Quad wiring perimeter. E-B is the fiberglass rod. C-D is the aluminum rod. The rod has an opening at its center to which a half-turn wire loop is temporarily attached. A dip-meter sensing coil is inserted into this loop during adjustment of the perimeter loop to the desired frequency. After the adjustment has been made, the opening is permanently shorted and becomes the attachment point for the gamma-match system. See Figure 4. (In actuality, the single spreader shown in Figures 3 and 4 assumes a vertical position—pointing down—in the finished quad.)

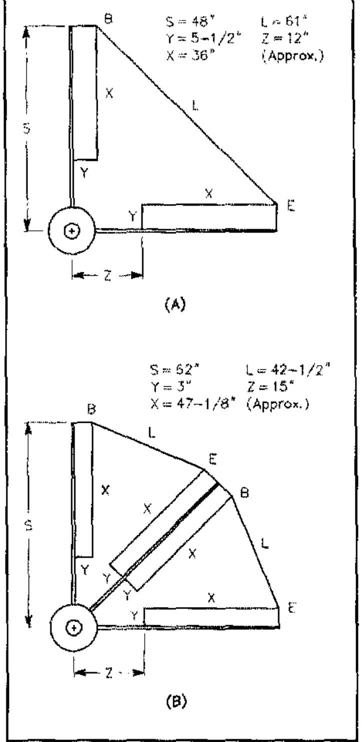


Figure 5—Dimensions of the various sections of the 12-meter Maltese Quad (A), and the 20-meter Maltese Double-Cross Quad (B). Points B and E are for wire attachment (see text).

Don't attempt to use a single piece of wire for the perimeter wire and the wires in the linear-loading sections! Use one wire from B to C, and another from D to E. The wire identified as L between points E and B in Figures 5A and 5B should be a separate, single piece of wire.

Tune-Up

I used a dip meter in conjunction with a calibrated receiver to adjust each element to resonance. Whatever method you use, be sure to separately adjust each element to resonance. Then, assemble the driven and reflector elements on the boom. Once at-

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[1] Notes appear on page 31.

tached to the boom, the individual element resonant frequencies will change because of their mutual interaction. This can be rectified once the antenna is matched to its feed line, and you'll be able to check the resonant frequency of the driven element as indicated by the lowest SWR point. You can then adjust the driven-element perimeter length—shorten it, or lengthen it—to raise or lower its resonant frequency. The reflector element perimeter can also be adjusted to increase forward gain, or front-to-back ratio, as you prefer. (I used this system in the frequency adjustments of all three Maltese Quads.)

When you get to the final perimeter adjustment, in order to achieve desired resonance (again, see Figure 3), move one inner yardarm away from, or toward, the hub center, removing or adding wire as required. I use the inner yardarm that is in line with the feed-point inner yardarm.

Figure 5A shows one quadrant of the Maltese Quad. The linear loading method for the 12-meter Maltese Quad requires about 17.5% more wire (7 feet) than a standard 12-meter quad. (I used 16-gauge [0.050-inch-diameter] copper wire.) The dimensions are those I used for the driven element of my 12-meter Maltese Quad, with a center frequency of 24.940 MHz. The perimeter of the driven element is 48 feet. The total perimeter for the driven element of a standard 12-meter quad is 40 feet 1 inch.

A standard 20-meter quad and a 20-meter Maltese Double-Cross Quad on 8-foot booms, using the diamond orientation, have turning radii of 13 feet and 6 feet 7 inches, respectively.

I mounted my Maltese Quad and Maltese Double-Cross Quad in the diamond position. This favors the height of the quad above ground, and there is less tendency for ice build-up on the wire perimeter. In addition, this orientation provides an ideal site for the feed point and its associated matching network.

Going Down?

I've concluded that a Maltese Double-Cross Quad for 40 meters is feasible—and a tempting project! A standard 40-meter-quad element requires a spreader diagonal of about 49 feet, whereas a Maltese Double-Cross Quad needs a spreader diagonal of *only 19 feet!*

Going Up?

Applying the Maltese Quad linear-loading system for 2, 6, and 10-meter quads—using aluminum tubing rather than copper wire for the radiator—would provide a self-supporting, rigid quad frame. You'd need a plastic hub (spider), but no spreaders would be required (see Figure 6).

Empirical Conclusions

Not having access to an antenna range, the conclusions regarding the performance of a Maltese Quad versus that of a standard quad have been made through on-the-air contacts. One of the most valuable series of these tests was that with Joseph J. Belson, K2ANR, of Riverhead, Long Island, New York, whose location is some 60 miles from mine. Through the years, both of us have realized the importance of a superior antenna system, and we both "graduated" (changed our direction of antenna experimenting from Yagis to quads) at the same time, except that Joe uses a standard quad.

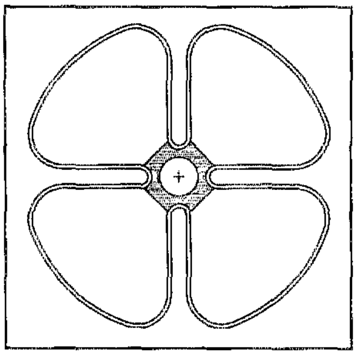
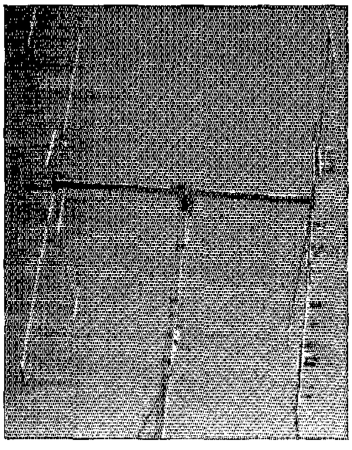


Figure 6—Front view of a quad element for the 2, 6, or 10-meter bands applying the Maltese Quad linear-loading system. Rather than copper wire, aluminum tubing is used for each radiator in conjunction with a plastic hub (spider). This would provide a self-supporting, rigid quad frame without the need for spreaders.



PETER BOWERS, G0HES

Figure 7—Peter (G0HES) Bowers' version of the Andy (K1KLO) Pfeiffer compact linearly loaded Maltese Quad for the 12-meter band. The spreaders are solid 10-mm fiberglass rod, as are the outer yardarms. The inner yardarms are 10-mm aluminum tubing. High-density plastic forms the hub and the boom is made of 44-mm-OD fiberglass tube. For the perimeter wire, 1.25-mm-diameter enameled copper is used.

Over a two-year period, Joe and I conducted many three-way QSO comparison tests on the 12- and 17-meter bands under all conditions covering all continents. We've concluded that the two quads perform equally. That is, front-to-back ratio, forward gain and side rejection are the same. (Joe's quads are a good deal higher than mine, but the height difference didn't seem to matter. The lowest point of my quads is about one-half wavelength or less above ground. Joe and I both ran about 100 W during these experiments.

My 2-element, 20-meter monoband Maltese Double-Cross Quad on an 8-foot boom has been in service since March 1993. It's made a good account of itself, signal reports being consistent with those received from the 12- and 17-meter Maltese Quads on their respective bands.

It's a real pleasure for one person to be able to assemble a 2-element 20-meter quad, attach its boom and mast, and then easily carry the complete assembly to its location!

I use a gamma match on all my quads; I feel comfortable with this system. The 2:1-SWR bandwidth of the 12- and 17-meter Maltese Quads is in excess of 500 kHz. The 20-meter Maltese Double-Cross Quad presents no problem to my TS-180S on frequencies from 14.0 to 14.350 MHz.

After hundreds of contacts with the Maltese Quads, I can say without hesitation that my system has not degraded the bandwidth or the other favorable characteristics of the standard quad. The standard quad is a very low-Q antenna—most forgiving; the Maltese has not altered these important qualities.

An Acknowledgment

As with my past antenna experiments, I'd not have been able to see this project through to its conclusion without the invaluable aid of my wife, Marianne, who through the years has weathered the brainstorms of K1KLO.

Experimentation

For those of you who'd like to experiment with the linear-loaded quad configuration, here are some notes, hints and thoughts.

In determining the dimensions for a 2-element monoband quad, I used the formulas found in Bill Orr's cubical quad handbook.*

Total driven element perimeter:
$1000 \div f_{MHz}$ (Eq 1)

Total *reflector element perimeter:*
$1032 \div f_{MHz}$ (Eq 2)

Element spacing: $118 \div f_{MHz}$. (Eq 3)

I've not developed formulas that determine the relationship between the driven and reflector elements for the Maltese Quads. Therefore, I translate the length of one side of the standard reflector element to the frequency

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*W. Orr, *All About Cubical Quad Antennas* (Wilton, CT: Radio Publications Inc, 1971), 2nd edition, 2nd printing.

that this length establishes. For example, in selecting a frequency of 14.2 MHz for a standard 20-meter quad, one side of the reflector is equal to 18.169 feet. This dimension, applied to the formula 250 ÷ 18.169 feet, results in a frequency of 13.759 MHz. I then adjust the reflector element to that frequency.

During my first experiments for this project, I used an element with four 8-foot-long spreaders. The four inner yardarms were fixed 12 inches from the hub center. The four outer yardarms were adjustable at 6-inch intervals from the spreader tips down to 3 feet from the hub center. Then I wired up the element perimeter with the outer yardarms set at the 8-foot locations and measured and recorded the element's resonant frequency. Next, I moved the four outer yardarms 6 inches toward the hub center and rewired the element. I checked this new element's resonance and recorded it, repeating the process down to the 3-foot marks. This data is valuable in planning Maltese Quads for other bands.

I applied this same testing technique in experiments with the Maltese Double-Cross Quad, with the following two changes: The eight inner yardarms were fixed at 15 inches from the hub center, and all yardarms, inner and outer, were reduced in length to 6 inches (see Figure 4B).

When conducting Maltese Quad experiments, be aware that the diameter of the perimeter wire used for the finished quad *must be the same as that used in the experimental antenna.* For example, if you use 20-gauge wire (0.032-inch diameter) in your experiments, and use 14-gauge wire (0.064-inch diameter) in your finished quad (using the same total length of wire), you'll find that the finished element's resonant frequency is about 1 MHz higher.

Until his retirement in 1986, Andrew (Andy) Pfeifler, K1KLO, was a self-employed medical research instrumentation consultant in the field of neurophysiological studies. Through the years, his consuming interests in Amateur Radio have been antenna experimentation, design and construction. A few of his other interests are hiking in New Hampshire's White Mountains with his wife, Marianne, and amateur archeology and archery. Andy is unable to answer written inquiries, but can be reached by phone at 203-434-5621.

New Books

THE BEGINNER'S HANDBOOK OF AMATEUR RADIO

By Clay Laster, W5ZPV

TAB Books, McGraw-Hill Inc, Blue Ridge Summit, PA 17294-0850; tel 717-794-2191, fax 717-794-2103. Third edition, March 1994. Paperback, 416 pp, 210 illus, 7×10 inches, retail price $21.95. ISBN 0-8306-4354-0.

Reviewed By Brian Battles, WS1O
QST *Features Editor*

It's a nonham's introduction to Amateur Radio! It's a license manual! No, it's both! Well, actually, it's not quite either.

It can be frustrating for a seasoned ham to try to entice someone else to join him or her in the hobby because the reasons for wanting to be a ham are notoriously intangible. What a job it is to make Amateur Radio clear and enticing to an outsider. Telling someone how much fun it is to make two-way radio contacts with random strangers can make Amateur Radio appear downright pointless or frivolous to nonhams. In a one-on-one, face-to-face conversation, this can be difficult; in print it's an enormous challenge at best. (So far, the only book I've ever read that pulls this task off well is *Ham Radio Horizons: The Book* by Peter O'Dell, WB2D; see New Books, November 1993 *QST*, page 49).

You're not going to buy a copy of this book for Aunt Flo to transform her from someone who thinks there's lightning inside the wall sockets to a devoted ham radio enthusiast. This isn't an invitation to the hobby for nonhams, it's not positioned as a license manual and it's not a comprehensive reference to electronic communication like *The ARRL Handbook.* Many experienced hams can't imagine how little the average person knows about basic electricity, let alone radio electronics, and it takes a deft touch to introduce it without frightening a beginner with complicated lessons and details. Because there are excellent texts available that focus on license exam preparation, it can be argued that another one isn't necessary.

Having stated what this book *isn't*, can we determine what it *is*? This is where it's confusing: Laster's book is a bit advanced to be an introduction to Amateur Radio for people who know little about the hobby. On the other hand, it's sort of a preparation manual for the entry level FCC licenses. When Laster explains what ham radio is, he does a good job of making it sound like a fun and intriguing pastime. Instead of concentrating on that theme, however, he segues right into FCC Rules, basic electronics theory, equipment and antenna topics, and operating details beyond the typical needs of the person who's a true beginner in Amateur Radio. In some ways, the technical information he discusses isn't quite complete enough to recommend this as an in-depth license-preparation manual, and in some areas it takes you considerably beyond the material covered in the exam question pools. Laster goes a few steps toward serving as an Elmer who can "clue in a new guy," and he often refers to the ARRL and what it has to offer new and potential hams—in several places he recommends that readers take advantage of League information, resources and services.

So what *is* the *Beginner's Handbook of Amateur Radio*? I'm not sure, and I don't think Laster is, either. It appears that what may have begun as a means to express his enthusiasm for explaining everything to newcomers went beyond the scope of what a prospective or beginning ham may want to read about—in fact, it's not hard to imagine a reader starting out with a sense of excitement, but becoming slowly turned off by the technical material. It's not poorly written, there are plenty of interesting and useful photographs and illustrations, and some worthwhile tutorial and reference material for hams, but it needs a slightly tighter focus. As it stands, it seems best suited as a nice supplemental resource for someone who's studying for a ticket or has just received a new license. Perhaps retitling it *The Beginning Ham's Handbook of Amateur Radio* would be a step in the right direction.

ENCYCLOPEDIA OF ELECTRONIC CIRCUITS, VOL 4

By Rudolf F. Graf and William Sheets

TAB Books, McGraw-Hill Inc, Blue Ridge Summit, PA 17294-0850; tel 717-794-2191, fax 717-794-2103. First edition, first printing, 1992. 650 pp, B&W schematic diagrams. 9¼×7¼ inches. Retail $29.95. ISBN 0-8306-3895-4.

Reviewed By Bruce Hale, KB1MW/7
2238 168th Ave NE
Bellevue, WA 98008

The Encyclopedia of Electronic Circuits is a hefty 650-page collection of circuits pulled from a variety of electronics magazines, including *QST*, *73 Amateur Radio Today*, *Popular Electronics* and *Elektor Electronics*. Each circuit is presented with only a brief description and average one circuit per printed page—as many as three circuits are presented on some pages; other circuits require two pages.

The circuits are an interesting collection, ranging from audio mixers through power supplies and battery chargers to FM wireless microphones and model-train controllers. Some are simple and easy to build, while others are practically impossible without referencing the original article (some descriptive text is missing or the parts are difficult to find).

Graf and Sheets could have been more selective in choosing circuits. I haven't seen volumes 1-3—perhaps the well has run a bit dry and those volumes are more useful. I did find some interesting circuits—I built a NiCd battery charger and there are a few other circuits I'd like to try. But at $29.95, I can't recommend that you buy this book without taking a good look at it first.

Notes

[1] For some additional information on standard quad construction, see W. Stein, "A Five-Band, Two-Element Quad for 20 through 10 Meters," *QST*, Apr 1992, pp 52-56; K. Wellenius and Björn Wellenius, "A Light and Sturdy Quad for 10 and 15 Meters," *QST*, Jul 1993, pp 30-32, and *The Radio Amateur's Handbook*.

[2] I recently had the distinct pleasure of working Maltese Quad to Maltese Quad! *That* was an experience I'll never forget! Peter Bowers, G0HES, was kind enough to send me photos of his version of the Maltese quad. (See Figure 7.)

[3] Fiberglass rod, tubing and square-sided tubing is available from Max-Gain Systems Inc, 221 Green Crest Ct, Marietta, GA 30068, tel 404-973-6251. Solid and tubular fiberglass fishing-rod blanks are available from: Netcraft Co, 2800 Tremainsville Rd, Toledo, OH 43613; for orders, tel 800-638-7238; customer service, 419-472-9826.

Using a VU Meter for Phone-Patch Adjustment

Fiddling with a patch's phone-line output until it "sounds okay" just isn't good enough. Here's how to use the right tool for this audio-measurement job.

By Alfred Lorona, W6WQC
1810 Muirfield Dr
Oxnard, CA 93030

Recently, as a fellow amateur and I installed and adjusted a telephone patch, we discovered that current Amateur Radio references do not describe a method for adjusting phone patches for proper telephone-line drive on speech signals. This article fills that gap by discussing how to adjust a patch for speech with the help of a VU (volume unit) meter.[1]

VU Meter Basics

You can easily measure the level of a continuous audio tone with just about any meter capable of accurate indication if the signal's strength and frequency stay the same. You just hook your meter to the tone source, wait for its pointer to settle down at a constant indication, and take the reading.

Measuring speech-audio levels is comparatively tricky, however, because speech varies continuously in frequency and amplitude. Connected to such an audio source, a meter keeps moving and never settles down to a steady reading.

A meter's *ballistics*—how its mechanism acts when in motion—play a major role in how it indicates changing speech and music levels. Different meters respond to sudden level changes differently: Some change indication almost instantly, but greatly overshoot the proper indication, coming to rest only after a frustratingly long period of diminishing overshoot/undershoot cycles. Some meters take longer to respond to level changes, but settle down to a steady reading relatively rapidly, with little overshoot.

Telephone company specifications call for driving a phone line with no more than –9 dBm (about 0.13 mW) of audio. Using a meter to adjust a phone patch for proper phone-line drive therefore requires a meter capable of providing acceptably accurate readings on voice signals at powers in this range. That's where the VU meter comes in.

First adopted in 1939 as an industry standard, the VU meter resulted from a joint effort by the Bell Telephone Laboratories, the Columbia Broadcasting System and the National Broadcasting Company to develop an audio level meter with standardized ballistics. The goal was a meter that would indicate complex speech waveforms in a way that would correspond closely with a listener's subjective impression of the signal's perceived loudness.

This was not easy to do. Although meters can be made to register the average power or the instantaneous power peaks of the signal, there is no simple relationship between these two qualities of a complex speech waveform and its loudness. Nevertheless, by making the meter indicate a value somewhere in between average and peak power, it was possible to provide a close correlation between the meter indication and loudness.

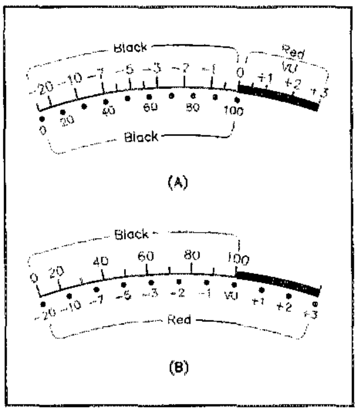

Figure 1—Both VU-meter scales span the same range of signal-strength change. The A scale is of most use for audio-level adjustments because it emphasizes its decibel calibration. The B scale *emphasizes percentage of transmitter* modulation. VU meters are often available at flea markets, swap meets and surplus outlets, such as Fair Radio Sales (address in Table 42 of Chapter 35 in the 1993 *ARRL Handbook*).

You've probably seen audio level meters marked *VU* in consumer audio, recording and public-address equipment. But not every audio meter marked *VU* meets the specifications for a true VU meter.[2]

VU Meter Characteristics

A true VU meter consists of a 200-µA, 3.9-kΩ coil d'Arsonval movement equipped with a copper-oxide full-wave bridge rectifier. (The rectifier is required to transform ac voltages into dc voltages because the basic meter movement responds only to dc.) The meter is intended to be used in series with a 3.6-kΩ resistor for connection across 600-Ω audio lines (the standard impedance for phone lines, and long-distance audio transmission in the broadcasting and recording industries). This results in a total instrument impedance of 7.5 kΩ—a value that, when *bridged* across a 600-Ω line, causes a level reduction of only 0.4 dB.

VU Meter Scales

Because they were developed for two main uses—(1) broadcasting and (2) recording, test equipment and telephone applications—VU meters are available in A- and B-scale form. Both scales (see Figure 1) cover the same span of signal-strength change. The meter's 0 to 100 scale indicates percentage of transmitter modulation, its –20 to +3 dB scale indicates decibels relative to 0 VU (0 dB). (One *volume unit* equals a 1-dB level change in a complex waveform.)

The A scale emphasizes the –20 to +3 dB calibration and is preferred for recording and test-equipment applications; the B scale, which emphasizes the 0 to 100% calibration, is preferred for broadcast applications. Other than this difference in appearance, A- and B-scale VU meters are electrically and ballistically identical and can be used interchangeably.

Sensitivity

Connected in series with a 3.6-kΩ resistor, a VU meter indicates 0 VU when driven by a sinusoidal, 1-kHz, 1.228-V

[1] Notes appear on page 33.

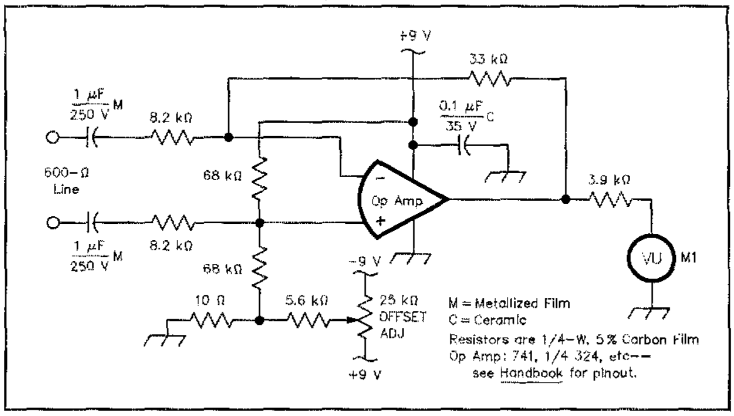

Figure 2—How to wire a general-purpose op amp to provide the 13 dB of gain necessary for a VU-meter indication of 0 VU when a −9 dBm signal is present. (See text.) M1 is a standard (3.9-kΩ coil, 200-μA full scale) VU meter. There is nothing critical about constructing this circuit; the op amp can be a 741 or ¼ of a 324, for instance. Set the OFFSET ADJ trimmer so the meter reads zero with no signal present.

RMS (+4 dBm) signal. Without its 3.6-kΩ series resistor, the basic VU meter movement indicates 0 VU for a standard power level of 1 mW in 600 Ω (0.775 V RMS across 600 Ω).

With the series resistor in place (which brings the instrument's total resistance up to the 7.5-kΩ value called for by the VU meter standard), the meter indicates −4 dB with 1.228 V RMS applied. This 4-dB reduction is of no consequence if we remember to mentally add 4 dB to the pointer indication, and is a satisfactory arrangement for measuring speech levels in the 0 to +4 dBm range.

Ballistics

When a steady 1-kHz sine wave is suddenly applied to a VU meter, it indicates the signal's true value to 99% within 0.3 second and overswings the true value by not more than 1.5%. The required outboard 3.6-kΩ resistor plays an important part in maintaining these ballistics: The meter responds more slowly if less external resistance is used.

Making a VU Meter Cover the Range We Want

The level to which a speech circuit should be adjusted depends on two conflicting requirements. The level should be high enough to provide an acceptable signal-to-noise ratio at the receiving end, but not more than the telephone system allows. This means that the signal should be substantially stronger than line hum and noise, but not high enough to cause crosstalk or overload. The line level of −9 dBm, maximum, complies with these requirements.

If we want our meter to indicate 0 VU when connected to a line operating at −9 dBm, amplification is necessary. Without amplification, the meter will indicate −13 dBm, equal to −9 dBm plus the 4-dB loss contributed by the meter's outboard 3.6-kΩ resistor. (Yes, a VU meter's scale already covers −13 dBm, but in a part of the scale where resolution is poor.) We therefore need 13 dB of amplification to make our meter indicate 0 VU at our desired signal level of −9 dBm. Figure 2 shows a circuit that can do this.

So far, I've talked about a VU meter with a 3.6-kΩ series resistor, but the 3.6-kΩ value assumes that the meter is bridged across a 600-Ω line terminated with a 600-Ω load—300 Ω with both values paralleled—to ultimately load the meter with 3.6 kΩ + 300 Ω, or 3.9 kΩ. Such is not the case in Figure 2 because the output impedance of an operational amplifier is a few tens of ohms at most. The meter series resistor in Figure 2, therefore provides all of the 3.9-kΩ resistance necessary to ensure the meter's proper ballistic response.

In addition to providing 13 dB of gain, the preamplifier circuit must lightly load both sides of the telephone line, neither of which is grounded. This is accomplished by taking advantage of the op amp's differential input feature. Sharp-eyed readers will note that the resistive loading on the lines is unequal, but this is of no consequence because both resistances are so much higher than 600 Ω that their difference contributes insignificant imbalance.

Connect the meter amplifier's input to the phone line at your phone patch with your phone off the hook or while the patch is disconnected from the phone line. Treat the phone-line wires as if they carry high voltage—which they do, especially when ringing signals arrive (nearly 100 V RMS at the customer end of a phone line).

Using the Meter

Now comes the most difficult part: learning how to read and interpret the meter indication. The correct way to read the meter is to note the average indication of the three highest peaks during a 10-second period while disregarding occasional extreme peaks. This is not as formidable as it sounds. Just recite the following two sentences twice in succession in a normal speaking voice while observing the meter pointer's peak swings: "Joe took father's shoe bench out" and "She was waiting at my lawn." These two sentences contain all of the fundamental sounds of the English language and, when spoken twice, take about 10 seconds to say. (Try it!)

Once you're sure of how to read your VU meter, you can readily adjust your patch's output for −9 dBm of phone-line drive. That's all there is to it. Your phone patch is now optimally adjusted.

Notes

[1] Pages 28-13 through 28-16 of the 1993 *ARRL Handbook* describe how to use an audio tone in calibrating a nonstandard VU meter to indicate 0 dBm at 600 Ω. (For suggestions on how to modify the response of nonstandard VU meters that respond too quickly to level changes, see G. Schleicher, "Measuring Phone-Patch Levels Accurately," *QST*, Feb 1972, pp 24-26.) The *Handbook*'s phone-patch coverage, which I recommend as an overview of phone-patch and phone-line interface topics, includes a statement that applies here: "You are advised to obtain a copy of FCC Part 68 from the Government Printing Office, and read Sub-Part D before connecting any home-made device to the telephone line."

[2] ANSI/IEEE 152-1953, abbreviated title: *Volume Measurements of Electrical Speech and Program Waves*.

Alfred Lorona has been licensed since 1941 and holds an Extra Class license. His main interests are QRP CW ragchewing and home-brewing.

New Products

ROOF-MOUNT TOWERS

◊ No room for a full-size tower on your property? Put a tower on the roof. The manufacturer of the Hazer antenna tram system has added two roof towers to its line of aluminum antenna-support products. The premier model is the RT-936, a nine-foot, four-leg tower capable of mounting antennas up to 28 square feet of wind load (85 mi/h). It weighs 78 pounds and retails for $378.75. The lighter model RT-832 is an eight-foot, four-leg tower that supports wind loads up to 8 square feet. It weighs 37 pounds and retails for $189.95. Both roof-mount towers come with rotator mounting supports and top plates, and are constructed of 6061-T6 angle aluminum with stainless steel hardware. Gary Martin, WB0WMQ, Glen Martin Engineering, Rte 3, Box 322, Boonville, MO 65233; tel 816-882-2734; fax 816-882-7200.

On Center-Fed Multiband Dipoles

Is the G5RV really an all-band antenna?

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Since the opening of the 30, 17 and 12-meter bands, interest in the center-fed dipole with tuned feeders has seen a revival. Particularly popular is the G5RV,[1] a multiband center-fed dipole with a particular feed-line arrangement. This dipole can be used on the amateur bands 3.5 MHz and above. Its arms are somewhat shorter than a quarter wavelength on 80 meters, which makes it an attractive antenna for some, since it will fit on many city lots. Many amateurs regard this antenna with its so-called *special feed-line arrangement* as a panacea, particularly when it is used in a drooping dipole (sometimes called *inverted V*) configuration. There is, however, nothing magical or superior about the antenna:[2] It is merely a center-fed dipole with a particular feed-line arrangement, which the newcomer to Amateur Radio may or may not want to duplicate. In fact, the performance of a multiband drooping dipole can be inferior to a dipole at the same apex height. In preparing this article, we conducted a critical analysis of center-fed multiband dipoles, and developed data on calculated radiation patterns for horizontal and drooping configurations.

Center-Fed Dipoles with Tuned Feeders

The center-fed dipole with tuned feeders was a simple and widely used multiband antenna in the 1930s and 1940s, and various versions of it are still in use today. Because each half of the "flat top" is the same length, the feeder currents are balanced at all frequencies, except for any imbalance introduced because one half of the antenna is closer to the ground than the other. Antenna length is not particularly critical, nor is the feed-line length; however, some combinations allow easier impedance matching to the transmitter over a wide frequency range.

The dipole was generally fed with an open-wire transmission line. Antenna lengths of 135 feet and 70 feet were typically used, with feed-line lengths ranging from 40 to 75 feet. In the years before affordable coaxial cable, the entire transmitting and radiating system was balanced. Because modern transceivers and antenna tuners are unbalanced devices, coaxial cable is now the preferred feed line. Whatever the feed-line arrangement, a balun is required, which can lead to difficulties, as the load may be a very reactive mismatch at some frequencies.

G5RV Version

In 1946, Louis Varney, G5RV, anxious to get on the air after the war, designed and erected a multiband antenna which would fit his average-sized backyard. The antenna consists of a 102-foot flat-top, split in the center and fed by tuned feeders. Two versions were tested: one using full length open-wire feeders (Figure 1A) and the other using a 34-foot open-wire *stub* fed at its base by either transmitting-grade 72-Ω twin lead or 72-Ω coax (Figure 1B). The length of the stub was designed to be a half wavelength at 14 MHz (the reason is discussed later). An alternative to using an open-wire stub was also tested, using 300-Ω ribbon. In this case, the stub was shortened by the velocity factor for the ribbon. The stub version, Figure 1B, has become known as the *G5RV Antenna*. It was first described in a note by him in the July 1958 RSGB *Bulletin*.

With a suitable tuner, this antenna can be used on all HF bands from 3.5 to 30 MHz. Varney referred to this stub as a *matching section*. (The italicization of *matching section* and other words to follow in this para-

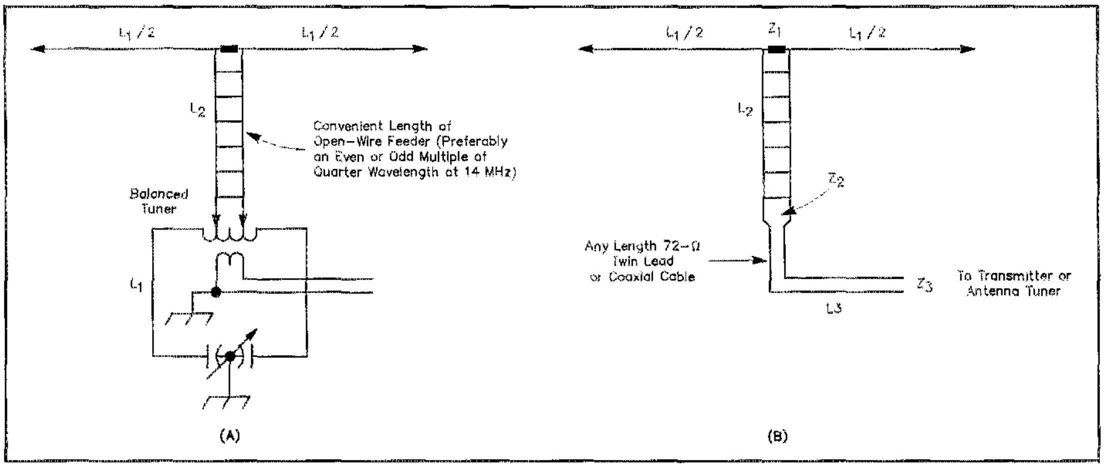

Figure 1—At A, a multiband dipole fed directly with open-wire line. The line is matched with a balanced tuner. At B, L3 is a length of 72-Ω twinlead or coaxial cable.

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[1] Notes appear on page 36.

graph is Varney's.) The statement *matching section* has led to considerable over-the-air discussion and confusion over the years. In retrospect, Varney concluded that it was probably a mistake to refer to this feeder arrangement as a *matching section* when it was only for the 20-meter band that it functions as one. (The dipole impedance of approximately 100 Ω at the center of the 3/2-λ flat-top was transformed by the λ/2 open-wire line (length, 34 feet) to the base of this section, thus providing an acceptable match for a suitable length of 72-Ω twin lead or coax *feeder* needed to reach the transmitter.[3]) More accurately, this *matching section* is a series-section impedance transformer. The *feeder* is also a series-section impedance transformer, a factor that in most cases is completely ignored. Varney recommended using open-wire tuned feeders to reduce losses due to the large-amplitude standing wave on this section of transmission line, and while he did not specify the impedance, the constructional details given correspond to a transmission-line impedance (Z_0) equal to 523 Ω.

Clearly, this feed-line arrangement is *special*, since multiband dipoles are more usually fed by means of a single feed line of constant characteristic impedance. The G5RV in effect uses a combination feed line: the open-wire line as described above in conjunction with the low-impedance *feeder*, which connects the antenna to the transmitter. This complicates discussion about the antenna. The confusion over the use of the words *special, matching section* and *feeder* persists today. Some antenna experimenters hold that the G5RV can be used without a tuner on all traditional amateur bands. It is curious how this idea came into the minds of the radio amateurs using this antenna, because Varney never said that it could be used without a tuner.

The G5RV antenna is typically not a perfect match on *any* amateur band, and since the SWR can be high on some bands, Varney recommended that no balun be used. This is another topic of controversy. A balun should be used to feed a balanced antenna by an unbalanced transmitter; however, the use of a balun can lead to difficulties when the SWR is too high. Without a balun there will be some feed-line radiation, which can be particularly severe on some frequency bands.

Dipole Orientation

Dipoles are often not installed in a horizontal straight line. Half-wave dipoles are generally tolerant of bending, sloping or drooping to fit the antenna site. But this is not always the case for multiband dipoles. We will consider here horizontal and drooping dipoles; we will not consider bent dipoles.

Louis Varney has always illustrated his multiband dipole mounted as a classical horizontal dipole. But radio amateurs frequently use multiband dipoles in a drooping configuration, since only one support mast is needed. In fact, this configuration is illustrated in *The ARRL Handbook*, which has for years given details for such a multiband dipole with 50-foot arms, fed by an open-wire transmission line. The caption under the sketch giving constructional details states that "the included angle (A) between the arms of the dipole for best efficiency should be between 90° to 110°."[4] The authors have not seen substantiating data to support this recommendation. In fact, as we see, the optimum angle for a multiband dipole is 180° (a horizontal dipole).

Measured Impedance Characteristics

The authors' G5RV (in spite of what was said previously) was installed as a drooping dipole. The included angle (A) between the arms of the dipole was about 127°; the apex height, about 30 feet. The dimensions for the authors' G5RV were those given by Louis Varney: L_1=102 ft; L_2=34 × 0.95=32.3 ft of 450-Ω transmission line. The factor 0.95 is the velocity factor for the 450-Ω ribbon. The antenna was resonant (reactance zero) on five frequencies in the 3- to 30-MHz range: 3.49, 7.52, 14.15, 19.5 and 24.6 MHz. It is interesting to note that the antenna was resonant in the 20-meter band, at 14.15 MHz, as G5RV intended. The antenna's resistance at these frequencies was 16, 31, 148, 48 and 162 Ω, respectively. This illustrates, in part, the problem in achieving a low SWR for harmonic-resonance frequencies of the antenna's impedance-versus-frequency response. For a 50-Ω feeder (see the following), the SWR at the corresponding resonant frequencies would be 3:1, 1.6, 3.0, 1.04 and 3.24.

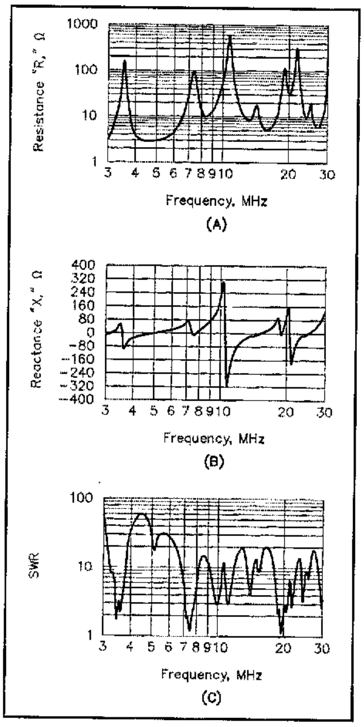


Figure 2—At A, input resistance versus frequency for the antenna tested by the authors; at B, input reactance versus frequency; at C, input SWR.

Our G5RV antenna was fed by an additional length of 50-Ω coax transmission line. We used a 31-foot length of mini-foam coax (the total length of which included the length of a 300-bead current balun; see below). The antenna-system impedance is changed by this feeder, depending on its length and characteristics (impedance and attenuation factor). In Figure 2A and B, the measured input resistance and reactance versus frequency is given for the antenna with a coaxial feeder and balun. Note in particular the anti-resonant responses in the 80-meter band, and just above the 40-meter band. The G5RV itself was resonant near these frequencies.

Figure 2C shows the SWR versus frequency. Although the SWR was low (less than 2:1) in three bands in the 3 to 30-MHz range, unfortunately it is high for most amateur bands. We should note that since a balun was used, there will be insignificant current on the outside surface of the feed-line coax, and therefore SWR should be independent of the length of the coax.

The G5RV antenna is clearly a multiband antenna. It is harmonically resonant on a number of frequencies in the 3 to 30-MHz range. Comparisons of measured SWRs we obtained with those reported by others showed low SWRs in different bands, though sometimes outside amateur bands.

Drooping versus Horizontal Dipoles

Drooping the arms of the λ/2 dipole has only a small effect on gain. When the length of the dipole is greater than λ/2, the place on the arms of the dipole where the current is a maximum is displaced from the center of the dipole. In this case, these current maxima occur at a lower height when the arms of the dipole are drooping. We might, therefore, expect that the effective height of the dipole would be decreased, and hence the launch angle increased. But the pattern and gain changes are more complicated than this. When the dipole is horizontal, the takeoff angle decreases continuously with increase in frequency (since the electrical height of the dipole increases with increase in frequency), and when the pattern becomes multi-lobed, the maximum takeoff value has the same value for all the lobes in the azimuthal plane. But this is not the case for the drooping dipole, and the differences become greater as the droop is increased.

A multiband dipole, if used at frequencies where its electrical length is appreciable (greater than λ/2) should be installed as a horizontal dipole, or as nearly horizontal as possible.

Conclusions and Recommendations

When reading about what has been written about the G5RV, or modified versions *based on the G5RV principle*, we must not think of the transmission-line segment L_2 as a *matching section*; more generally, it is a series-section transformer. This view is particularly important when considering the combined effect of the open-wire line and the coaxial feeder. Clearly the coaxial feeder of length L_1 is an add-on series-section impedance transformer, which may or may not include the impedance-transfer characteristics of the balun. The analysis approach (previously published) has considered only the impedance-transfer characteristics of the open-wire part, but has ignored the effect of the coaxial lead-in. This complication was avoided by considering only SWR rather than impedance. While there is nothing fundamentally wrong with this approach, in our view this confuses our understanding of the characteristics in the antenna system. Calling Figure 1A's antenna a dipole with tuned feeders, and Figure 1B's antenna a dipole with stub tuning, are also misnomers. Since an antenna tuner is used with the Figure 1B antenna, this antenna system is also a dipole with tuned feeders.

Despite several studies by those who have tried to optimize dimensions for the G5RV type of multiband dipole with its combination feeder arrangement, there is, in our opinion, little to be gained by this endeavor. Louis Varney, in correspondence with author Belrose said: "While I think that Brian Austin's ZS6BKW designs[5] of a multiband antenna are interesting, in view of the fact that whatever bands [on which] he manages to achieve an SWR less than 2:1, there will be other bands for which the SWR is greater than 2:1, and therefore the use of a suitable antenna tuner is essential. I would prefer more simply to use an antenna tuner." The authors agree with this view held. Attempts to optimize the antenna system are not worthwhile, except if operation on a particular band (or bands) for field use without a tuner is desired.

The length of the dipole depends on the frequency band of interest, and since the radiation pattern changes with frequency, the best length for the dipole as judged by a particular user should take this factor into account. The 102-foot length is a dimension that is approximately 3/8 λ long at the lowest frequency of operation, 3.5 MHz. This is a reasonable length. A dipole with shorter arm lengths quickly becomes impractical. The multi-lobed pattern for frequencies greater than 10 MHz is, in our view, undesirable. For communication in nonspecific directions a multiband dipole meets the requirements. If communication to a particular remote location is desired, however, it is important that the azimuthal pattern not have a null in the direction of interest. Dipole lengths greater than 1.25 λ generally do not meet this requirement. For operation on the higher bands, a scaled-down multiband dipole could be used, but other, quite different antenna types could be used instead. For example, a simple all-band antenna that avoids the difficulty of pattern change with frequency, for frequencies greater than 10 MHz, is a compact horizontal loop. Another simple antenna type, which the authors favor, is to use several dipoles in parallel (a so-called *fan* or *stagger-tuned* dipole).

Dipole orientation is rather unimportant at frequencies where the dipole arms are not appreciably greater than λ/2, whether the dipole is horizontal or drooping. When the dipole's arm length becomes greater than 1.25 λ, though, it should be mounted horizontally, or as horizontally as possible (Λ > 127°).

In our view, the correct feed-line length for a multiband dipole is that required to go from the output terminals of the antenna tuner to the antenna terminals, because—regardless of the length of the feed line—both the antenna and the feed line are made resonant by the tuner. Our recommendation, based on personal experience, is to use open-wire line for the *total length* of the required feeder. This will result in lower losses. An additional advantage in using a full-length feed line is that any necessary balun can be inside the station, easing evaluation of its performance, or you can use a balanced tuner.

It is interesting to note that the 100- to 150-Ω impedance of the 1.5-λ, 31.1-m-long dipole, when used for the 20-meter band, could be matched by employing two lengths of coaxial cable to form a balanced line. Varney designed his antenna to be resonant at 20 meters, so this arrangement would have suited him. This type of transmission line has been used by us in our studies of the performance of dipoles and baluns. The use of a low-impedance line limits the extremes of impedance variations, and such a line is easy to fabricate and feed into the station. In addition, its losses are not high (we recommend foam-dielectric coax) unless you need a very long feed line. (Our 300-bead 1:1 balun of Mix 73 ferrite on miniature Teflon-dielectric coax is the best balun we have found for use with HF multiband dipole antennas, where the mismatch impedance can be high.[7])

Notes

[1] P. Hawker, "More on the G5RV/ZS6BKW Antennas," Technical Topics, *Radio Communications*, Jan 1993, pp 43-45; Feb 1993, p 34; and Apr 1993, pp 53-54.

[2] W. Maxwell, *Reflections* (Newington: ARRL, 1990), pp 20-13 to 20-16.

[3] L. Varney, "An Effective Multi-Band Aerial of Simple Construction," RSGB *Bulletin*, Jul 1958.

[4] R. Schetgen, ed, *The ARRL Handbook for Radio Amateurs*, 1993 ed (Newington: ARRL, 1992), p 33-9, Figure 15.

[5] B. Austin, "Computer-Aided Design of Multi-Band Dipole Based on the G5RV Principle," *Radio Communication*, Aug 1985, pp 614-617, 624.

[6] J. Belrose, "An Update on Compact Transmitting Loops," *QST*, Nov 1993, pp 37-40.

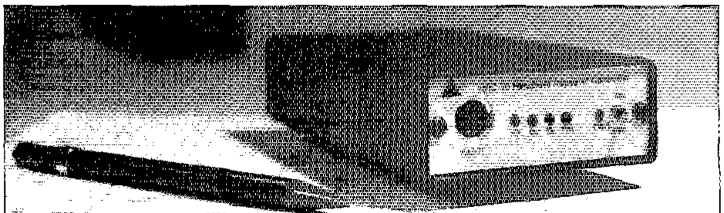
[7] Small quantities of 50- and 100-Ω Teflon-dielectric cable, as well as 50-bead W2DU and VE2CV 1:1 and 4:1 baluns in ready-to-use or kit form, can be obtained from The Wire Man, 261 Pittman Rd, Landrum, SC 29356, tel 803-859-4195. We have noted that radio amateurs in North America and the UK have experienced difficulty in locating a source for Teflon-dielectric coax—particularly the 100-Ω type necessary for the VE2CV 4:1 balun.

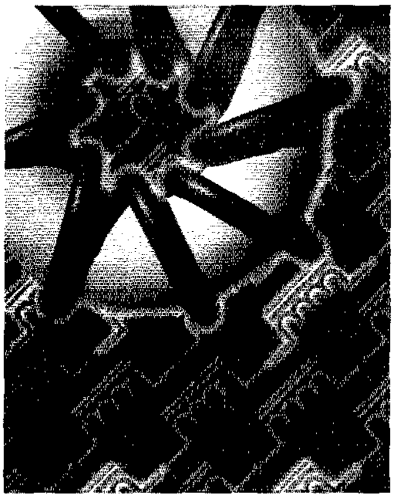
New Products

COMPACT CROSSBAND REPEATER CONTROLLER

◊ Many modern dualband transceivers offer crossband-repeater functions, but lack important functions, such as automatic IDer, hang and time-out timers, telemetry tones, private voice mail and Digital Voice Operated Squelch (DVOS). The HRC-10 is smaller than most hand-held transceivers, and converts a hand-held rig into a full-featured simplex or duplex repeater system. The compact controller connects quickly to the speaker and mike connectors, and can be powered by an internal battery or external dc supply. Optional carrier-operated squelch input and PTT output allows the unit to be used as a "repeater maker," and provides basic ID and timer functions for a repeater system. Retail price is $299 plus s/h. Spectrum Electronic Products, 4740 Scotts Valley Dr, Scotts Valley, CA 95066; tel 408-438-2788, fax 408-438-6027.

Choose crossband repeater frequencies carefully. Check with your local spectrum management or repeater coordination group for guidance—*Ed.*





Under the Hood IV: *Inductors*

They may not be common in audio and dc applications, but our radios wouldn't be radios without these important components.

By Bryan Bergeron, NU1N
Decision Systems Group
75 Francis St
Boston, MA 02115

Photos by Kirk Kleinschmidt, NT0Z

Inductors are critical to the operation of virtually every electronic communication device. Take your transceiver's power supply, for example. In addition to a power transformer, it probably includes one or more electromagnetic interference (EMI) filters, and possibly a filter choke. In your radio's receiver front end and intermediate frequency (IF) section, you'll likely find inductors used for tuning and impedance matching, as components of resonant filters, and for coupling signals from one amplifier section to the next. Similarly, your transmitter's circuitry likely uses inductors for filtering, signal coupling, and impedance matching. Inductors in the transmitter's radio frequency (RF) output stage are used for tuning, impedance matching, and antenna coupling.

Outside your transceiver, inductors match impedances and transform voltages in audio and power applications, as well as controlling potentially interfering signals that could emanate from the antenna feed lines and power, audio and data cables, speaker and microphone cables.

Inductance

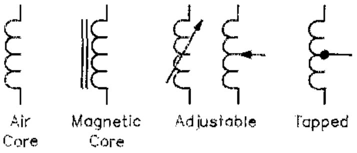
Inductance, which is expressed in *henries* (H, after physicist Joseph Henry), is a property exhibited to some degree by all conductors. Current flowing through any wire, cable, component lead, or circuit board tracing generates a magnetic field around the conductor. This magnetic field represents stored energy that can be harnessed in a variety of ways. For example, an inductor can be used as a *choke* to oppose changes in the current flowing through the conductor—in other words, to reduce or disallow ac flow on that conductor. Alternatively, a second conductor can be placed in the field to capture some of its energy and convert it to current flow. *Transformers* are based on this principle.

In that they tend to resist *changes* in current flow, inductors are the inverse of capacitors: They allow dc to flow while blocking or reducing ac. An inductor's ac resistance, its *inductive reactance*, increases with increasing frequency. The formula $X_L = 2\pi f L$ expresses this relationship, where X_L is inductive reactance in ohms, f is the frequency in hertz, and L is the inductance in henries.

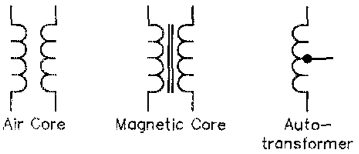
In theory, inductive reactance increases linearly with increasing frequency from dc to UHF and beyond. Such a predictable relationship between inductance and frequency is never observed in practice over very wide frequency ranges, however. Just as ideal resistors and capacitors exist only as ideas, there is no such thing as a perfect inductor. In practice, an inductor's reactance and inductance depend not only on its design, but also on factors such as current flow, frequency and temperature.

You can get an idea of an inductor's complexity by examining its equivalent circuit (Figure 1). The series resistance, R_s, represents the finite resistance of the inductor's wire. An inductor's quality factor, *Q*, depends on the ratio of its reactance (X_L) to its effective series resistance (R_s). Minimizing an inductor's resistance by using larger diameter wire, and plating the wire with silver or gold, therefore maximizes Q. High inductor Q is not always desirable, however. Low Q inductors are often preferred for power supply, noise filtering and interference suppression applications; higher Q inductors are used for sharply tuned circuits in which good efficiency is important.

An inductor exhibits distributed or parasitic capacitance, C_p, because capacitance exists between its adjacent coil turns. Distributed capacitance turns an inductor into a complete tuned circuit, making it *self-resonate* at one or more frequencies. Distributed capacitance can be reduced by constructing inductors with single layer windings whenever possible, and, when multiple layers must be used, by using special winding techniques, such as keeping the beginning and end of its winding(s) as far apart as possible.



Coils and Chokes



Transformers

Core Permeability and Inductance

Every wire exhibits inductance, but many radio circuits operating below a few hundred megahertz often require more inductance than circuit interconnections of ap-

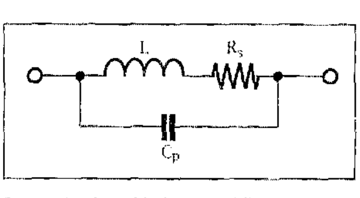
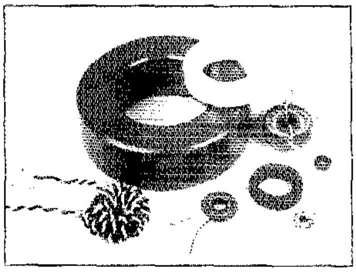


Figure 1—A real inductor exhibits more than just inductance: Its equivalent circuit reveals the presence of parasitic shunt capacitance (C_p), which makes the inductor act as a tuned circuit and *self-resonate* at one or more frequencies, and series resistance (R_s), which affects the inductor's efficiency and Q (quality factor).

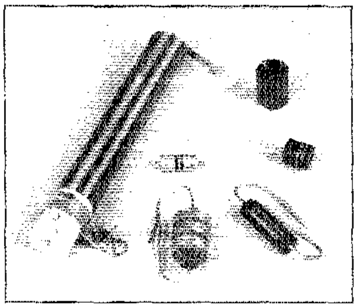


With magnetic cores, larger size means higher power-handling capability. These are cores and coils in torus form—*toroids*. Toroidal inductors are largely self-shielding and allow high component density without coupling magnetically to adjacent components.

propriate length can supply. Instead of relying on tens or hundreds of feet of wire to obtain specific values of inductance, we use compact, *lumped* inductances. Lumped inductors usually use wire that has been *coiled* to concentrate their magnetic fields and so increase their inductance.

In addition to coiling, a wire's inductance can also be increased by winding it around or through a *core* of iron, powdered iron, ferrite, or other material of high *permeability*. Permeability is a measure of how much better a given material is than air as a path for magnetic lines of force. Air is defined as having a permeability of 1. Inserting a core of higher permeability raises its inductance. Threading the core and coil to allow the core to be screwed in and out of the coil makes the inductor variable. Many radios use variable RF coils and transformers of this type.

Inductor cores vary widely in Q, frequency range, stability, and permeability, depending on their composition, shape and size. *Toroidal* (doughnut-shaped) cores are commonly used in today's equipment because they are compact and largely self-shielding.



Choke is often used as a generic name for any two terminal inductor—even an inductor that isn't used as a choke! This group includes a *plate choke* (the largest coil), intended to block RF energy while carrying high voltage dc feed to a vacuum tube plate (anode); a *pie-wound* RF choke (the three section component); a ferrite cased inductor (the black cylinder) for use at audio and low radio frequencies, and encapsulated chokes.

Magnetic cores may *saturate* at elevated energy levels. That is, they have a finite capacity for magnetic energy. Exceeding this capacity may cause the core to overheat. Depending on the material, overheating may permanently change the core's permeability.

Inductor Types

Most inductors can be classified as coils, chokes or transformers, based on their construction.

Coils and Chokes

Coils and chokes are single winding inductors. The main difference between them is how they're used. A coil is usually used with capacitance to resonate or tune a circuit to some desired frequency. A choke usually does just what its name implies: It *chokes* (reduces or eliminates) ac flow between two circuit points while allowing other signals (dc, or ac signals significantly different in frequency than those "choked") to pass.

Chokes rarely exist as a simple coil of wire with solder leads attached, but are generally wound on some sort of core or form for stability, ease of handling, and, in many cases, increased inductance. Low inductance designs use forms made of nonmagnetic materials, such as plastic, ceramic or glass. Intermediate inductance designs use powdered iron forms. Ferrites, which are compounds of iron oxide and other metallic oxides combined with a ceramic material, are used when high inductance is required.

Coil and choke types you may hear about include:

RF chokes. These vary considerably in design. Molded RF chokes, shielded axial lead inductors encapsulated in epoxy, are designed for minimum coupling in high density circuit configurations. For less densely populated circuit boards, inexpensive axial-lead RF chokes wound on phenolic or powdered-iron cores are available (Figure 2). When relatively high values of inductance are required, multisection windings are often used instead of a single, multilayer winding to reduce distributed capacitance. These are traditionally called *pi-wound* or *pie-wound* chokes because each section is sometimes called a *pi* or *pie*.

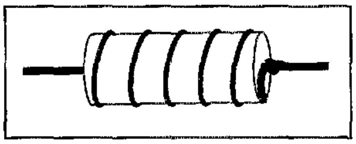


Figure 2—A single-layer RF choke is just a wire *solenoid* wound on a phenolic or ferrite core.

Ferrite beads, fitted over the leads of a component, serve as a popular and inexpensive form of low inductance, low Q RF choke. A ferrite bead has a center hole; when placed on a current-carrying conductor, the bead works with the wire to act as a choke. The choke's inductance is a function of the bead's material, frequency, and size. Although single-hole beads are most popular, multiple-hole beads are also available for using in achieving larger inductance values and building miniature transformers.

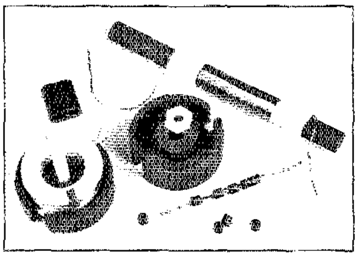
Filter chokes attenuate rectifier output ripple in power supplies. The 20 to 30-A power supplies common in ham stations nowadays usually don't contain filter chokes. Chokes were more common in the higher voltage, lower current supplies used in vacuum tube based, vintage Amateur Radio gear. Such chokes generally exhibit inductances from 1 to 50 H, with resistances up to several hundred ohms. A filter choke in a modern high current, low voltage application might provide up to 30 mH or so of inductance with a dc capacity of 12 A or more, and a resistance of less than 1 Ω.

Common-mode chokes. RF or noise energy that flows on both of a cable's wires in the same direction—in the *common mode*—causes many EMI problems. Choke that current flow by winding several turns of the cable through or around a ferrite core, and you choke the interference. This is what "clamp on" EMI suppression chokes do.

Roller inductors (Figure 3), used in some antenna tuners, include a rotating, bare wire coil contacted by a grooved, spring loaded wheel. You can usually wire the moving contact so that it progressively short-circuits the coil or acts as a movable tap.

Miniductor (Figure 4), a trademark of Barker & Williamson, has nearly become a generic for self-supporting, air-wound coils held in shape by molded plastic bars. B&W now also markets the comparable Air-Dux coils once made by a competitor, Illumitronic Engineering.

Surplus telephone toroids are low to medium Q inductors once used by telephone



Magnetic cores come in other than toroidal form, including *ferrite beads*, which can tame interference and stabilize circuits when slipped over wires, as shown here; *cup cores*, inside which a nylon bobbin carries the winding or windings; and rods (your AM radio probably uses one in its built-in antenna).

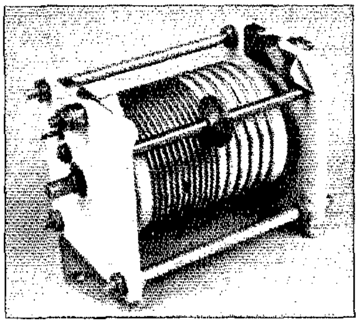


Figure 3—Some hams covet *roller inductors* as a means of adjusting the inductance with infinite resolution in antenna tuners. But the relatively high expense of roller inductors and the turn-counting dials necessary to drive them make other alternatives more attractive for high power variable inductances.

companies for line equalization. Usually available in 44/11 or 88/22 mH form, they enjoyed a surge of popularity for audio filtering use in the 1960s and 1970s. Their increasing age and the availability of self shielding inductors in comparable standard values has rendered them nearly obsolete, even in ham applications.

Transformers

As I mentioned earlier, transformers use the principle of capturing the energy in the magnetic field surrounding one inductor and converting it to current flow in a second inductor. Part of the electrical energy in a transformer's input inductor (*primary* winding) is transformed into a magnetic field, which is, in turn, transformed back into electrical energy by the second inductor (*secondary* winding). In a transformer with multiple input and output windings, all of input windings are called primaries and all output windings are called secondaries.

A transformer translates a given voltage, current, or impedance level to another, in proportion to the ratio of the number of wire turns in its primary and secondary windings. Some transformers have the same number of turns in their primaries and secondaries. This equates to a turns ratio of 1:1, or *unity*. Such transformers are used mainly to isolate—break direct wire connections between—two circuits. Telephone isolation transformers, which usually have 600-Ω primaries and secondaries, are a common example of these.

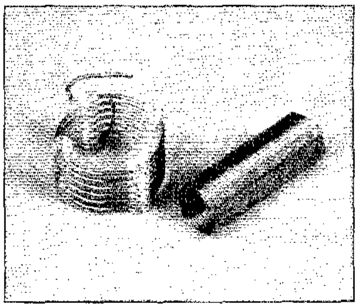


Figure 4—Self-supporting air core coils called *Miniductors* were popular with hams before toroidal inductors took over with the coming of miniaturization. Some Miniductor stock isn't so mini: A few types are 6 inches in diameter!

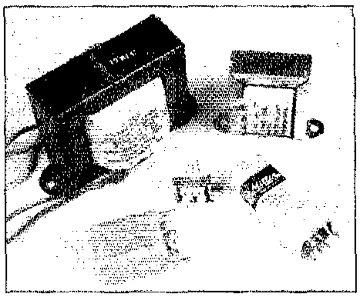
As with chokes, transformer cores vary widely in shape, size and composition. Most RF transformers in today's radios use ferromagnetic cores of some type; air core RF transformers were common in the vacuum tube gear of yesteryear. Cores of laminated steel plates or wound steel tape are the norm at audio and power frequencies. Toroidal cores are sometimes used in audio and power transformers, too.

Power transformers change ac line voltage to a voltage more suitable for conversion to dc. They're present in the ac-to-dc power supplies associated with base station radio transceivers, virtually all ac powered peripherals and test equipment, and stand-alone dc power supplies. *Filament transformers* are power transformers once intended to heat the cathodes of vacuum tubes: nowadays, we rectify and filter their low voltage output for powering solid-state circuitry. (Some power transformers, especially those used in instrumentation, include an interwinding *electrostatic shield* that keeps high frequency line noise out of secondary windings.) *Wall transformers* are power transformers that plug directly into wall ac outlets. They may contain just a transformer or a transformer and rectifier/filter circuitry.

Audio transformers do the same basic job as power transformers—after all, ac line energy is just high-power audio at 50 or 60 Hz—but over a broader frequency range, often with attention paid to low distortion and hum immunity. A *modulation transformer* is an audio transformer intended for use in between a modulated RF power amplifier and the audio amplifier that modulates it.

Autotransformers are unusual in that they use a single winding for primary and secondary. Nowadays, autotransformers are most commonly used to match ac equipment and power sources that operate at different voltages. *Variable autotransformers* allow continuous output voltage adjustment by means of a movable secondary tap. Variable autotransformers capable of producing 0 to 120 or 0 to 130 V from the 120 V ac line are traditionally referred to by the genericized trade names *Variac* or *Powerstat*.

Balun (balanced-to-unbalanced) and *unun* (unbalanced-to-unbalanced) RF transformers are widely applied in double-balanced mixers, directional couplers, signal splitters, and signal distributors and impedance matching circuitry.



As a class of components, transformers operate from megawatts to microwatts and come in shapes and sizes that span a comparable range. Here are just a few: a "filament" power transformer that steps 120 V down to 6.3 V; a hermetically sealed, high fidelity audio transformer; a Realistic "line-to-voice-coil" audio matching transformer, a tiny "transistor" transformer intended for PC board mounting; and an IF transformer from vintage, vacuum tube gear. (Its transistor equivalent is 1/5 the height and 1/2 the depth and width!)

Flyback transformers, also called *horizontal output transformers*, provide the high horizontal scanning and anode voltages for TV and computer monitor cathode ray tubes (CRTs).

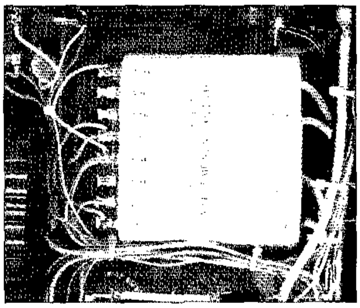
Inductor Characteristics

Inductance

Most manufacturers specify inductance for RF parts in millihenries (mH) or microhenries (μH) within a given tolerance and at a specific test frequency. Filter chokes carry values in henries or millihenries. Tolerance values vary from a few percent for precision fixed inductors to 10 to 30% for general purpose inductors.

Q

Typical Q values range from 10 to several hundred, depending on inductor design and application. Especially for tuned circuit applications, Q is most useful when



The transformer in your transceiver's power supply may include taps that allow it to operate at different ac line voltages encountered around the world.

measured at or near the frequency at which the inductor will operate, because Q generally varies significantly with frequency.

Other Frequency-Related Characteristics

Manufacturers sometimes specify an inductor's *self-resonance frequency*, especially for fixed RF chokes. *Frequency response* (frequency span equated with amplitude response "flatness" [a ± figure in decibels], usually together with a power level specification [watts or milliwatts]) is often stated for audio transformers.

Resistance

An inductor's dc resistance is an important characteristic if it must carry dc in addition to choking or transforming ac, or while serving as a tuned circuit inductor.

Application-Dependent Specifications

Depending on the inductor type and application, some specifications are more important than others. For example, few of us care about the inductance, self-resonant frequency or Q of a power transformer. Instead, we want to know its power handling capability (and/or its secondary's current carrying capacity), the breakdown voltage of its primary to secondary insulation, its maximum operating temperature (such as 105°C), its expected lifetime in thousands of hours, its maximum primary-to-secondary leakage current (such as 10 µA for a typical isolation transformer), its frequency rating (50, 60 or 400 Hz, or a range of frequencies) and, for critical applications, its stability. Similarly, relevant specifications for audio transformers may include frequency response, degree of magnetic shielding, maximum operating level, percentage of waveform distortion, and insertion loss in decibels. RF baluns are commonly rated by impedance, insertion loss, frequency response and power handling capability.

Inductor Labeling

Inductors and transformers large enough to carry plain text verbal labels usually do, as some of the photos indicate. Color bands based on the value/multiplier scheme used for resistors may label encapsulated RF chokes as shown in Figure 5.

Summary

Induction and inductors would have plenty of work to do even if humankind had not discovered and harnessed radio, because induction lies at the heart of most devices that use electrical energy to produce motion, and devices used to convert one ac voltage to another. These inductor functions are important to our radios' operation, but arguably not as important as the inductor's ability to *resonate* when used with capacitance—to *tune* a circuit to one ac frequency instead of another. Whenever you receive or transmit radio signals, inductance and inductors are on the job. Now you'll be able to spot inductors at work the next time you pop the top off your radio and take a peek under the hood.

Variable Inductors

Making an inductor variable usually involves moving a magnetic core in and out of a fixed coil. Moving the core farther into the coil increases the coil's inductance, and vice versa. (Brass cores were sometimes used to make inductors variable—moving brass into a coil *reduces* inductance—but this practice fell out of favor because it lowers a coil's Q.) Radios can even be tuned this way: Ten-Tec's current 555 Scout uses just such a PTO (*permeability tuned oscillator*), as did much vintage Collins and R. L. Drake radio gear.

Inductors can also be made variable by expanding and compressing their turns; for a given number of turns, closer spacing means more inductance, wider spacing means less. An inductor can also be made variable by splitting it into two windings and moving one part relative to the other to *boost* (add inductance) or *buck* (subtract inductance) through addition or subtraction of their magnetic fields. And an inductor can be made variable by short-circuiting its turns with a multipoint switch or moving contact, as in roller inductors.—*WJ1Z*

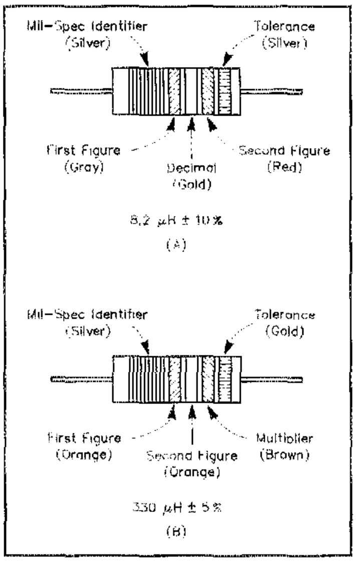


Figure 5—Color coding for encapsulated RF chokes. The color code is equivalent to that of the resistor code, but the bands are reversed relative to those on resistors to help tell the two component types apart. At A, an 8.2-µH, 10% tolerance part; at B, a 330-µH, 5% tolerance part.

New Products

KITS AND PARTS

◊ A line of compact transceiver kits designed by Rick Littlefield, K1BQT, features full-range VFOs with 8:1 reduction drives, crystal-filter superhet receivers, RIT, front-panel selectable audio filters, adjustable-delay break-in, built-in sidetone oscillators and attractive Ten-Tec cabinets. A series of QRP transceivers for 15, 17, 20, 30 and 40 meters provides 5 watts output (specify model, by band: QRP-15, QRP-17, QRP-20, QRP-30 or QRP-40). Retail price is $99.95 each, including shipping in the continental US. The compact 20-meter Travelradio transceiver features a sensitive superhet receiver, 9-MHz crystal filter with a front panel selectable audio filter, RIT, sidetone, adjustable-delay break-in, full-range VFO with an 8:1 reduction drive and a Ten-Tec enclosure. The 30-watt model is $149 and the 15-watt rig is $139.

K1BQT's SSB transceiver kit for 3.8-4.0 MHz (easily retuned to any portion of 75 meters) features 15 watts output at 12-14 V dc and 35 watts output at 24-28 V dc, with a superhet receiver, 9-MHz filter, wide-range AGC, speaker, S meter, amplified ALC and a cabinet. Retail price is $179. For $189, the kit is available with an optional crystal-control module for MARS/CAP operation on up to six frequencies in any 300-kHz segment from 3.5-5.0 MHz.

The QRP 3-Bander kit designed by ARRL Lab Engineer Zack Lau, KH6CP, as featured in October 1989 *QST* and the *ARRL Handbook*, puts out 4 watts on 18, 21 and 24 MHz, with a VXO-controlled transmitter and a direct-conversion receiver, full QSK, sidetone, spotting and an attractive enclosure. Retail price is $89.95.

The *QRP Classics* 6-watt VXO Transmitter (3.5-21 MHz, specify band) is available for $54.95. The *QRP Classics* Universal QRP Transmitter provides 1½ watts output on 160-80 meters (basic kit without crystal or cabinet $22.95, complete kit $38.95), 40-10 meters (basic kit $20.95, complete kit $36.95). The Neophyte Receiver for 80 or 40 meters retails for $34.95. The *QRP Classics* 50-mW rig with a 7.060-MHz crystal is $19.95. The Optimized 7-MHz transceiver from the 1992 *ARRL Handbook* retails for $49. K1BQT's Micro Receiver for 20-meter CW and SSB is $49.

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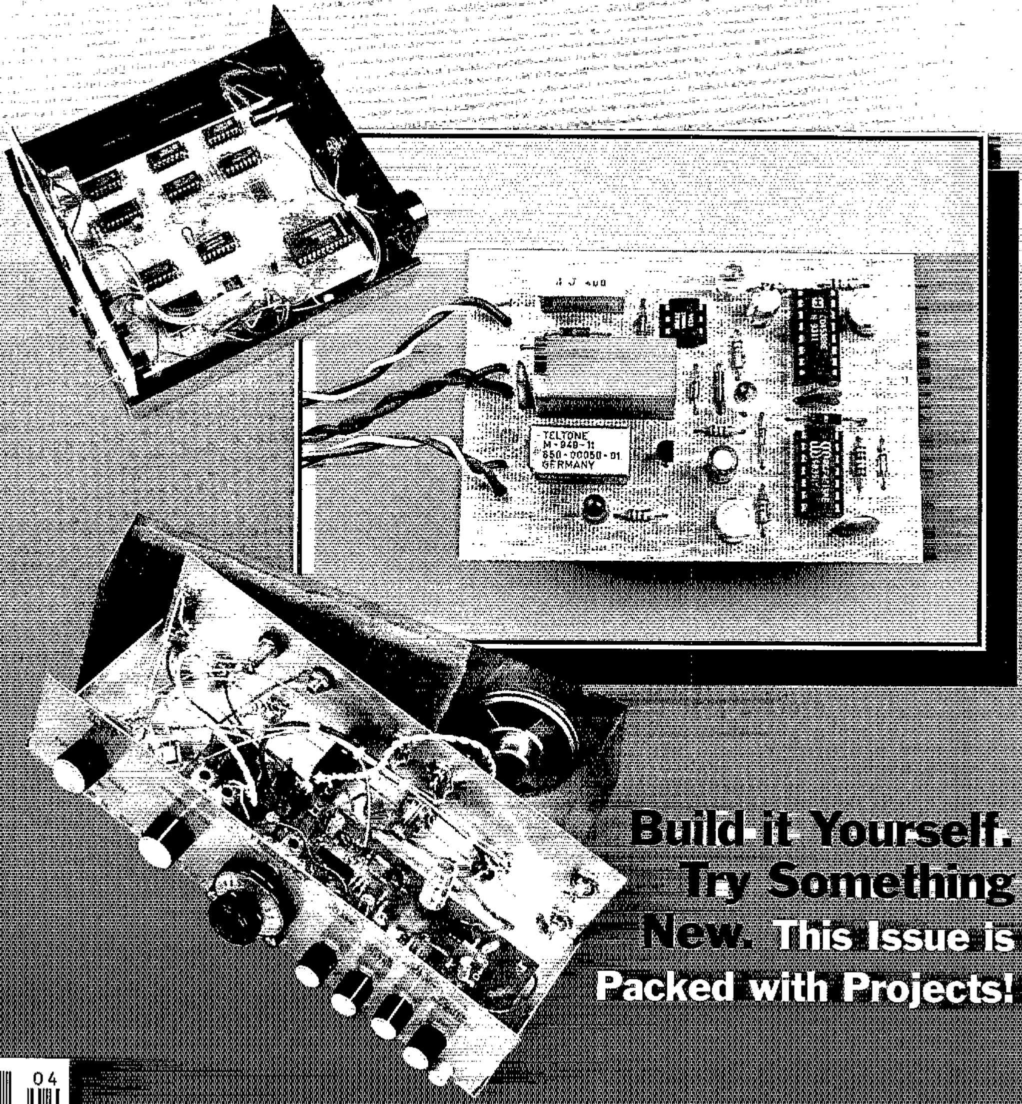

Build it Yourself. Try Something New. This Issue is Packed with Projects!

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OUR COVER

Spring out of the winter doldrums with these fresh new home-brew projects. W9XD describes a flexible Tone Alert device (top) that monitors a frequency for group call-ups, the ARRL-endorsed LiTZ system and other DTMF sequences; the WA4TEM Ring Master (center) can protect repeater/autopatch users from interception by Caller-ID systems; and WD4PLI shows you how to build his *Lowfer* Transceiver to slide *way* down the dial for experimentation on 1750 meters (160-190 kHz). *(photos by Kirk Kleinschmidt, NT0Z).*

CONTENTS

April 1994
Volume LXXVIII Number 4

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

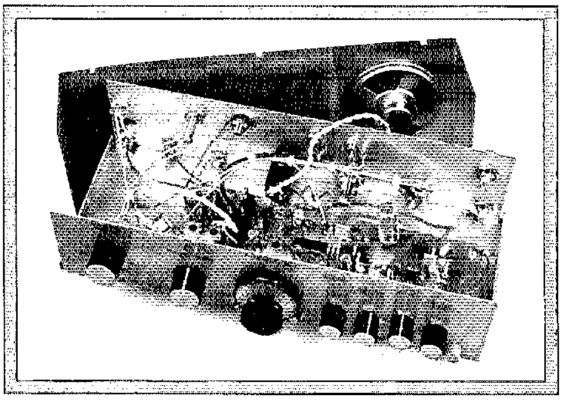
OPERATING

DEPARTMENTS

Build Your Own *Lowfer* Transceiver

Explore the 1750-meter band with this high-performance CW transceiver.

By David Curry, WD4PLI
737 N Fairview St
Burbank, CA 91505



Black crinkle finish, 82 pounds, 1944. What do these words have in common? Why the *RBL* low-frequency receiver, of course! And what makes the RBL LF/VLF receiver so special? The RBL symbolizes the Low Frequency/Very Low Frequency experience for the people who explore the depths of the 1750-meter band (160 to 190 kHz). They're the *Lowfers*, radio experimenters—many of them hams—who enjoy this band and manage to carry on useful communications despite low-power restrictions and high noise levels. *No license is required to operate on 1750 meters.* Anyone can use the band. All you need is the proper equipment.

Which brings us back to the RBL...

My first encounter with an RBL was at Fred Wilson's electronic shop in 1971. He had one sitting on a shelf, next to an old Tektronix 'scope. You could feel the warmth of the tubes and smell the capacitors when you opened the door to his shop. There was something alluring about his RBL receiver. Even the dials were impressive, especially when you turned the big frequency vernier.

One day we connected a wire antenna and heard...noise! Lots of noise. Fred sold it to me later for $40 and I used a garbage cart to take it around the block to my house. Sometime later Ken Cornell, W2IMB, gave me some details on how to make a simple loop antenna that worked very well with the RBL. Sure enough, I received dozens of distant aero/marine beacons late at night in the bedroom closet, lit only by the yellow lights of the RBL.

In due time I became a Lowfer after building my first 1750-meter station and working Todd Robert's "ABC" in January 1974. The "feel" of 1750 meters is in harmony with the RBL. The Lowfer band is filled with mystery, challenge, antiquity, and much history. These low frequencies are where radio communication began. During WW II the RBL was a symbol of the state of the art. Many of these receivers served our nation's interests, and remain to this day in sunken hulks at Pearl Harbor, and in ill-fated submarines.

The RBL design was the inspiration for this simple Lowfer CW transceiver—the *CW-893*. You can build it in a couple of evenings and you'll soon be on the air on 1750 meters!

Description

The receive portion of the CW-893 is a virtual duplicate of the RBL, but in modern form! The RBL had unbeatable stability and sensitivity with its regenerative detector. The transceiver described here does not use a regenerative detector, but a *direct conversion* approach instead. The front end preselector uses a tunable two-pole Chebychev bandpass filter to remove unwanted signals. Noise is always a problem at these frequencies, so two noise limiters are included that provide very effective limiting of man-made and natural noise.

Audio filtering is included, with variable frequency and bandwidth controls for precise filtering of the desired signal. Ample audio output drives headphones and most speakers. The CW-893 is capable of providing over 100 dB of receive gain with virtually no power supply hum.

The transceiver generates 1 W of input power with its Class-E MOSFET amplifier. FCC Rules limit Lowfer power to 1 W, which is why CW is the mode of choice for this radio. Semi-break-in operation is provided with an adjustable time delay. The frequency is VFO controlled.

Although this transceiver is basic in concept, it incorporates all the required features for successful two-way operation on 1750 meters. Your range will extend from about a mile to 200+ miles, depending on band conditions and the quality of your antenna.

As you probably know, purchasing components these days can be expensive, which was a major concern when I designed the CW-893. All parts are off the shelf, with part numbers provided.

Construction Notes

Several parts are soldered directly to the component side of the circuit board. This provides the ground connection for many components. Be sure to solder these leads to *both sides* of the PC board. All parts are easy to identify. The capacitors are disc or round shaped, while electrolytics are round with the polarity marked. Transistors are designated by the half moon shape, or round with a key. ICs are rectangular with notches at the ends.

I recommend soldering the ICs first. Notice that some pins must be soldered on the component side. The next step should be soldering L1, L2 and T1. Switch S1 should be installed after you mount the potentiometer. The switch is mounted on the solder side of the circuit board.

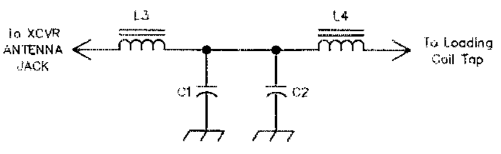


Figure 1—An auxiliary low-pass filter for the CW-893 transceiver. Installation is optional. L3 and L4 are made from 43 turns of #20 wire wound on an Amidon T-94-3 core. C1 and C2 are 0.01 µF polystyrene capacitors (Mouser 23PS310)

The auxiliary low-pass filter described in Figure 1 is not strictly necessary. It does offer a substantial reduction in harmonic output, however. To make absolutely certain that your transceiver conforms to FCC Part 15 Rules, I recommend that you build and install the filter in the enclosure.

Transformer T2 must be wound by hand on a T-68-3 toroid core available from Amidon Associates and other sources. The kit manual provides detailed instructions. In addition, see the *ARRL Handbook* for more information on winding toroid transformers. The T2 primary consists of 93 turns of #30 enamel-insulated wire. The secondary is made of 49 turns of #25 enamel-insulated wire. Wind the turns evenly and firmly. After you've finished, cut the wires so that about one inch remains from the toroid to the end of each wire. Remove the enamel insulation from the ends with sandpaper.

The CW-893 is available as a kit complete with a circuit board, parts and instructions. All you have to supply is a suitable enclosure and various knobs. You can purchase the kit directly from me for $94, shipping included (California residents please add sales tax). If you want to shop for your own components, you can buy only the circuit board at a cost of $22. I can provide information sheets on the CW-893 in return for a self addressed, stamped envelope and three units of First-Class postage.

Checkout and Alignment

Refer to Figure 2 and locate the following points:

A: 50-Ω transmit/receive antenna jack.
B: CW key jack.
C: Speaker/headphone output.
COMM: Common terminal for auxiliary relay.
D: Frequency monitor port.
E: Audio output for external amplifier, or AB1 deluxe audio board (available as a kit from the author).
GND: Ground.
JP1: Receive input select. Short JP1A and JP1B to use an antenna at port A for receive. (Receive-only antennas connect to JP1B.)
N/C: Normally closed terminal for auxiliary relay control.
N/O: Normally open terminal for auxiliary relay control.
VCC: 12 to 18 V dc.

Connect a 12-V power source to the V_{CC} points and ground. A frequency counter or receiver covering 150 to 250 kHz will be required for the following steps.

Connect a frequency counter to point **D**. Switch the transceiver on. Adjust tuning capacitor C10 to its maximum clockwise position. Turn the slug in T1 until the frequency reads 189 kHz on the counter.

If you don't have a frequency counter, don't worry. You can also use a long-wave receiver, general coverage receiver, or ham transceiver that can accurately tune to 189 kHz. Attach a small piece of wire to the antenna jack. Tune your monitor receiver to 189 kHz. Listen for a tone while turning the slug of T1. Keep turning the slug until a zero beat is heard.

Next, align the preselector. Inductors L1 and L2 must be tuned to the same frequency. If you have a signal generator, adjust it for a low level (approximately 100 µV) at the antenna jack. Turn the **Preselector** and the **Filter Frequency** controls to their 12 o'clock positions. Rotate the **Series Limiter** and **Filter Bandwidth** controls fully counter-clockwise. Tune the **Frequency** control until you can hear the generator's signal. Adjust the slugs on L1 and L2 for maximum volume in your speaker or headphones, decreasing signal generator output as the generator tone becomes louder.

If you don't have access to a signal generator, connect a long piece of wire to the antenna jack and listen for any signals you can find (even an interference signal will do!). Once you find a usable signal, turn the **Preselector** capacitor to the same general setting as the **Frequency** capacitor. Now adjust the slugs in L1 and L2 for maximum signal strength.

Operating Tips

The audio gain stage (adjusted through the use of the **Volume** control) has a built-in limiting function. This can be used to increase the gain of a signal that's buried in man-made noise, effectively cutting off the peaks of the noise while leaving the signal unaffected. The **Series Limiter** clips any remaining distortion from the shunt limiter and lowers the volume to a comfortable level. The audio **Filter Frequency** and **Bandwidth** controls are adjusted for the amount of filtering desired.

While there is plenty of audio power available to drive a speaker, I recommend that you use headphones whenever possible. Direct-conversion receivers derive much of their gain in the audio amplifier stages. As a result, a high-gain audio amplifier—such as the one used here—may begin to oscillate if you turn it up too high. (You'll know this is happening if you hear a howling sound!) By using headphones, you'll hear the weak signals much better and you won't need nearly as much audio power.

Direct-conversion receivers are also prone to *microphonics*—amplification of mechanical vibrations. Care should be used if you install a speaker near the chassis or circuit board to avoid any audio feedback.

An important feature of the CW-893 receiver section is the input **Preselector** control. The preselector filter is very sharp, allowing only a small slice of the band to be received. If, for example, the signal you want to hear is on 180 kHz, tune the **Frequency** control to either 179 or 181 kHz. The signal will be heard as a 1-kHz tone (180 kHz − 179 kHz = 1 kHz, or 181 kHz − 180 kHz = 1 kHz). Choosing whether the upper or lower frequency is best depends on which provides the clearest reception. During two-way operation, for example, you might be transmitting at 182 kHz with the preselector peaked to your friend's signal at 182.4 kHz. By the same token, your friend's preselector would be peaked at *your* frequency (182 kHz). As you can see, tuning the preselector above and below your center frequency provides a lot of flexibility.

Transmitting with the CW-893 is as easy as plugging in a CW key, selecting a clear frequency, and using a *resonant* vertical transmitting antenna. Consider a call sign that uses the last two or three letters of your Amateur Radio call sign. (It's considered poor practice to use your full call sign.)

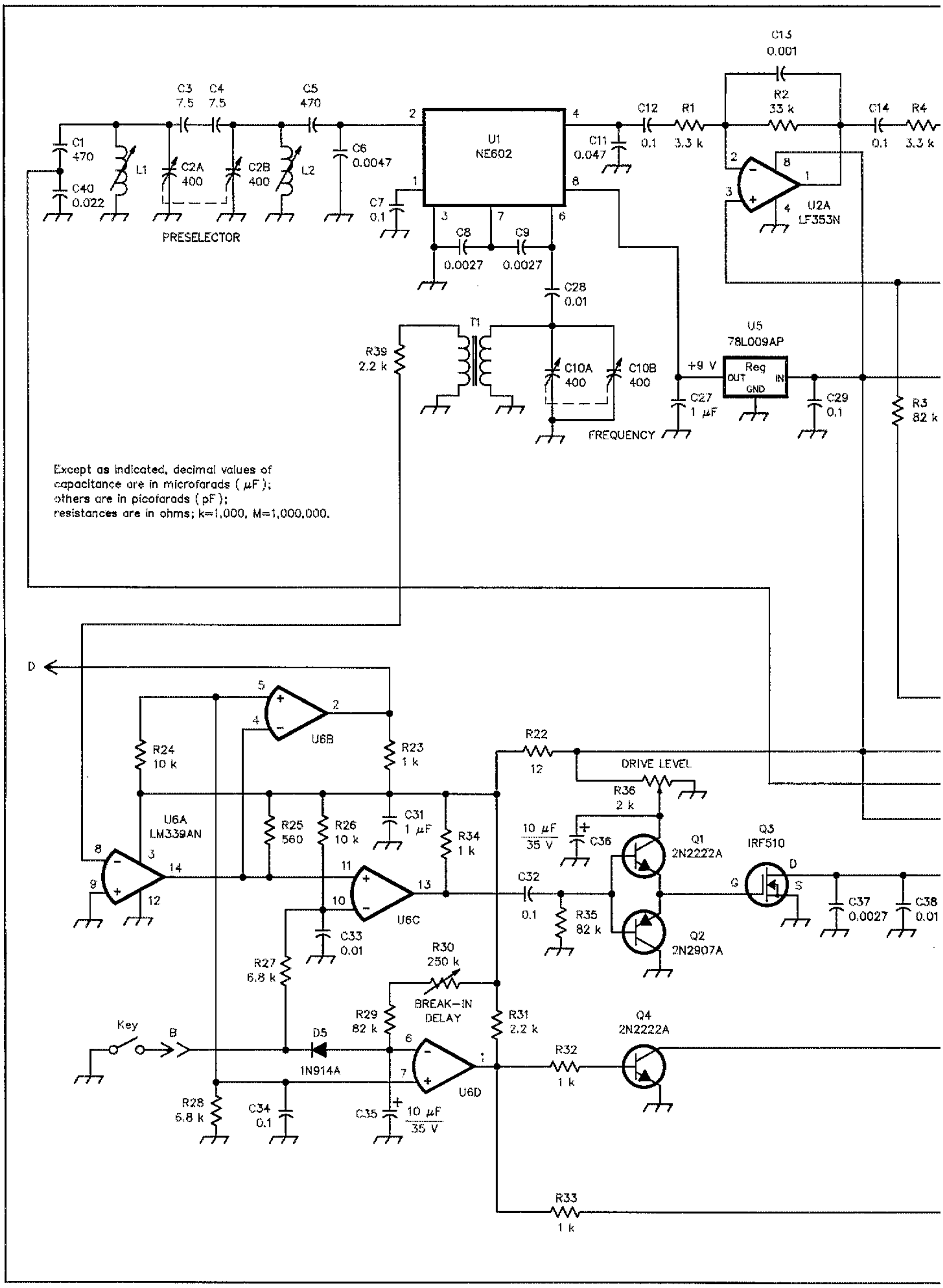
When transmitting, you'll want to adjust your TR delay potentiometer (R30) for the desired keying delay. In addition, the power amplifier drive control (R36) should be set for desired input power. You could install a 1-mA meter in the enclosure to monitor the power amplifier current. Meters can be expensive, however, so a VOM or VTVM can be used instead. Connect this to the meter − and + points on the circuit board. The voltage indicated corresponds to the input current to the power amplifier. One watt of input power is 83 mA at 12 volts, or *83 mV* on a VOM or VTVM connected to the − and + points.

When you're not enjoying a conversation, you can use the transceiver to send a beacon signal. Beacons are helpful to other Lowfers who want to know if they can hear you. It also helps with antenna experimentation and propagation tests. The transceiver is easy to use as a beacon. Simply connect your beacon ID generator (a CW keyer set to the "repeat" mode, for example) to the key input (point "B").

Antennas

The type and location of your 1750-meter antenna are of paramount importance. FCC Rules restrict the length of your *transmitting* antenna system to 15 meters. Just about any type of antenna design will do for casual, short-range experiments (less than one mile).

If your goal is optimum transmitting performance over greater distances, you *must* use a *resonant* vertical antenna (see Figure 3A). This involves resonating the antenna with a *loading coil* at the feed point and using a good ground system. The ground system can be composed of eight (or more) radial wires, each 30 feet or more in length. Terminate the radial ground wires with four copper pipes used as ground rods. If a ground system is required in rocky or sandy soil, or you want to roof mount the antenna, use a counterpoise resonant radial system. The cold water pipe in your home is an alternative if



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

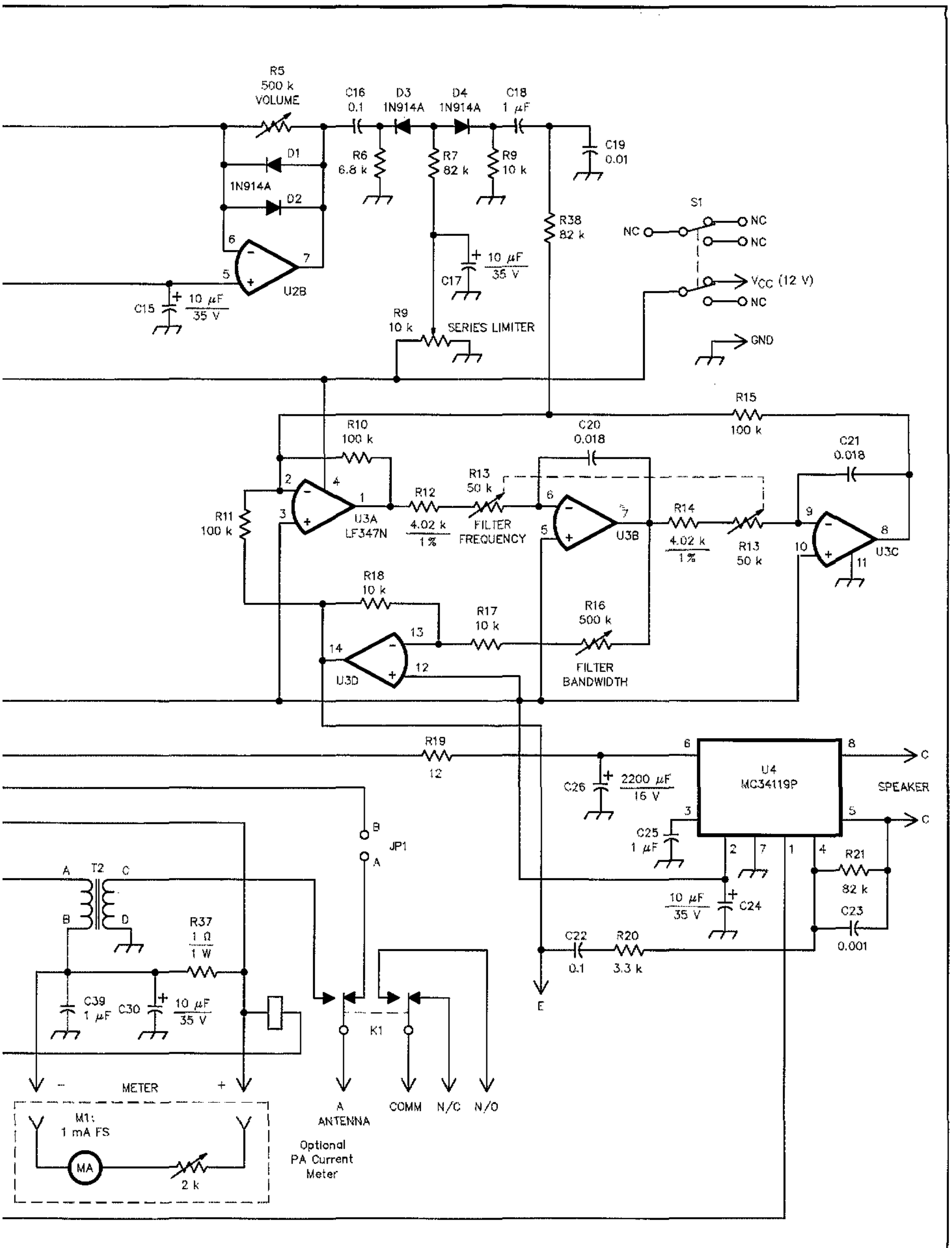


Figure 2—See caption on the next page.

Fig 2—Schematic diagram of the CW-893 transceiver. Resistors are ¼-watt, 5% tolerance carbon-composition or film except as noted below.

C1, C5—470-pF polystyrene (Mouser 23PS147).
C2, C10—400-pF variable (Mouser 24TR218).
C3, C4—7.5-pF ceramic disk (Mouser 21CB008).
C6—0.0047-µF polystyrene (Mouser 23PW247).
C7, C12, C14, C16, C22, C29, C32, C34—0.1-µF ceramic disk (Radio Shack 272-135).
C8, C9, C37—0.0027-µF polystyrene (Mouser 23PS227).
C11—0.047-µF film (Digi-Key P4521).
C13, C23—0.001-µF polystyrene (Mouser 23PW210).
C15, C17, C24, C30, C35, C36—10-µF, 35-V electrolytic (Radio Shack 272-1025).
C18, C25, C31, C39, C27—1-µF monolithic (Newark 90F1907).
C19, C33—0.01-µF ceramic disk (Radio Shack 272-131).
C20, C21—0.018-µF polypropylene 12% (Digi-Key P3183).
C26—2200-µF, 16-V electrolytic (Radio Shack 272-1020).
C28, C38—0.01-µF polystyrene (Mouser 23PW310).
C40—0.022-µF polystyrene (Digi-Key P3223).
D1, D2, D3, D4, D5—1N914A diode (Radio Shack 276-1122).
K1—DPDT relay (Digi-Key Z768-ND).
L1, L2—1.5-mH variable inductor (Digi-Key TK3203).
Q1, Q4—2N2222A (Radio Shack 276-2009).
Q2—2N2907A (Radio Shack 276-2023).
Q3—IRF510 power MOSFET (Radio Shack 276-2072).
R1, R4, R20—3.3-kΩ (Radio Shack 271-1328).
R2—33-kΩ (Radio Shack 271-1341).
R3, R7, R21, R29, R35, R38—82-kΩ.
R5, R16—500-kΩ, pc-board pot (Mouser 31CW505).
R6, R27, R28—6.8 kΩ.
R8—10-kΩ, linear taper, pc-board pot (Mouser 31CW401).
R9, R17, R18, R24, R26—10 kΩ (Radio Shack 271-1335).
R10, R11, R15—100 kΩ, 1% (Mouser 29MF250-100K).
R12, R14—4.02 kΩ, 1% (Mouser 29MF250-4.02K).
R13—50-kΩ, ¼-W, dual audio taper (Calrad 25-411).
R19, R22—12 Ω.
R23, R32, R33, R34—1 kΩ (Radio Shack 271-1321).
R25—560 Ω.
R30—250-kΩ trimmer (Mouser 32RM503).
R31, R39—2.2 kΩ (Radio Shack 271-1325).
R36—2-kΩ trimmer (Mouser 32RM302).
R37—1 Ω, 1 W (Mouser 29SJ901).
S1—DPDT switch (Digi-Key EG1003-ND).
T1—0.63-mH transformer (Digi-Key TK1201).
T2—Toroid transformer (see text).
U1—NE602 mixer/amplifier (Digi-Key NE602AN).
U2—LF353N low-noise op amp (Mouser 511-LF353N).
U3—LF347N quad op amp (Mouser 511-LF347N).
U4—MC34119P audio power amplifier (Newark MC34119P).
U5—78L009AP 9-V regulator (Mouser 333-78L009AP).
U6—LM339AN quad comparator (Mouser 511-LM339AN).

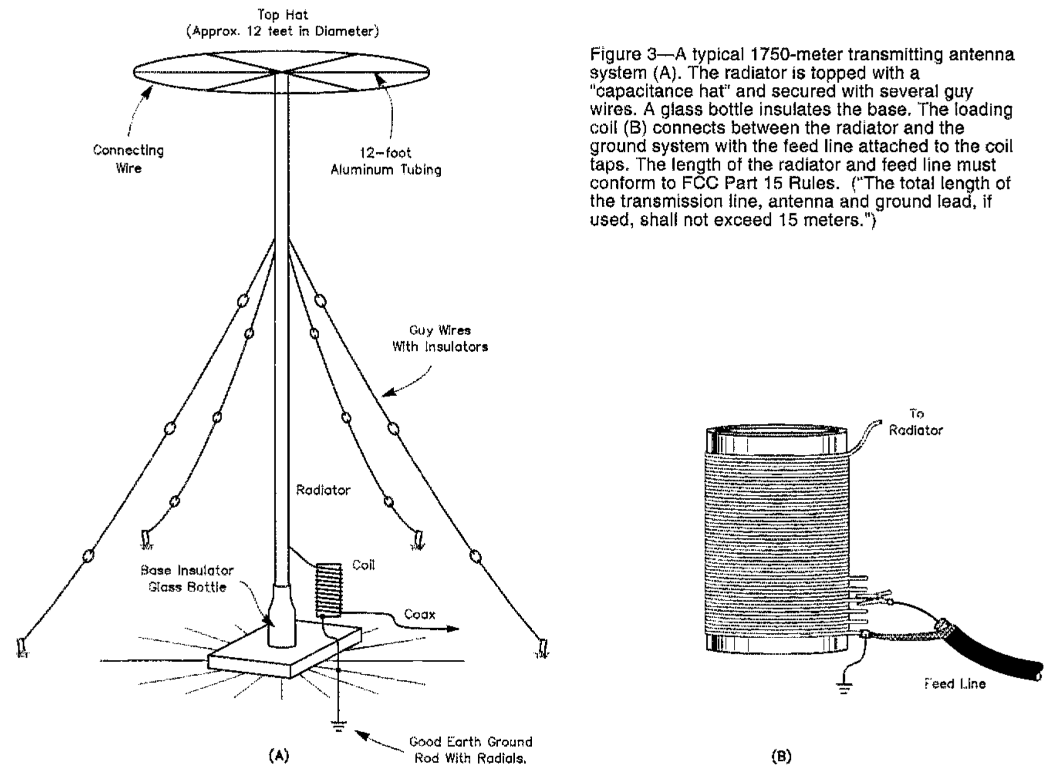


Figure 3—A typical 1750-meter transmitting antenna system (A). The radiator is topped with a "capacitance hat" and secured with several guy wires. A glass bottle insulates the base. The loading coil (B) connects between the radiator and the ground system with the feed line attached to the coil taps. The length of the radiator and feed line must conform to FCC Part 15 Rules. ("The total length of the transmission line, antenna and ground lead, if used, shall not exceed 15 meters.")

you have space limitations (performance will suffer greatly, though).

You can make the loading coil from a piece of *white* PVC pipe or Plexiglas tubing approximately six inches in diameter (see Figure 3B). Use #18 enamel-insulated wire, winding it tightly on the pipe to a length of five inches. Create taps at the cold end of the coil by sanding away the insulation of the first 10 turns. The braid of your coaxial feed line is attached to the ground system and cold end of the coil. The center conductor is soldered temporarily to the fifth turn of the coil. The top of the coil attaches to the antenna. Several small NE-2 neon bulbs can be soldered in series and used as a voltage detector. While transmitting a carrier, touch the bulb string to the antenna and watch for an indication. Place your hand near the antenna and note the illumination. If the capacitance of your hand makes the bulbs brighter, *add* more turns to the coil; remove turns if the bulbs dim. Once the best number of turns is determined, experiment further to find the best tap point. When you're finished, paint the coil with clear marine varnish.

It's important to note that there are no restrictions on the type and size of your *receiving* antenna. With that in mind, you may want to consider separate receiving and transmitting antennas. For example, many Lowfers use a broadband high-impedance probe with a built-in preamp called an *active whip* antenna. The small size of the whip antenna makes it convenient for it to be placed in a location where noise is at a minimum. A separate ground rod should be used as a noise-free isolated ground reference for this antenna. Active whip antennas are considered the best overall LF/VLF receiving antennas and can be purchased from LF Engineering, 17 Jeffery Rd, East Haven, CT 06513; tel 203-248-6816.

Lowfers also use loop antennas because of their ability to reduce or eliminate noise by rotating the antenna *null* for minimum noise pickup. Magnetic noise from residential wiring is a problem in city or suburban environments, so a loop is not recommended for these areas. But in a clear area such as a park, a loop can effectively remove strong carriers and noise that block the signal you are trying to receive.

Receiving and transmitting antennas should be located away from power lines, trees or buildings. Noise may wipe out reception completely, discouraging the casual listener from using this band. Smart antenna design and placement will significantly reduce the noise to an acceptable level. Man-made noise tends to rapidly diminish as you move away from the source. Light dimmer interference, for example, will radiate throughout the wiring in the home, eliminating useful reception from an antenna on the roof, or within 20 feet of the home. If the receiving antenna can be moved near the curb, however, the noise may disappear.

Conclusion

Much more needs to be said about Lowfer antennas and Lowfing in general, but I'd need a lot more space! The CW-893 kit manual includes a detailed discussion of antennas with several suggested designs. I also recommend that you pick up a copy of Ken Cornell's book, *The Low and Medium Frequency Radio Scrapbook*. It's available by mail directly from Ken at 225 Baltimore Ave, Point Pleasant, NJ 08742. The cost is $17.50 (shipping included). You'll find a partial list of 1750-meter CW beacon stations in WB8IMY's article, "Lowfing on 1750 Meters" in the October 1993, *QST*, page 67.

Lowfer newsletters are also available. If you join the Longwave Club of America, you'll receive the *Lowdown*. For details, write the LWCA at 45 Wildflower Rd, Levittown, PA 19057.

David Curry, WD4PLI, has been an active Amateur Radio operator and Lowfer for more than 20 years. He is a regular on the Los Angeles area 1750-meter SSB Net Saturday mornings at 9 AM on 183.5 kHz. He also participates in the 80-meter Lowfer Net Sundays at 8 AM on 3927 kHz. David is a professional musician and enjoys snow skiing, wind surfing and fine dining.

New Books

SATELLITE COMMUNICATIONS AND DBS SYSTEMS

By James Wood, G3VG

Focal Press, 80 Montvale Ave, Stoneham, MA 02180; tel 617-438-8464. First edition, hardcover, 6 × 9 inches, 277 pp. $79.95.

Reviewed By Steve Ford, WB8IMY
Assistant Technical Editor

Can you condense the history and current technology of satellite communications into a relatively small book? James Wood, a licensed amateur since 1938, a Chartered Engineer, and Fellow of the Institution of Electrical Engineers, has made a worthy effort in *Satellite Communications and DBS Systems*.

The emphasis of this book is on commercial satellite systems, although Amateur Radio satellites are mentioned briefly. In particular, the author concentrates on the delivery of television signals via satellite. He describes commercial satellite systems in detail, but manages to do so without diving too deeply into arcane technospeak. Instead, he takes a highly specialized subject and makes it interesting.

Satellite Communications and DBS Systems is full of unique tidbits of information. For example, the Russian *Proton* SL-13 rocket's first stage develops 334,000 pounds of thrust and burns for 130 seconds. Chapter 16 lists every commercial broadcast satellite presently in orbit, with a summary of each one's capabilities.

A chapter is devoted to the subject of high-definition television (HDTV). Many in the broadcast community believe that direct-broadcast satellites (DBS) are prime candidates to supply HDTV programming to consumers. Wood also discusses a new multiplexed analog component (MAC) television standard. It's part of a family of television signal standards designed to eliminate the deficiencies in the existing PAL, SECAM and NTSC formats. The proposed MAC/packet standard can accommodate high-fidelity digital audio and HDTV. It's presently under consideration as a worldwide DBS standard.

The only deficiency of the book, from an American point of view, is its emphasis on European systems. For example, Wood, who lives in Bracknell, England, devotes a number of pages to a discussion of the Astra and Eutelsat satellites. Considering their strictly European signal coverage, this is of limited interest to us "Yanks." The same is true of his chapters on PAL and SECAM broadcast formats and video recording techniques.

Although it's not primarily an Amateur Radio-oriented text, the technology is fascinating and *Satellite Communications and DBS Systems* is worthwhile reading if you're a satellite TV enthusiast or if you'd like a glimpse of the future of direct satellite broadcasting.

A Trigonal HF Beam

Not an April Fool article! A directional HF gain antenna that doesn't need a rotator.

By Dick Bird, G4ZU/F6IDC
Malves 11600, France

Rick, a local CBer, was interested in operation on the 10-meter band once he got his amateur license. He asked me whether it would be possible to retune his ground-plane vertical from 27 to 28 MHz, or use pieces of it to make a 10-meter beam. Pulling out my pocket calculator, I found that the radials of a 27-MHz ground plane were about 3 to 4% longer than a quarter wavelength at 28 MHz, so it seemed that two of his radials butted together would be just about the right length for the reflector of a 28-MHz beam. With a little further thought I realized that all we needed to make a 28-MHz beam was to hang a simple wire dipole between two of the radials. The two radials support the dipole act as a (bent) reflector, as shown in Figure 1. All Rick would need to do is extend the radials with short fiberglass rods (about 6 inches) so that the dipole could hang more-or-less horizontally. The ground plane itself could be left intact for CB use.

Bent Reflector

Rick was skeptical about the use of a bent reflector, so I suggested he read my article about the "Jungle Job."[1] I told Rick he'd need a rotator to turn the beam for 10-meter use. The next day, I realized that even that problem could be overcome, by using the three existing radials to support three separately fed wire dipoles. This scheme provides three independent 10-meter beams that can be selected to provide almost complete 360° coverage (Figure 2).

I found this arrangement quite attractive as, although it requires three separate feed lines, it doesn't require a rotator. All the same, I felt that further study was required, to ensure there were no harmful interactions. From initial tests with a 6-meter model I knew the design was viable. I optimized the dimensions with the aid of the MN 4.0[2] program, which indicated potential gain of nearly 9 dBi, with a 30-dB front-to-back ratio!

On 80 and 40 meters, many DX operators rely on a couple of suitably phased coax-fed quarter-wave verticals. If the feeder lengths are correctly chosen, this can provide both forward gain and useful front-to-back ratio. By switching in a delay line, you can also reverse the direction of fire.

There are three main disadvantages of such an installation:

- Feed lengths are fairly critical
- Lack of 360° coverage
- A large number of radials is required for good performance

Using Dipoles Instead

To overcome the need for radials and to provide coverage over a wider range of azimuths, a number of radio amateurs have experimented with beams using horizontal dipoles, rather than quarter-wave verticals. For example, given three suitable support points, such as a couple of trees and a house chimney, a center-fed wire dipole could be backed up by a pair of similar-length reflector elements (Figure 3A). This antenna would provide a forward gain similar to that of a couple of phased verticals, plus quite useful discrimination against unwanted signals arriving from the sides or the rear.

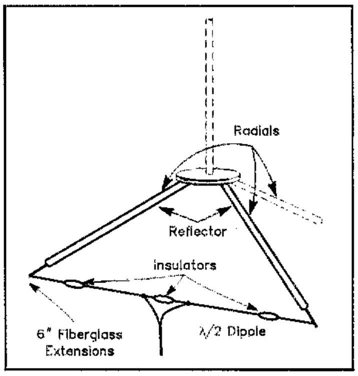


Figure 1—A 10-meter dipole suspended between radials on an 11-meter ground plane. The radials are just about the perfect length to act as a reflector for the dipole.

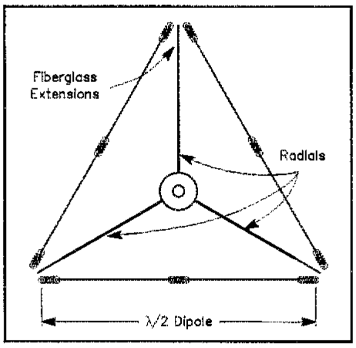


Figure 2—Bird's-eye view of the Figure 1 antenna, with three dipoles in place.

An even more interesting approach would be to use three driven dipoles in a very similar formation (Figure 3B). The dipoles not in use form the reflector for the dipole being fed. For the antenna shown in Figure 3B, you can see that by switching feed lines you can easily aim the signal in 120° steps. This method has the additional advantages of being faster and cheaper than a rotary Yagi on a tower. Antennas of this type are used to good effect by hams all over the world.

Just as with phased verticals, however, it is essential that all three feeders be cut to precise electrical lengths, to keep unused feeders from detuning their respective reflectors. I was tempted to experiment with the Figure 3B design, but I decided to first make a detailed computer analysis, to determine optimum phase delay and appropriate feeder lengths.

In the course of this study I found that, quite apart from feeder length problems, the system had many other somewhat unexpected shortcomings. For example, to make the reflectors fully effective, they really ought to be 3% to 5% longer than the driven element, and much more closely spaced than in Figure 3B. Even then, the feed lines would still need to be cut to very precise electrical lengths. After examining a number of alternatives, I finally settled for the rather curious three-wire reflector system shown in Figure 4. You easily can see the similarity to the antenna I worked up for Rick (Figure 2).

As far as I am aware, this "trigonal reflector" is entirely new. It is, in effect, three half-wave V-shaped reflectors coupled back-to-back. Each section embraces an arc of up to 120° in azimuth. For

[1] Notes appear on page 34.

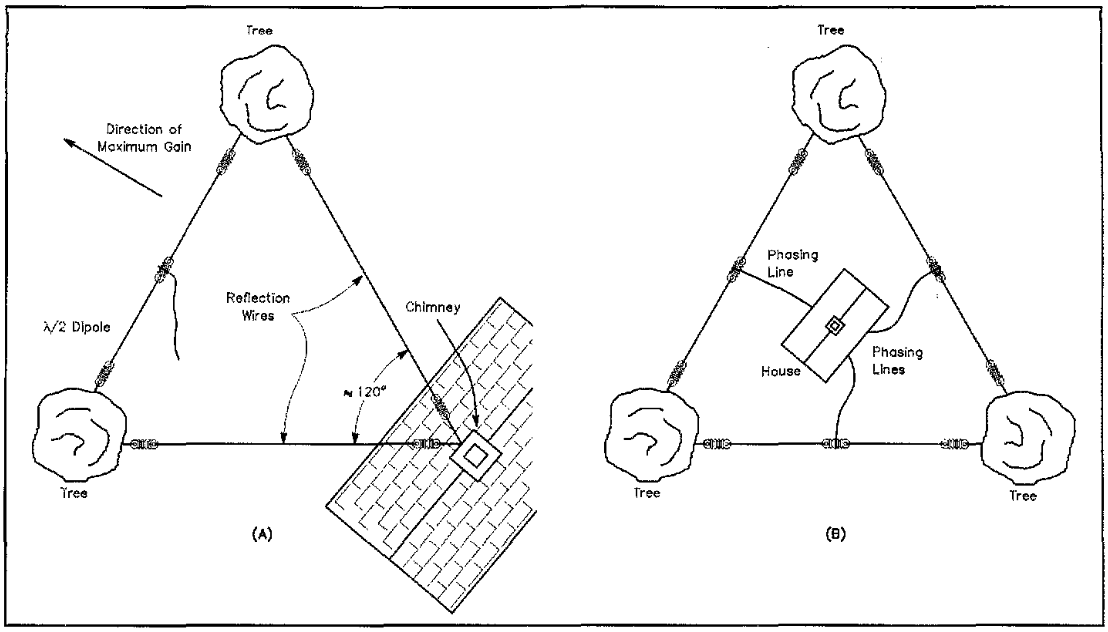


Figure 3—A large-scale arrangement good for a single arrangement is shown at A. B shows an array of phased dipoles suspended from trees. It helps to live in the woods. The disadvantage of the arrangement at B is the difficulty of properly phasing the dipoles.

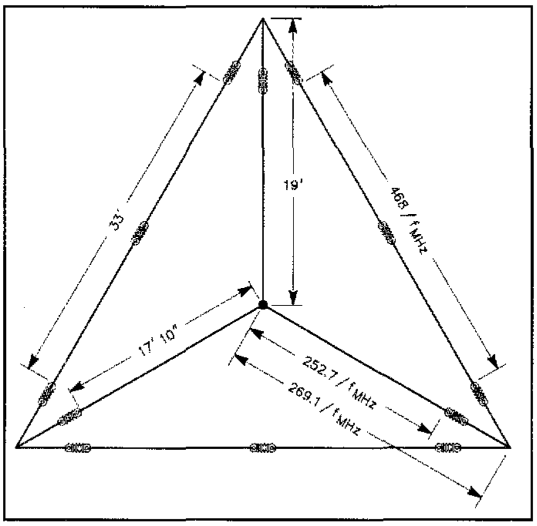


Figure 4—G4ZU's final design uses three dipoles and a trigonal reflector to switch the antenna pattern in three directions. The reflector can be made of wires or aluminum tubing. The dipoles are standard ½-λ designs. Each leg of the reflector is about 8% longer than ¼-λ. If you use a rigid reflector or three common supports for the dipoles and a wire reflector, you'll need to add some insulating material to the reflector tips to get the ends far enough apart so that the dipole fits. Generic dimensions are given in wavelengths so you can scale the design to any band. The specific dimensions shown are for 20 meters, and the performance of this antenna is shown in Figure 5.

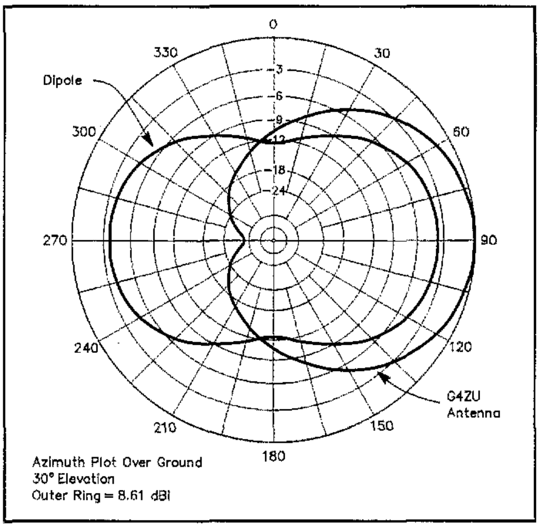


Figure 5—Azimuth plot generated by *ELNEC*, an antenna-modeling program, shows the pattern of the G4ZU Trigonal Beam antenna compared to a conventional dipole. This plot assumes the other dipoles are in place but aren't in use. This is a 20-meter version 25 feet above ground.

best results, each leg of this trigonal reflector must be cut to a length of 1/4 wavelength plus 3%. After they are soldered together at the center, the tips can be secured to appropriate support points.

Using the same support points, I then strung three half-wave center-fed wire dipoles around the trigonal reflector. With three feed lines, this arrangement gave me radiation over three independent arcs, each of 120°, any one of which could be selected at the flip of a switch. What's more, the feed lines can be of any random lengths, because the active segment is independent of and completely screened by its reflector from the two other segments that aren't in use. The suppression of radiation between adjacent segments was, in fact, found to be more than 30 dB.

If you'd prefer a more sophisticated look, a trigonal reflector could be fabricated from tapered aluminum tubing and mounted on a single mast.

Technical and Performance Notes

If you seek a very high front-to-back ratio, the spacing between the tips of each radiator and its reflector can be adjusted in accordance with the "critical coupling" techniques outlined in the article cited in Note 1. My first experimental model had a forward gain of nearly 9 dBi and a front-to-back ratio of more than 34 dB.[3] Its performance is better than some conventional 3-element Yagis. The lobe width of each section is broad enough to cover an entire continental area, so by switching from one bay to another, you have almost 360° coverage. Figure 5 is a polar plot of this antenna as modeled with the *ELNEC* program.

At a modest height of only 25 feet above ground, the measured feed impedance was 55 Ω, and the SWR was less than 1.2:1. Even more gain should be available if the antenna is 1/2 wavelength or more above ground. These measurements were made on the 20-meter band, although the trigonal beam should be even more attractive when constructed for use on 40 meters. On 40, only the lucky few can manage a Yagi and the necessary tower and rotator. A G4ZU Trigonal Reflector beam may be more manageable.

Notes

[1] G. A. Bird, "New Techniques for Rotary Beam Construction," Hall, ed., *The ARRL Antenna Compendium Volume 2* (Newington: ARRL, 1989, pp 58-60.

[2] *MN4.0 Antenna Analysis Software,* by K6STI, available from Brian Beezly, K6STI, 507½ Taylor, Vista, CA 92084.

[3] ARRL Technical Advisor James W. ("Rus") Healey, NJ2L, used another antenna modeling program, *ELNEC*, to study the G4ZU antenna. He obtained similar gain-over-isotropic values, but lower F/B ratios (on the order of 22-25 dB), which are still quite respectable. *ELNEC* is available from Roy Lewallen, W7EL, 5470 SW 152 Ave, Beaverton, OR 97007.

Strays

QST de GALLERY

◊ Our thanks to Robert M. Gallery, of Gaithersburg, Maryland, for his recent donation of *QST*s from the 1920s and '30s. They are from the collection of his father, Robert A. Gallery, W3CTQ, who died in 1954. The senior Gallery had been commander of the carrier USS *Randolph* under Admiral Halsey. These issues will be used to replenish our stock of back issues for in-house use.

New Books

WHAT IS YOUR TNC DOING?

By Gloria Medcalf, KA5ZTX

ZM Expressions, 1544 N 1000 Rd, Lawrence, KS 66046-9610; tel 913-842-6808. 1993, 121 pp; B&W diagrams and tables; 6×9 inches. Retail $15 from dealers or directly from ZM Expressions.

Reviewed By Steve Ford, WB8IMY
Assistant Technical Editor

In the legion of beginner-level packet books, there's one subject that isn't always given the treatment it deserves: how a packet terminal node controller, or *TNC*, actually *works*. To many amateurs, TNCs are black-box devices. They understand that TNCs are the nerve centers of their packet stations, but that's about as far as it goes. They plug in their cables, load their software (or fire up their terminals) and they're on the air.

Understanding the inner workings of your TNC gives you a valuable edge in the packet radio world. You can use this knowledge to optimize your TNC and make it much more efficient on the air. You can interpret those odd messages you see from time to time and fix a potential problem.

Gloria Medcalf, KA5ZTX, gets you started at ground zero in *What is Your TNC Doing?* She explores this potentially deep subject in small, easy-to-understand steps. Gloria doesn't assume that you're a computer expert. In fact, she begins with the assumption that you're as new to your computer or data terminal as you are to your TNC. Chapter 1 (The Computer) begins with, "Voltages change and things happen. That's how a computer works." You can't get more elementary than that!

The learning curve ramps up quickly. Soon you're tackling the typical voltage/signal format in an RS-232 serial cable. That's the umbilical between your computer or terminal and the TNC. There are wiring diagrams for several types of serial cable connectors.

Once you're connected to your TNC, you have to talk to it. To this end, Gloria describes terminal software and typical communication settings (stop bits, parity and so on). This is a potential stumbling block for new packeteers and she does a fine job of leading readers through it.

Chapter 4 finally gets into the meat of the issue: the TNC itself. Without overwhelming you with technospeak, Gloria explains the operation of this complex device. The hookup diagrams alone are a major asset. The book illustrates typical speaker and microphone plug diagrams for several transceivers, including a number of hand-helds. Ample time is spent discussing TNC and radio settings. For example, transmit/receive switching time is explained in detail.

Chapter 6 is devoted to sample packet conversations. The emphasis is on what packet users may see on their screens during live, keyboard-to-keyboard chats. Although still aiming at the beginner, this section may be tough sledding for some. There's a lot of discussion about the frames exchanged during packet communication. The average user won't see most of these frames, but the discussion is worthwhile. At least the beginner will have a better sense of what's going on during a conversation, whether it's with a bulletin board or a live operator.

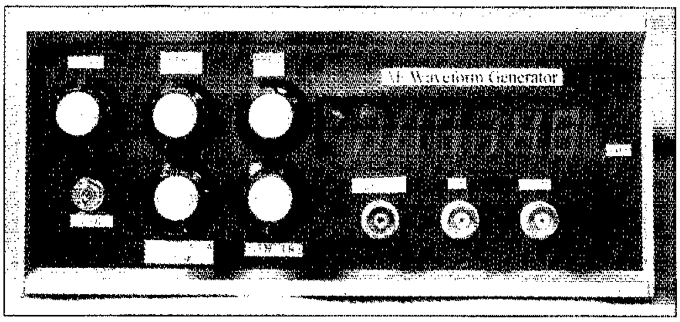
What is Your TNC Doing? ends with a convenient troubleshooting guide. This may be the most valuable chapter of the book for beginners. For example, some may wonder why their TNCs refuse to key their transceivers. How many will think to check the XMITOK command? As the guide points out, if XMITOK is off, the TNC won't switch to the transmit mode.

What is Your TNC Doing? lacks operational detail. For instance, there's no discussion of nodes and specialized networks such as DX *PacketClusters*. Packet bulletin boards aren't described at all. These are things beginners will face from the moment they activate their TNCs.

But *What is Your TNC Doing?* isn't an operational packet book. It concerns itself strictly with the nuts and bolts of TNCs and the packet protocol itself. For beginners who want to expand their understanding of how packet really works at the hardware and software level, *What is Your TNC Doing?* is an excellent choice.

A Function Generator with a Frequency-Counter Digital Readout

"Low cost," "easy to build" and "functional" describe this project. There's no ragged frequency guesstimation by using the knob-pointer-on-panel-mark method, either! It uses an honest-to-goodness frequency meter!



By Ben C. Spencer, G4YNM
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Swindon, SN2 2BN
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Photos by the author

Of the many pieces of test equipment Amateur Radio experimenters and operators alike should have in their shacks is one that can produce audio test signals. A sine-wave generator is the mainstay, with a square-wave generator now being just about as important what with all the digital circuitry surrounding us. A *function generator* is a box that produces more than one type of waveform that can be varied in frequency and amplitude.

General Description

This low-budget function generator is based on the tried and true Intersil ICL8038 waveform generator IC. A useful bench instrument, it provides sine, square and triangle waveforms covering frequencies from 10 Hz to 80 kHz in four ranges. (Table 1 provides a list of the generator's specifications.) A built-in frequency counter with a 6-digit, 7-segment LED display shows you the waveform's frequency. The square wave is fed directly to the front-panel output connector without buffering or amplification. The sine and triangle waveforms are routed to a dual op amp that provides one 600-Ω output and one low-impedance output that simulates an ideal voltage source.

Table 1
Function Generator Specifications

Frequency:
Range 1: 10 Hz to 100 Hz;
Range 2: 100 Hz to 1 kHz;
Range 3: 1 kHz to 10 kHz;
Range 4: 10 kHz to 80 kHz.

Distortion:
Sine wave: 1% (when correctly adjusted).
Triangular wave: 0.1% linearity (10 Hz to 10 kHz).

Output Levels:
Sine wave: 20 V P-P (0-dB variance from 10 Hz to 100 kHz).
Triangular wave: 20 V P-P (10 Hz to 20 kHz).
Square wave: 20 V P-P (0-dB variance from 10 Hz to 80 kHz from the DIRECT output).

Output Impedances:
Square: 200 Ω (with a load drawing less than 5 mA).
Low: Approximately less than 1 Ω.
600 Ω: Approximately 600 Ω.
Duty cycle: 50%.

Many of the components can probably be found in your junk box or purchased from any of the mail-order supply houses. Smart shoppers can bring in this project (with enclosure and PC boards[1]) for about $50. Solely because an oscilloscope is the best way to test and calibrate the unit, I rate this project at the intermediate level.

Circuit Description

We can divide this project into three blocks, the power supply (Figure 1), waveform generator (Figure 2) and the digital fre-

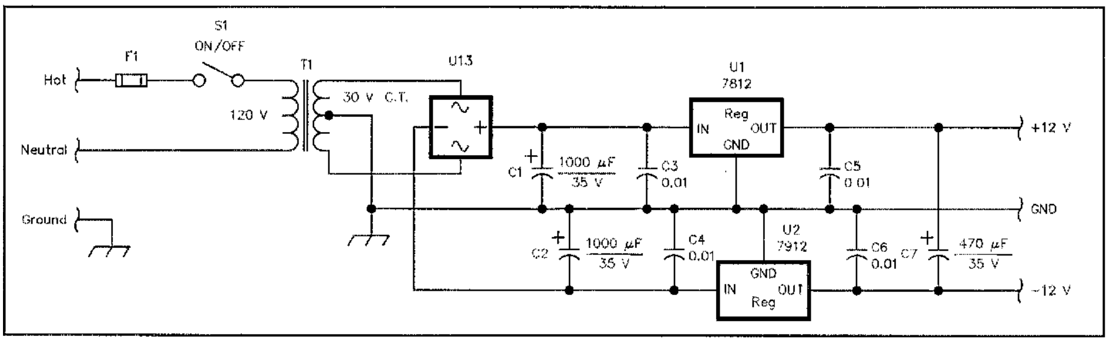

Figure 1—Schematic of the power supply. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

F1—125-V, 1.5-A fuse.
S1—SPST toggle switch.
T1—120-V primary, 30-V, 1-A center-tapped secondary.
U1—7812, positive 12-V, 1-A regulator, TO-220 case.
U2—7912, negative 12-V, 1-A regulator, TO-220 case.
U13—50-V, 1-A bridge rectifier.

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[1] Notes appear on page 39.

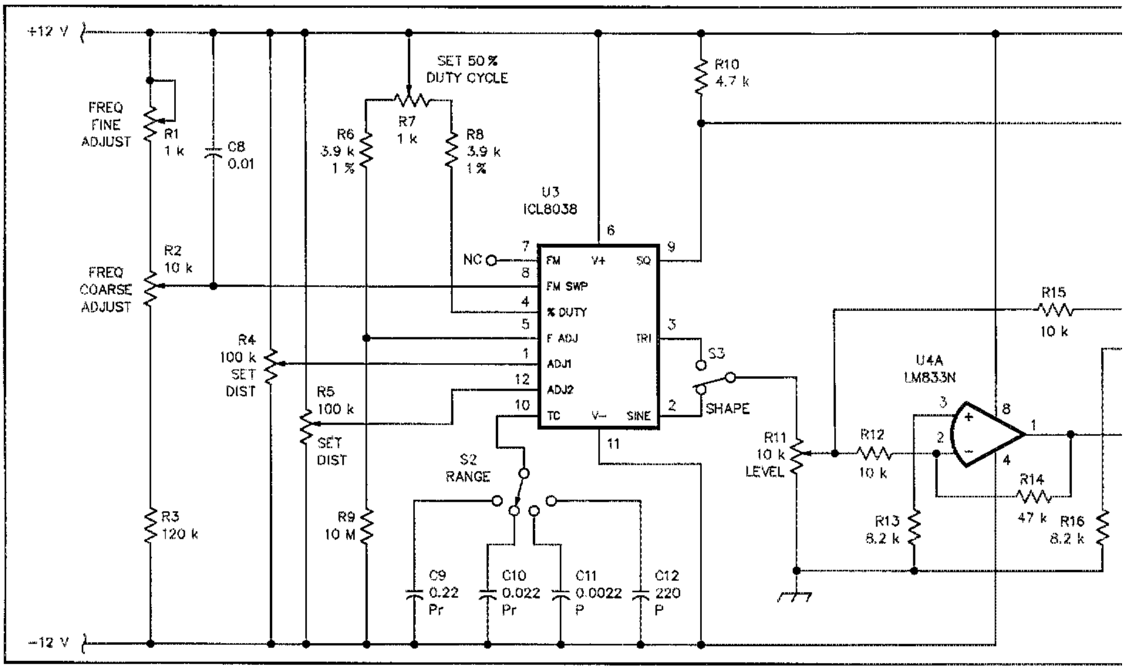


Figure 2—Schematic of the AF waveform generator. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

R1—1-kΩ linear-taper, panel-mount potentiometer.
R7—1-kΩ PC-mount trimmer potentiometer.
R2, R11—10-kΩ linear-taper, panel-mount potentiometer.
R4, R5—100-kΩ PC-mount trimmer potentiometer.
R6, R8—3.9 kΩ, 1%-tolerance.
S2—Single-pole, 4-position rotary switch.
S3—Single pole, 2-position rotary switch.
U3—ICL8038 function generator.
U4—LM833N dual op amp (available from Newark Electronics, which has many locations throughout the US; check your telephone book for a branch near you. Main office: 4801 N Ravenwood Ave, Chicago, IL 06040-4496, tel 312-784-5100, fax 312-784-5100, ext 3107).

quency counter (Figure 3). Some builders may find a use for the frequency counter itself in other circuits.

The Power Supply

Refer to Figure 1. The mains input passes through F1 and **ON/OFF** switch S1. T1 steps down the mains voltage and bridge rectifier U13 produces positive and negative voltages that are symmetrical about the ground reference.

Voltage regulator U1 produces the +12 V rail and U2 develops the −12 V rail. C7 is necessary to reduce a small spurious spike that would otherwise appear on the generator's sine-wave output (see C8 discussed later). The digital frequency counter uses the +12 V and 0 V rails only.

A single supply rail (ie, ground and +V) can be used for the ICL8038, however, as the IC output waveform is centered on |V/2|, a dc decoupling capacitor would be required. That would be undesirable because of the frequency sensitive nature of capacitors. By using +V and −V supplies, the waveform is centered on 0 V eliminating the need for an output decoupling capacitor.

Waveform Generator

See Figure 2. U3 is the function-generator IC. A triangle waveform is created by using internal current sources to charge and discharge a selected external capacitor (C9 through C12). The square wave is derived from the triangle waveform and finally, the sine wave is produced by feeding the triangle waveform into a nonlinear network.

U3 can be made to change its operating frequency by applying an ac sweep voltage to U3 pin 8. If, instead of ac, this sweep voltage is dc, then a simple means of varying the frequency becomes realistic. R1 and R2 are front-panel-mounted potentiometers that respectively provide **FINE** and **COARSE** frequency adjustment. C8 removes a small spurious spike that would otherwise appear on the generator's sine-wave output as a result of the nonlinear network (see C7, discussed earlier).

To obtain a 50% duty cycle, it's necessary to have equal values for R6 and R8, so R7 is included to trim the duty cycle, and R9 is included to reduce the variance of the duty cycle with frequency. Over large frequency excursions (ie, 10 Hz to 80 kHz), the duty cycle still varies slightly. If required, you could mount R7 on the front panel.

As the sine converter part of U3 comprises a nonlinear network, the sine-wave output is an approximation of a pure sine wave. R4 and R5 are included to minimize the sine-wave distortion during the initial set up.

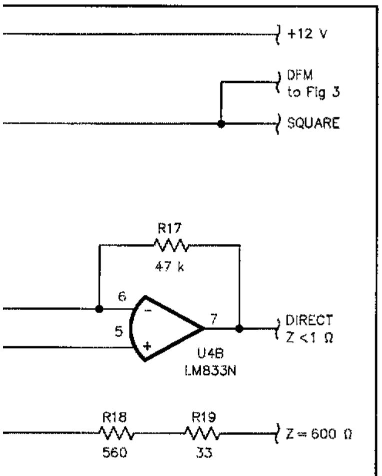
C9, C10, C11 and C12 determine the unit's four frequency ranges. These ranges overlap slightly, but are nominally: 10 Hz to 100 Hz; 100 Hz to 1 kHz; 1 kHz to 10 kHz; and 10 kHz to 80 kHz. S2 (**RANGE**) selects which one of the four capacitors is connected to U3 pin 10.

The sine and triangle waveforms from U3 are fed to S3 (**SHAPE**) which selects which of the two waveforms is fed via R11 to buffer-amplifiers U4A and U4B. Because these buffers are identical in all respects except for their output impedances, I'll describe only one buffer.

U4's gain is set to approximately 5 (14 dB) by R12 and R14. R13's value is chosen to be the parallel equivalent of R12 and R14, reducing the op amp's dc output offset voltage to about 0 V.

The output impedance of U4A is very low, effectively acting as a reasonably ideal voltage source. A common impedance used in audio circuits is 600 Ω, so R18 and R19 bring the equivalent output impedance up to 600 Ω. The output of U4B is left as a low-impedance source as this can be equally useful on the workbench.

U4's slew rate limits the output waveform rise and fall time. As U4 cannot switch instantaneously between the voltage rails, as the operating frequency increases, the tri-



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

IC pins not shown are unused
NC=Not Connected
P=Polystyrene
Pr=Polyester

angle-wave output becomes distorted; that is, the triangle waveform develops a squiggle. A higher-performance (read higher cost) dual op amp could be used to improve the top-end operation, but because most of the audio work I do is below 20 kHz, the existing op amp stayed!

The square wave output remains a true square wave over the unit's frequency range, but the output impedance is about 200 Ω (with a load current of less than 5 mA). The square-wave output from U3 pin 9 is an open-collector output, so R10 is used to provide a load. This output is also tapped to provide an input for the digital frequency counter circuit (to be described).

If you need to have a variable control for the square-wave output, you could use a dual-gang potentiometer for R11. Use one of the potentiometers to control the sine and triangle waveforms and the second potentiometer to control the square-wave output. I chose not to do this.

Digital Frequency Counter

See Figure 3, the digital frequency counter schematic. The time base oscillator is provided by U5, which is configured as an astable multivibrator with the asymmetrical waveform output at U5 pin 3.

The time high is

$t_H = 0.693(R21 + R22 + R23)C13$ (Eq 1)

and the time low is

$t_L = 0.693\,(R22 + R23)C13$ (Eq 2)

This means that the output of U5 is high for a longer time than it is low.[2]

R22 provides for adjustment of the time base to ensure that t_L can be set to 1 second (more on this later) and t_H is approximately 2 seconds. Input-signal noise reduction is performed by Schmitt trigger gate U6A.

The frequency counter is formed around 4026 decade counter chips which directly drive the six 7-segment LED displays. A description of these chips is in order. Each time a clock pulse is received at the CLK input, the number shown on the respective LED display is incremented, after ten pulses have been received at the CLK input, the displayed number reverts to 0 and the carry-out pin (CO) becomes active.

The 4026 can be made to stop counting by pulling the CK EN pin low, and the display can be blanked by pulling the display enable input (DEI) pin high. Finally, the 4026 internal counter can be reset to 0 by pulling the reset pin (RST) high, if the CK EN is disabled and then enabled again without a reset, the count simply continues from where it left off.

The magnitude of the maximum count can be increased by connecting the first 4026's CO output to the next 4026's CLK input. Hence, with two 4026's, it is possible to count to 99; with three 4026's, 999, and so on. With six chips, the maximum possible count is 999999.

If the CK EN pins are tied together and active for precisely 1 second, the display readout is given in hertz, so the maximum *frequency* the meter can measure is 999999 Hz, or almost 1 MHz.

For the meter to operate correctly, these steps must occur in sequence:

1. Blank the display.
2. Reset the internal counters to ensure a measurement starts at zero.
3. Enable the counters for 1 second.
4. Disable the counters.
5. Enable the display to show the last measurement.
6. Wait a reasonable period to allow the operator to absorb the information displayed.
7. Repeat from Step 1.

Assuming the output at U5 pin 3 is high, the CK EN and DEI pins will also be high (via U6B and U6D), so counting is disabled and the display enabled, showing the last measured frequency. C14 blocks the output from U5 pin 3, so the RST pins are pulled low by U6C.

As U5 pin 3 goes low, a number of things happen: A very short pulse appears across C14, causing U6C to strobe the RST line which resets the internal counters of U7 to U12, this is delayed by the inherent propagation delay of U6C. The CK EN and DEI pins of U7 to U12 are pulled low, this disables the display and enables counting to commence, this action is delayed by the inherent propagation delays of U6B and U6D.

This does not conform precisely to the states originally given. However, by using the gates in this manner, it ensures that a reset occurs before a new count is initiated. In any case, the response time of the LED displays is such that the theoretically displayed 000000s are never noticed by the user.

Additionally, the count-time duration could be off by two propagation delays, which would affect the counter accuracy. This problem is eliminated by adjusting R22 so that the low time of U5 pin 3 is actually equivalent to one second plus two propagation delays. Again, you never notice this, as R22 is simply adjusted until the correct frequency is displayed.

After the 1-second count duration, U5 pin 3 goes high again. This time, no pulse appears across C14, as the U6C side of C14 is already high, so a reset strobe does not occur. Instead, the CK EN and DEI pins of U7 to U12 are pulled high, thus disabling further counting and enabling the current values stored in U7 to U12 to be displayed for a time period of t_H.

Hence the overall action of the counter is:

1. Blank the display for 1 second while measuring.
2. Display the frequency for 2 seconds;
3. Repeat from Step 1.

If the input count is greater than 999999, the CO output of U12 becomes active. DS4 indicates an overflow condition.

Finally, the frequency in hertz is "converted" to kilohertz by simply ensuring that the decimal point segment of the appropriate LED display is tied to the positive rail via R31.

Construction

Single-sided PC boards are available for this project (see Note 1). Several jumpers need to be installed on the waveform-generator board. These jumper locations are marked by the dotted lines on the component overlay (see Note 1).

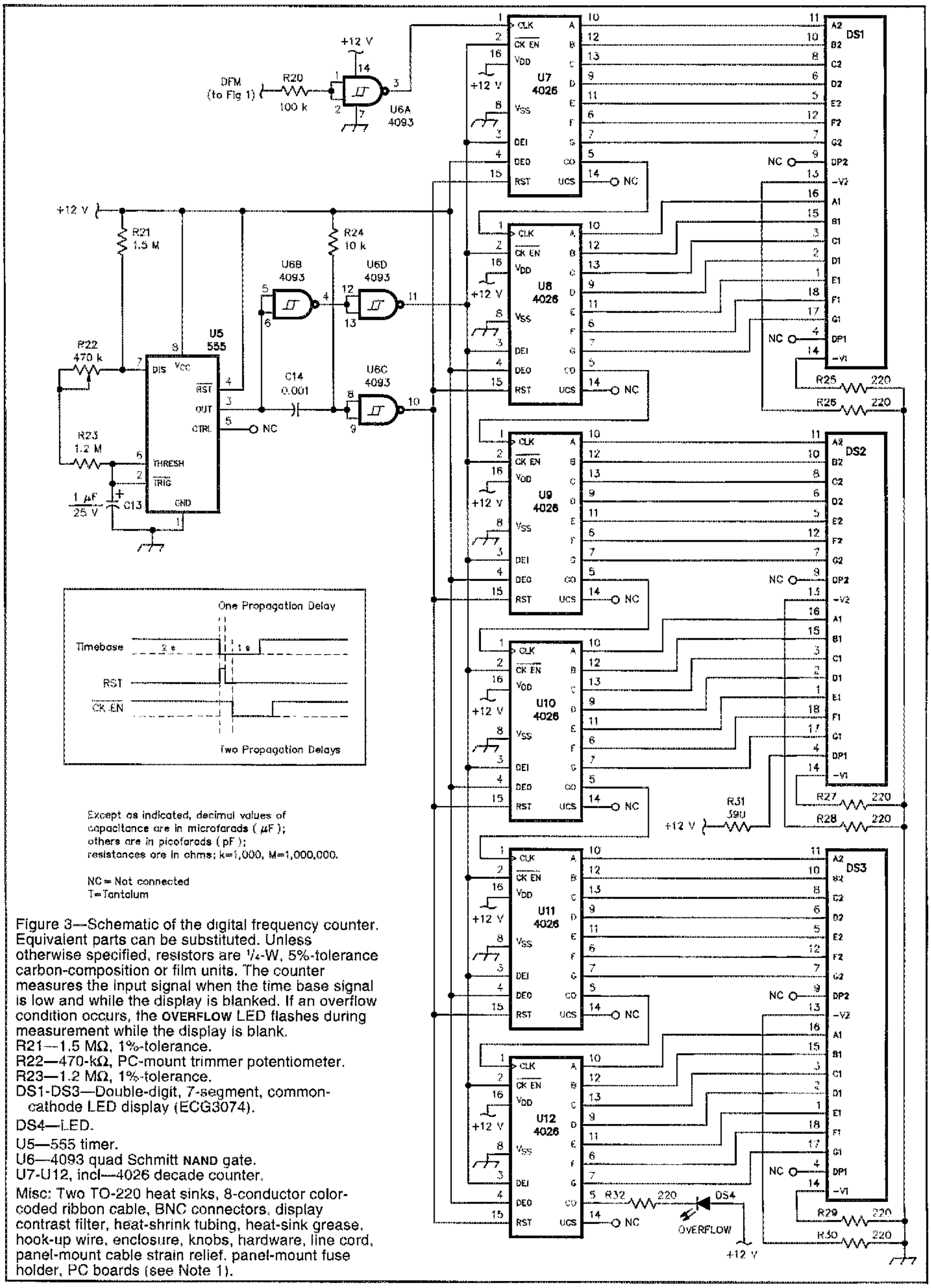
My prototype is shown in the title-page photo. Note that the prototype used separate boards for the waveform generator and digital frequency counter.

The following components are mounted on the front panel: R1, R2 and R11, S1, S2, S3 and the three output connectors. The connectors can be whatever suits your needs. I used BNC connectors because I have a box full of them and they make connects and disconnects quick and easy.

The fuse holder and ac-line cord are rear-panel mounted. For safety reasons, route all ac mains wiring connections and terminals to one end of the case and cover exposed connections with heat-shrink tubing. Route all dc wiring away from the mains wiring—ideally at the other end of the case. House the instrument in a metal case and connect the case to the mains earth. I used metal standoffs to mount the PC board, thus completing the ground circuits.

Fit U1 and U2 with TO-220 heat sinks (or home-brewed equivalents). *Do not allow the heat sinks to touch each other or anything else!*

The LED display PC board has six resistors (R25 to R30) mounted on the *solder side* of the board, this reduces the size of the board. The PC pins should be inserted from the



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

NC= Not connected
T=Tantalum

Figure 3—Schematic of the digital frequency counter. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. The counter measures the input signal when the time base signal is low and while the display is blanked. If an overflow condition occurs, the **OVERFLOW** LED flashes during measurement while the display is blank.
R21—1.5 MΩ, 1%-tolerance.
R22—470-kΩ, PC-mount trimmer potentiometer.
R23—1.2 MΩ, 1%-tolerance.
DS1-DS3—Double-digit, 7-segment, common-cathode LED display (ECG3074).
DS4—LED.
U5—555 timer.
U6—4093 quad Schmitt **NAND** gate.
U7-U12, incl—4026 decade counter.
Misc: Two TO-220 heat sinks, 8-conductor color-coded ribbon cable, BNC connectors, display contrast filter, heat-shrink tubing, heat-sink grease, hook-up wire, enclosure, knobs, hardware, line cord, panel-mount cable strain relief, panel-mount fuse holder, PC boards (see Note 1).

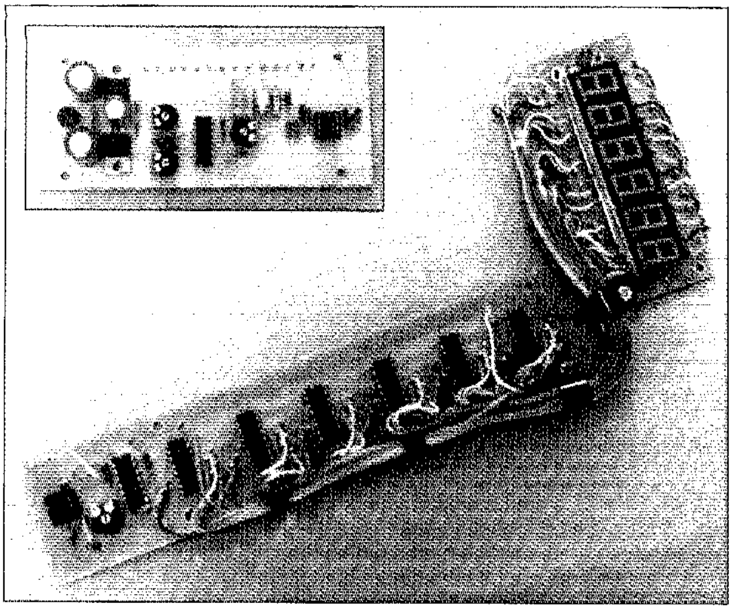


Figure 4—View of the prototype boards.

ponent side as this allows the on/off board wiring to be soldered on the "back of the board."

The main PCB and display PCB boards are interconnected by eight-conductor color-coded ribbon cable. There are two reasons for this approach: First, while UK readers will have no trouble obtaining the correct LED displays, some overseas builders may use a local supplier. In all probability, the local supplier's LED display pin-out will be different. Hence, the wire leads allow the use of almost any common-cathode display (even if it has to be built on perfboard). Secondly, by splitting the project into two separate boards, constructors can locate the display board on the front panel and place the main PC board wherever it suits them best.

Table 2 shows the interboard connections I used. On the main PC board, the PC pin labeled D0A indicates that this is digit 0, segment A, and so on. This corresponds with PC pin labeled 0A on the display PC board, and so on. Digit 0 is the least-significant digit (hertz); digit 5 is the most-significant digit (hundreds of kilohertz).

I first connected the ribbon cables to their respective decade counters (U7 to U12) and stacked the flat ribbon cables using **P** clips. So, when it came to joining the display PC board and counter PC board, it was easy to identify each digit and each segment within that digit.

Test and Calibration

To calibrate the function generator, you'll need a voltmeter and an inexpensive oscilloscope (with a bandwidth of 1 MHz). All measurements are made with respect to the instrument chassis (ie, 0 V is used as the reference ground).

Table 2
Interboard Wiring Convention

Wire Color	*4026/LED Interconnection*
Brown	Segment A to LED segment A.
Red	Segment B to LED segment B.
Orange	Segment C to LED segment C.
Yellow	Segment D to LED segment D.
Green	Segment E to LED segment E.
Blue	Segment F to LED segment F.
Violet	Segment G to LED segment G.

Apply the ac-line voltage to the unit and flip S1 to the **ON** position. With the voltmeter set to 30-V dc range, check for +12 V at pin 6 of U3. U3 pin 11 should show −12 V.

Set R4, R5 and R7 to their center positions; R1, R2 and R11 to their center positions and S2 to the **10-kHz to 80-kHz** range. The setting of S3 is irrelevant at this stage.

Connect a high-impedance oscilloscope probe to the square-wave output, (PC board pin E). adjust R2, R11 and the oscilloscope until a square wave of about 10 kHz is displayed. Adjust the **DUTY CYCLE PRESET** (R7) until a 50% duty cycle is obtained; you may need to adjust R1 (**FREQUENCY**) until this is accomplished.

Adjust R1 and R2 for precisely 5 kHz (200 µs as measured on the oscilloscope), adjust PC-board-mounted R22 until the front-panel digital frequency counter shows **5.000.** Alternatively, a bench digital frequency meter can be used instead of the oscilloscope for this alignment.

Connect the oscilloscope probe to the sine-wave output, (PC-board pin B) and *with the oscilloscope input set to dc coupling*, adjust R4 and R5 for the best sine wave you can obtain. In addition, you'll need to adjust R4 and R5 such that the waveform is centered around ground. This procedure may take a few minutes; it's a case of back and forth until you get a waveform you're happy with!

Essentially that completes the testing, except to check that the buffer amplifiers are operating correctly and that the four frequency ranges covered by the instrument are correct.

Summary

This function generator provides a sensible alternative to purchasing a commercial instrument costing ten times as much. Its electronic design complies with the philosophy of KISS (Keep It Simple, Stupid), and it should prove easy to build in a few bench-work sessions.

If you want to use the frequency counter section of this project for other projects—perhaps adding a MHz digit to enable it to be used with direct-conversion transceivers—the main PC board can be split, allowing the digital counter to be built separately.

Notes

[1] PC boards are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price for the three-board set is $16 plus $1.50 for shipping. The display board alone is $6, plus $1.50 for shipping.

A PC-board template package is available free from the ARRL. Address your request for the SPENCER FUNCTION GENERATOR TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

[2] Incidentally, the data books will tell you that you can *never* get t_H to be less than t_L with a 555 timer. Well, they're wrong! Here's a little diddle you may care to note for future reference. If you connect the CONTROL pin to ground the output inverts, that is, the formula for t_H swaps with that of t_L, which eliminates an inverter!

Strays

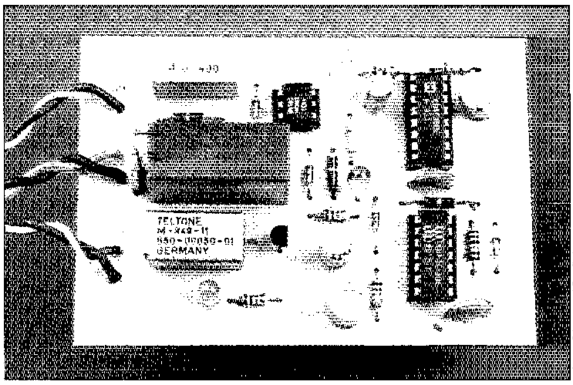
I would like to get in touch with...

◊ former shipmates (technicians, radio operators or crew) assigned to the US Coast Guard cutter *Pontchartrain* while ported in Wilmington, North Carolina, in 1972-73. Jay Chamberlain, KD4OOI (former ETN3), 27 Fox Run Ln, Fredericksburg, VA 22405.

◊ hams who served aboard the USS *Oriskany* (CV-34) before and/or during the Korean War. Harlow Beene, W5ZSL, PO Box 348, Glorieta, NM 87535.

The RingMaster Ring Detector

In some areas, Caller ID can mean reduced security for repeaters controlled by phone lines. Used in concert with one-line, multi-phone telephone service, this circuit preserves your control while keeping hackers out.



By Robin Rumbolt, WA4TEM
1202 Wilkinson Rd
Knoxville, TN 37923

The introduction of Caller ID by phone companies in many areas met with praise and criticism: praise for what it would let you see, and criticism for what it would let other people see. For Amateur Radio repeater operators, Caller ID presents a particularly alarming threat: It allows persons with Caller ID displays to make autopatch calls to themselves and see repeater phone numbers! For those of us who control our repeaters via phone lines, this opens the door for all kinds of hacker activity.

In many states, Caller ID can be blocked (for calls made from private numbers) by prefixing outgoing calls with *67, but here in Tennessee, our Public Service Commission does not allow the phone company to implement that option. So, we're stuck with a situation that would seem to allow any repeater's private number to be found out easily. Luckily, in our area—and likely in your area, too—there's a way to beat the system.

The answer is RingMaster[1] service, offered by the Bell telephone companies. This service allows several different phone numbers—a Main number and up to three RingMaster numbers—to be assigned to the same physical phone line. The ring timing associated with each number differs. Dialing the Main number rings its phone in the normal 2 seconds on, 4 seconds off sequence. Dialing a RingMaster number rings its phone in one of three distinctively different sequences, as shown in Figure 1.

The beauty of this is that when an *outgoing* call is placed, the receiving line's Caller ID unit displays only the Main number. None of the RingMaster numbers are ever displayed!

This article describes a circuit that senses the first RingMaster ring sequence.[2] When the circuit detects this sequence, it closes a relay and connects the phone line to the repeater equipment, which can then respond according to its usual programming. A call placed to the Main number—the number returned by Caller ID—results

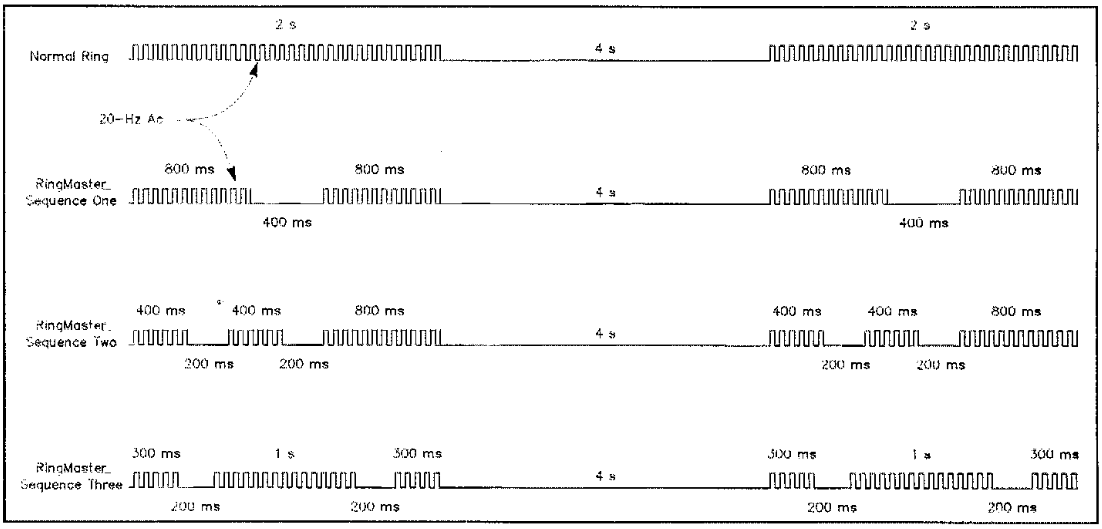


Figure 1—Four different ring sequences allow the use of up to four telephones on a single line. The RingMaster Ring Detector (Figure 2) responds to Sequence One.

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[1] Notes appear on page 42.

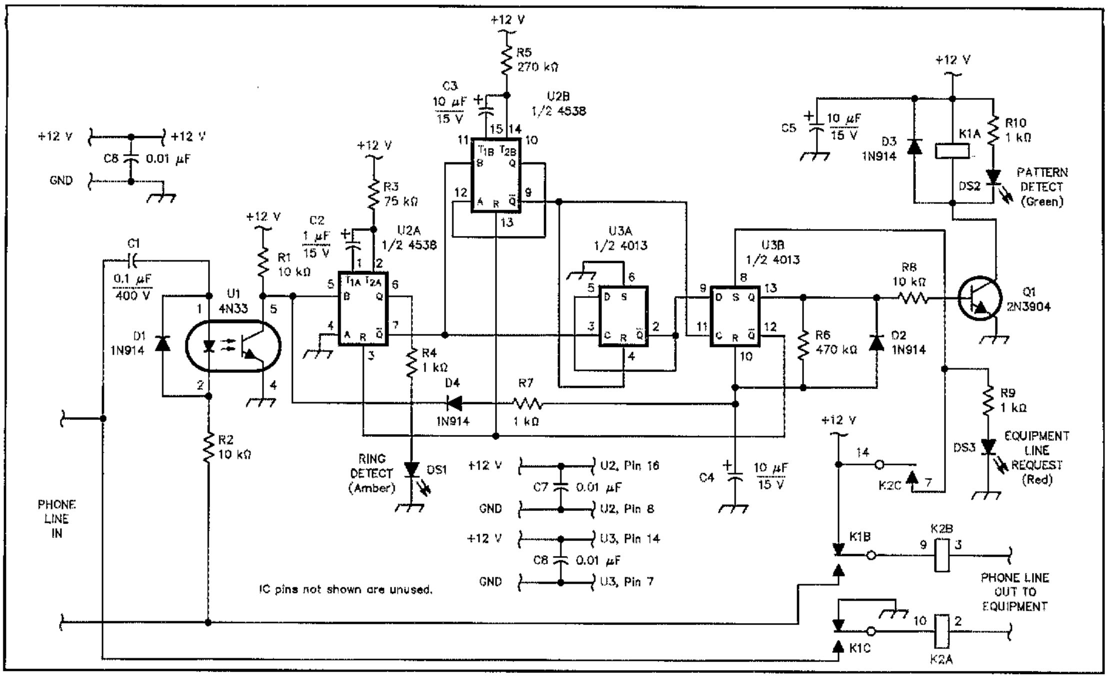

Figure 2—The RingMaster Ring Detector Circuit. Figure 3 shows the pinout for sensing relay K2.

C1—0.1-µF, 400-V Mylar.
C2—1-µF, 25-V electrolytic.
C3, C4, C5—10-µF, 25-V electrolytic.
C6, C7, C8—0.01-µF, 50-V disc-ceramic.
D1, D2, D3, D4—Silicon switching diode, 1N914 or equivalent.
DS1—Amber LED.
DS2—Green LED.
DS3—Red LED.
K1—DPDT relay, 12-V, 43-mA, 280-Ω coil. (Radio Shack #275-249 or 275-249A, or equivalent. The relays contained in packages marked *275-249* and *275-249A* are identical, but the pinout information shown on the package back is *incorrect* for parts marked *275-249* and *correct* [as implemented in the PC-board layout available per Note 3] for parts marked *275-249A*.)
K2—SPST line-current sensing relay. (Teltone M-949-11 or equivalent, available for $6.50 plus shipping in single quantities with no minimum order from High Technology Semiconductors, Suite 228, 3567 Benton St, Santa Clara, CA 95051, tel 408-246-5308, fax 714-544-4871.)
Q1—2N3904 or equivalent.
R1, R2, R8—10-kΩ, ¼-W, 5%-tolerance.
R3—75-kΩ, ¼-W, 5%-tolerance.
R4, R7, R9, R10—1-kΩ, ¼-W, 5%-tolerance.
R5—270-kΩ, ¼-W, 5%-tolerance.
R6—470-kΩ, ¼-W, 5%-tolerance.
U1—4N33 optoisolator.
U2—CD4538 DIP dual monostable flip-flop.
U3—CD4013 DIP dual-D flip-flop.

in a ringing, but always unanswered, line. Bingo! Freedom from hackers!

The Ring Detector Circuit

See Figure 2. U1, an optoisolator, isolates the circuit from the phone line and transforms the 100-V RMS ring pulses to 12-V logic pulses. U2A acts as a retriggerable monostable with a time-out period (≈75 ms) just slightly longer than the period of the 20-Hz ring pulses (50 ms). Once triggered by an incoming ring, it thus stays triggered as long as ring pulses are present.

U2B is configured as a non-retriggerable monostable with a period (≈2.5 seconds) greater than that of either the single ring (2 seconds) or double ring (0.8 second). U3A counts how many rings happen in the >2-second period timed by U2B. When U3A counts two rings and U2B times out, U3B triggers and turns on Q1, which turns on relay K1. This connects the repeater equipment to the line. U3B keeps Q1 turned on for about 5 seconds after the last ring pulse is received, and is reset by the network consisting of R6 and C4. The D4-R12 network discharges C4 with every ring pulse to keep U3B from timing out before the ringing stops. These networks make sure that the detector takes the repeater equipment off the line if it does not answer the phone or if the ringing stops.

If the detector senses the one continuous ring that occurs when the Main number is called, U3A doesn't count to two by the time U2B times out, U3B doesn't trigger, K1 doesn't activate, and the phone line remains unconnected to anything capable of answering it. The caller just hears a ringing line.

When a piece of repeater equipment places an outgoing call, it puts a low dc resistance across the terminals marked **PHONE LINE OUT TO EQUIPMENT**. This causes current to flow through K2A and K2B, making K2C close and applying 12 V to U3B's **SET** input. This turns on U3B, which in turn activates Q1 and relay K1. K1B and K1C then connect the repeater equipment to the outgoing phone line through K2A and K2B. This connection is maintained until the repeater equipment drops the line. Current flow through K2A and K2B then ceases and K2C opens, removing the voltage from U3B's **SET** input. The voltage at U3B's **RESET** pin then switches U3B, turning off Q1 and K1 and thereby disconnecting everything from the line.

K2, a Teltone type M-949-11, is a line-current sensing relay specifically made for this type of application. It's inexpensive and small, and has a 4-kV coil-to-contact isolation rating. I dislike construction articles that require weird, hard-to-get parts, but the Teltone relay is really necessary here, and I have found a distributor for it (see the Figure 2 part list) who will take credit-card orders via phone with no minimum order.

The detector includes three LEDs (DS1, **RING DETECT**; DS2, **PATTERN DETECT**; and DS3, **EQUIPMENT LINE REQUEST**) to show circuit status. DS1 (amber) lights as the phone rings, before the circuit detects the RingMaster sequence. DS2 (green) lights once the circuit detects the sequence,

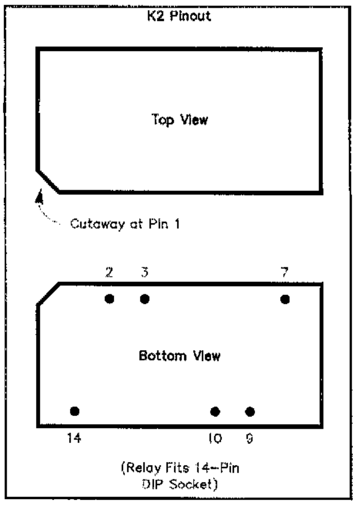

Figure 3—Pinout for the Ring Detector's Teltone M-949-11 relay.

More Than Just a Repeater Protector

I designed the Ring Detector to keep troublemakers safely isolated from sensitive equipment, but I'm sure you'll find many more uses for it. This circuit will not protect evildoers who try to use RingMaster service to cloak outgoing calls, because Caller ID returns the Main number for all outgoing calls made from RingMaster subnumbers. But the circuit does have other possible uses, including incoming and outgoing call restriction, isolating your home computer and/or fax machine from your main telephone number, ringing only your teenagers' phone on calls meant for them, and so on.

Leaving out U1 restricts activity to *outgoing calls only*, since incoming call ringing cannot be sensed with U1 absent. Calls that aren't detected won't be connected!

Removing K2 and putting jumpers in place of its coils (pin 9 to pin 3 and pin 10 to pin 2), and grounding pin 8 of U3B, restricts line activity to *incoming RingMaster calls only* since the Detector can no longer sense when a piece of equipment places an outgoing call and closes K2. To configure the circuit to accept *incoming calls only,* for the Main *and* Sequence One numbers, change C3 to 1 μF in addition to making the K2 and U3B changes.—*WA4TEM*

DS3 (red) may flicker whenever the repeater equipment receives ringing voltage, but shines steadily only when the repeater equipment answers or requests the line.

Construction

Since the Ring Detector deals only with low-frequency audio, its layout is noncritical. Building it on perfboard is fine, or you can use the PC-board layout I developed.[3] One caution, however: If your version will operate near a radio transmitter, the power pins of U2 and U3 must be properly bypassed for glitch-free operation. Solder the 0.01-μF power-lead bypasses shown in Figure 2 *directly across the IC supply pins* (16 and 8 of U2 and 14 and 7 of U3), keeping the capacitor leads as short as possible.

Checkout

To test the circuit, you must connect it to a line that has RingMaster service. Connect the Ring Detector to a phone line and apply 12 V dc. Wait for all of the Detector's LEDs to go out. Then have a friend place a call to your Main number. The normal ring should cause DS1, **RING DETECT**, to light with each ring-voltage burst. Nothing else should happen no matter how many times the phone rings.

Next, place a call to the RingMaster number. DS1 should light again with each ring. After the second quick ring, DS2, **PATTERN DETECT**, should light, and you should hear K1 close. With K1 closed, DS1 should no longer light with each ring, but DS2 should stay lit until the ringing stops. Now, hang up the phone. In about 5 seconds, the relay should open and DS2 should go out.

If DS1 lights properly, but DS2 doesn't act right, and you're sure your wiring is correct, try increasing R5's value, shown as 270 kΩ in Figure 2, to 330 kΩ or more until the circuit works properly.

If everything is working okay so far, connect a 220-Ω resistor across the **PHONE LINE OUT TO EQUIPMENT** terminals. As you connect it, K2 should close, and DS3 and DS2 should light immediately and stay lit as long as the resistor is attached. K2 should close as well. This simulates your repeater equipment (autopatch, modem, etc) going on-line. When you remove the resistor, the LEDs should go out and both relays should open immediately.

Be careful not to short-circuit the **PHONE LINE OUT TO EQUIPMENT** terminals. Shorting them or the line connected to them will cause too much current to flow through the K2's coils, possibly burning them out. K2 is a sensitive relay with a very low coil resistance (9 Ω). Current flow through its coils must be no greater than about 150 mA.

This completes the checkout.

The Detector at Work

Repeater control equipment must not only work well on the bench: It must be dependable day in and day out. Several RingMaster Ring Detectors have been operating successfully for many months now with no problems—and no hackers!

Notes

[1] RingMaster is a registered trademark of BellSouth. Other phone companies may offer this or a comparable system under other names; for example, Southern New England Telephone calls theirs by the service mark Star Ring.

[2] So far, I have neither needed nor designed circuitry to sense the other two RingMaster ring sequences, but such circuits shouldn't be too hard to design. Besides, ready-made equipment is available to do this from at least one mail-order house (Hello Direct, 5884 Eden Park Pl, San Jose, CA 95151-0062, tel 408-972-1990 or 800-444-3556, fax 408-972-8155) if you don't want to build it, and have money to spend!

[3] PC board patterns for the RingMaster Ring Detector are available free from the ARRL. Send your request to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. With your request for the RUMBOLT RING DETECTOR PC-BOARD TEMPLATE, enclose a business-size envelope with one First-Class stamp.

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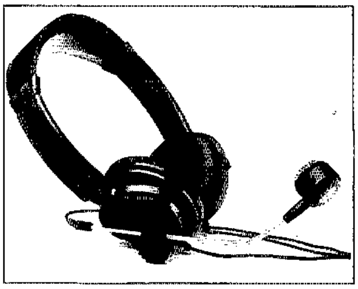


The Elkhart County Tone Alert

Use this DTMF decoder for group call-ups, for SELCALL and LiTZ alert.

By Dennis A. Drudge, W9XD
27901 CR 36
Goshen, IN 46526

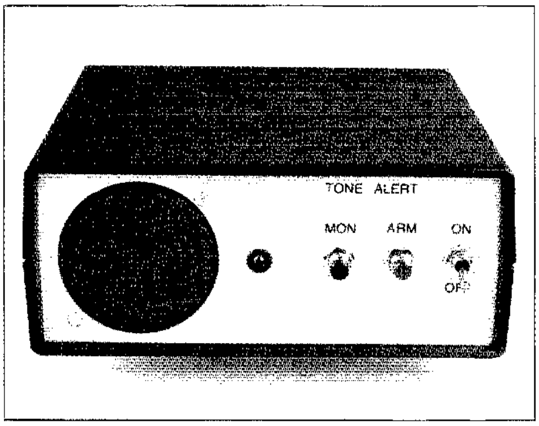
In early 1992, the leader of the SKYWARN program in Elkhart County, Indiana, determined a need for a tone-alert system to notify our members of impending severe-weather situations. Our requirements included low power consumption, the ability to respond to more than one tone sequence, audible and visual alarms and, of course, low cost. Our literature search revealed an excellent circuit in *QST*[1] that we used as our circuit's foundation. Using a minimum of readily available components, the total cost of this project (even if all parts are purchased new and in single quantities) is less than $50.

This unit responds to two 3-digit DTMF sequences, as well as the ARRL-endorsed Long Tone Zero (LiTZ) signal: A 3-second zero tone used on repeaters as a universal distress call.[2] When triggered, the system sounds an audible alarm for 5 seconds, sets a latching relay that switches received audio to a speaker, and enables a blinking LED.

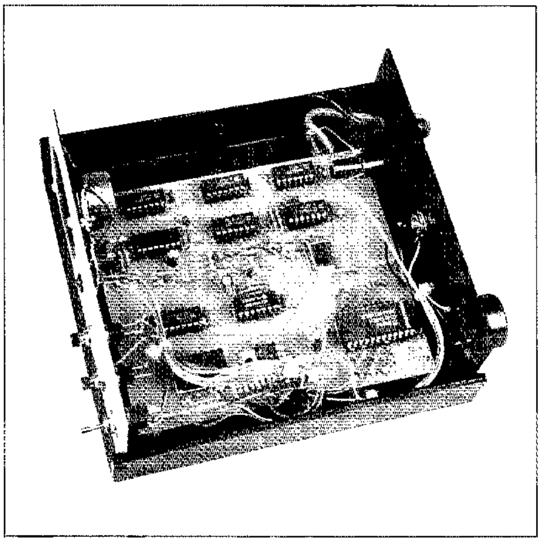
Circuit Description

Refer to Figure 1. Receiver audio output is fed to DTMF decoder U1. The decoded data, in hexadecimal format, is sent to U2, a 4 to 16-line decoder. The U2 outputs you select (0 through 9, A through D, * and #) are connected by manually installed jumpers to the D inputs of flip-flops U6, U7, and U8. You wire one of your selected three-digit sequences to inputs A1, A2, and A3, and your second three-digit sequence to inputs B1, B2, and B3. U6 through U8, inclusive, are sequentially enabled by U3, a digit counter that is clocked by the data valid (DV) strobe line of U1 through U5D. The DV strobe line also triggers monostable multivibrator U4B that provides an output pulse approximately 5 seconds after the end of the last tone fed to the input of U1. (The 5-second delay is determined by the values of R2 and C6.)

If the correct information appears at a D input of one of the flip-flops while it is enabled by U3, that flip-flop's Q output goes high. If a correct *three-digit sequence* is received, the Q outputs of the appropriate *three flip-flops* go high, causing the output of **AND** gate U9A or U9B to go high. This high turns on switch Q1, enabling the



Grille cloth covers the speaker hole at the left of the brushed aluminum front panel. A blinking LED, used as a visual alarm, is next to the speaker. Black (**MON**itor) and red (**ARM**) push-button switches come next, followed by the power **ON/OFF** toggle switch.



The PC board is supported on ½-inch-high nylon snap-on standoffs. Mounted on the rear panel are BZ1 (the piezo buzzer), the audio-input phono jack J1 and the fuse holder for F1, followed by a strain relief securing the dc-power-input leads. BZ1 is mounted on the *outside* of the rear panel, with its two leads passing through panel holes behind and beneath it. A close look at this PC board shows the as-yet-unwired multiple jumper-wire holes (near the large IC—U2—next to BZ1) for selecting any of the 16 DTMF tones. The easy-access clam-shell enclosure is made of black plastic except for the 1/16-inch-thick brushed-aluminum front panel. Two machine screws hold the cabinet top and bottom together, with the removable front and rear panels snugly secured in slots in the cabinet halves. (Similar enclosures are available through several mail-order suppliers such as Hosfelt Electronics, Inc, [no minimum order], 2700 Sunset Blvd, Steubenville, OH 43952-1158, tel 800-524-6464, 614-264-6464; fax 614-264-5413.—*Ed.*)

[1] Notes appear on page 46.

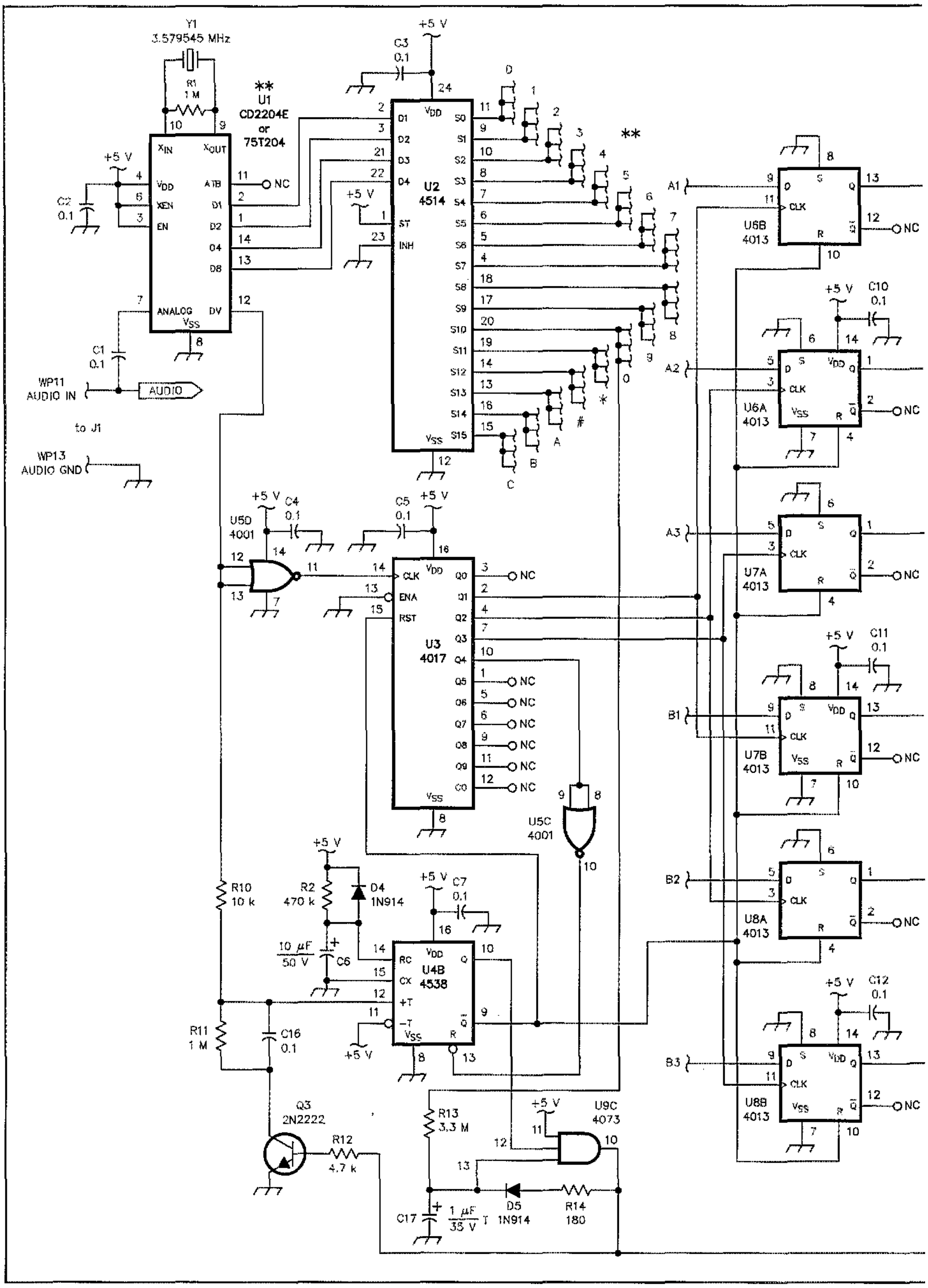
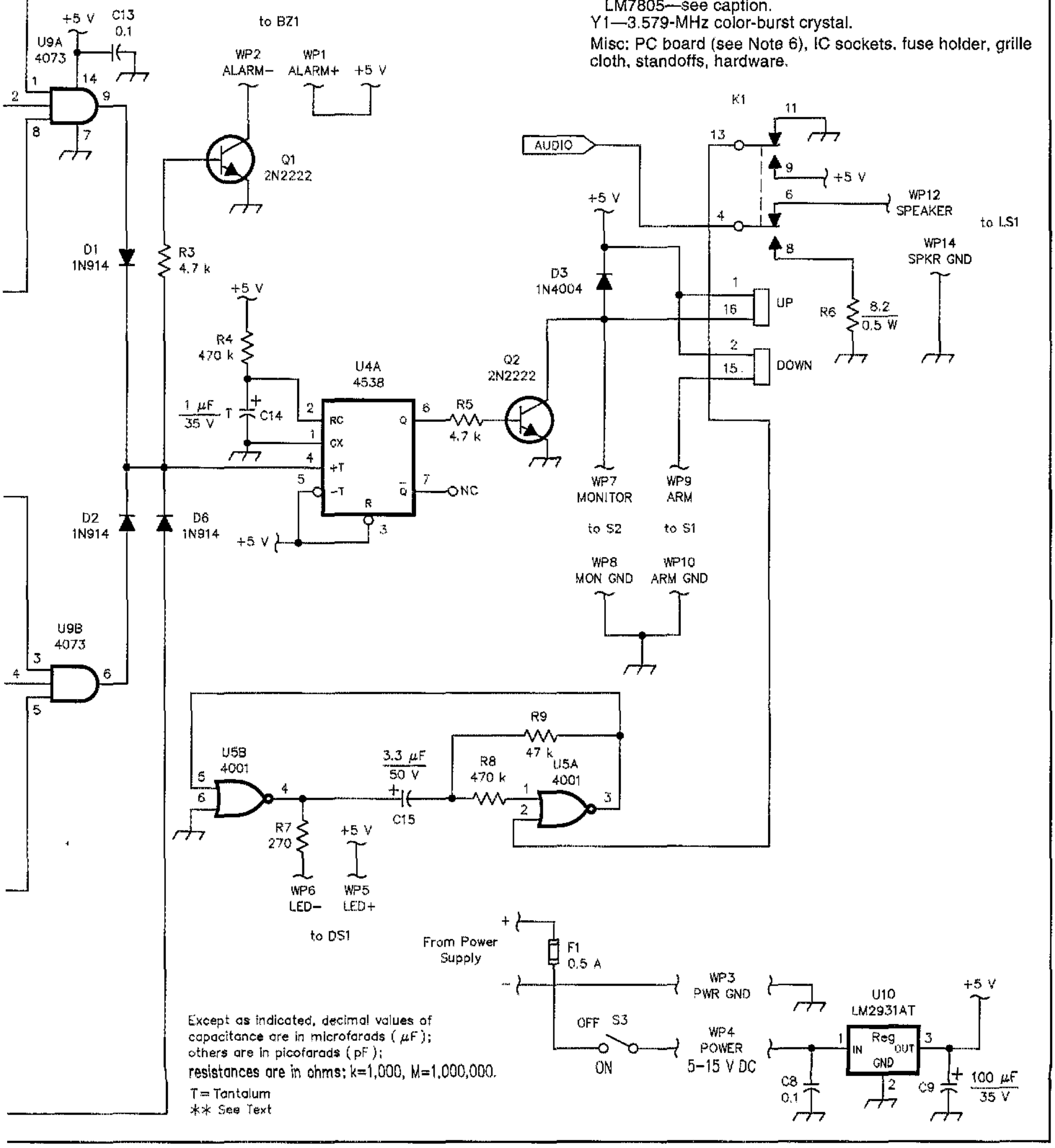

Y1
3.579545 MHz
R1
1 M
**
U1
CD2204E
or
75T204
XIN
XOUT
VDD
XEN
EN
A1B
D1
D2
D4
D8
ANALOG
DV
VSS
NC
+5 V
C2
0.1
C1
0.1
WP11
AUDIO IN
AUDIO
to J1
WP13
AUDIO GND
C3
0.1
U2
4514
ST
INH
S0
S1
S2
S3
S4
S5
S6
S7
S8
S9
S10
S11
S12
S13
S14
S15
**
D
1
2
3
4
5
6
7
8
9
0
*
#
A
B
C
U5D
4001
C4
0.1
C5
0.1
U3
4017
CLK
ENA
RST
Q0
Q1
Q2
Q3
Q4
Q5
Q6
Q7
Q8
Q9
CO
U5C
4001
R10
10 k
R2
470 k
D4
1N914
10 µF
50 V
C6
C7
0.1
U4B
4538
RC
CX
+T
−T
R11
1 M
C16
0.1
Q3
2N2222
R12
4.7 k
R13
3.3 M
U9C
4073
C17
1 µF
35 V T
D5
1N914
R14
180
A1
A2
A3
B1
B2
B3
U6B
4013
U6A
4013
U7A
4013
U7B
4013
U8A
4013
U8B
4013
C10
0.1
C11
0.1
C12
0.1
CLK
S
R
Q
Q̄

Figure 1—Schematic of the Elkhart County Tone Alert circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. The PC-board provides three wire holes for each of the 16 DTMF tones. U10 is a 5-V low-dropout voltage regulator. A more-commonly available LM7805 can be used instead, but if that's done, use a 1-µF tantalum capacitor at C9 for stability. Push-button, momentary-contact switches are used for the **ARM** and **MON**itor switches.(See Note 5 for addresses and telephone numbers of part suppliers. AE = Allied Electronics; CS = Circuit Specialists; DK = Digi-Key; HK = Hamilton Hallmark.)

BZ1—Piezo buzzer, 3 to 20 V dc.
DS1—LED.
F1—0.5-A, 125-V fuse.
J1—Single-hole-mount phono jack.
K1—DPDT latching relay, 5-V dc coil (OMRON G6AK-234P-ST-US-DC5; DK part number Z749-ND).
LS1—2-inch diameter, 8-Ω, 0.2-W speaker.
S1, S2—SPST normally open push-button switch.
S3—SPST toggle switch.

U1—DTMF decoder, Harris CD22204E or SSI 75T204 (CS: Harris CD22204E; AE and HK: Silicon Systems International [SSI] 75T204).
U2—4514 4 to 16-line decoder.
U3—4017 decade counter.
U4—4538 dual monostable multivibrator.
U5—4001 quad **NOR** gate.
U6-U8, incl—4013 dual type-D flip-flop.
U9—4073 triple, three-input **AND** gate.
U10—Low-dropout-voltage LM2931AT 5-V regulator (DK part number LM2931T-5.0-ND), or LM7805—see caption.
Y1—3.579-MHz color-burst crystal.
Misc: PC board (see Note 6), IC sockets, fuse holder, grille cloth, standoffs, hardware.



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

audio alarm (BZ1) for 5 seconds (the duration determined by U4B), and also triggers monostable multivibrator U4A, which provides a 1-second pulse (R4/C14) that sets K1, a DPDT latching relay, via Q2. One pole of K1 connects an 8-Ω speaker (LS1) to the audio line. The other pole enables a gated oscillator circuit[3] U5A-U5B, which causes DS1 to blink at an approximate 2-Hz rate. The speaker remains on and DS1 blinks until the circuit is manually reset by the **MON** switch, S2.

The tone alert also responds to a 3-second zero tone via **AND** gate U9C and its related circuitry. One input of U9C is tied to +V; the second input connects to the Q output of U4B (which goes high for 5 seconds when a valid tone is received); the third input is wired to U2's 0 output. When a continuous 0 tone is received, C17 charges in approximately 3 seconds, causing the input at pin 13 of U9C to go high. With all three inputs now high, the output of U9C (pin 10) also goes high, triggering the alarm as detailed earlier. Q3 and its related circuitry force the +T input of U4B to go low for a short time, then float high as C16 charges, in order to allow retriggering of U4B. D4 protects U4B by providing a discharge path for C6 when the unit is turned off.[4]

Construction

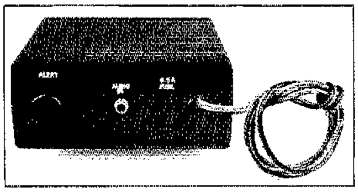
Components are readily available from several sources.[5] Your local Radio Shack outlet can supply many of the common parts. Part placement is noncritical; the use of IC sockets is recommended. The first eight prototypes were built on Radio Shack 276-154 universal circuit boards; now, a *double-sided* PC board and partial kit are available from the author.[6] The PC board fits tidily into a plastic 2½×6×6¼-inch (HWD) plastic, clam-shell enclosure.

Operation

The unit operates from power supplies delivering 6 to 16 V dc, drawing just 10 mA during standby. Current demand rises briefly (to about 100 mA) when K1 is energized. With power applied and audio connected, the **ARM** pushbutton (S1) places the unit in its ready state. This same button is used to reset the circuit after activation. The **MON**itor pushbutton (S2) latches K1 to activate the speaker for monitoring. To monitor incoming receiver audio with the unit's power off, press the **MON**itor switch and remove power by opening S3 (**ON/OFF**).

Summary

You can use one tone sequence for a group call-up, assigning the other as an individual SELCALL. Inclusion of the LiTZ response circuit allows you to hear someone unfamiliar with your repeater when they summon assistance. With this system, you can monitor a busy channel without distraction, yet be alerted to important calls.



Rear panel view of the Elkhart County Tone Alert.

Acknowledgments

Many thanks are due to David Evans, AA9DG, who drew the schematic, arranged for the PC-board design, and whose shrewd overall insight proved invaluable to the project. Thanks also to George Trenshaw, KD9UQ; Howard King, W9OCK; and Juli Mills (CAD operator), for their professional assistance, and to Sanford Swartzendruber, W9JOE; James Kehr, N9DUZ; Jon Slough, KB9ATR; Edward Charles, KB9BBI and John Aughey, N9IOX; for testing the early prototypes.

Notes

[1] V. Yakamavich, "Professional Quality DTMF Decoder and SELCALL System," *QST*, Feb 1988, pp 19-22.

[2] B. Battles, FM/RPT, *QST*, Oct 1992, p 82, P. Newland, "ZERO: An Alerting Device for Repeater Users." See also *QST*, Nov 1992, pp 108-110, and B. Battles, FM/RPT, *QST*, Jan 1993, p 104, for correction.

[3] D. Lancaster, *CMOS COOKBOOK* (Howard W. Sams & Co, 1977), p 234.

[4] *Motorola CMOS Databook*, 1978, p 7-504.

[5] Parts for this project are available from: Allied Electronics (minimum order $25 in the US, $50 outside of the US), tel 800-433-5700 in the US, fax 817-595-6444; 817-595-3500 outside of the US; Circuit Specialists, Inc (minimum order $15, unless order is prepaid by check or money order, in which case there is no minimum order), PO Box 3047, Scottsdale, AZ 85271-3047, tel 800-528-1417, 602-464-2485, fax 602-464-5824; Digi-Key Corp ($5 handling charge on orders less than $25), 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539, 218-681-6674, fax 218-681-3380; and Hamilton Hallmark (mimimum order $50), 10 S Centennial Dr, Peabody, MA 01960, tel 800-332-8638, 516-434-7490, fax 516-434-7491.—*Ed.*

[6] A 4½ ×5⅛-inch double-sided PC board with plated-through holes is available from the author for $22 postpaid in the US; a partial kit containing the PC board, K1 and U1 is $30 postpaid in the US. With all orders outside the US, please add 15% for surface-mail delivery. Make your check or money order payable to Dennis Drudge. Please allow three weeks for delivery within the US.

A template package containing the *double-sided* PC-board pattern and a parts overlay is available from the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please include a business-size SASE and identify your request for the ELKHART COUNTY TONE ALERT SYSTEM PC-BOARD TEMPLATE.

Dennis Drudge was first licensed as WN9ZKN in 1968 at the age of 15; he presently holds an Extra Class license. Dennis has a BA degree from Goshen College. He began experimenting with digital logic circuits about 15 years ago. Dennis is employed as a purchasing agent for the JAY-PARR division of JAYCO, INC, a recreational vehicle manufacturer. In his free time, Dennis enjoys almost all facets of HF and VHF operation from a home station that has been totally *solar-powered since 1983.*

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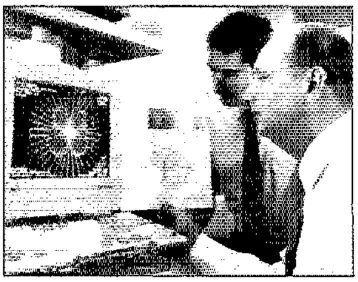

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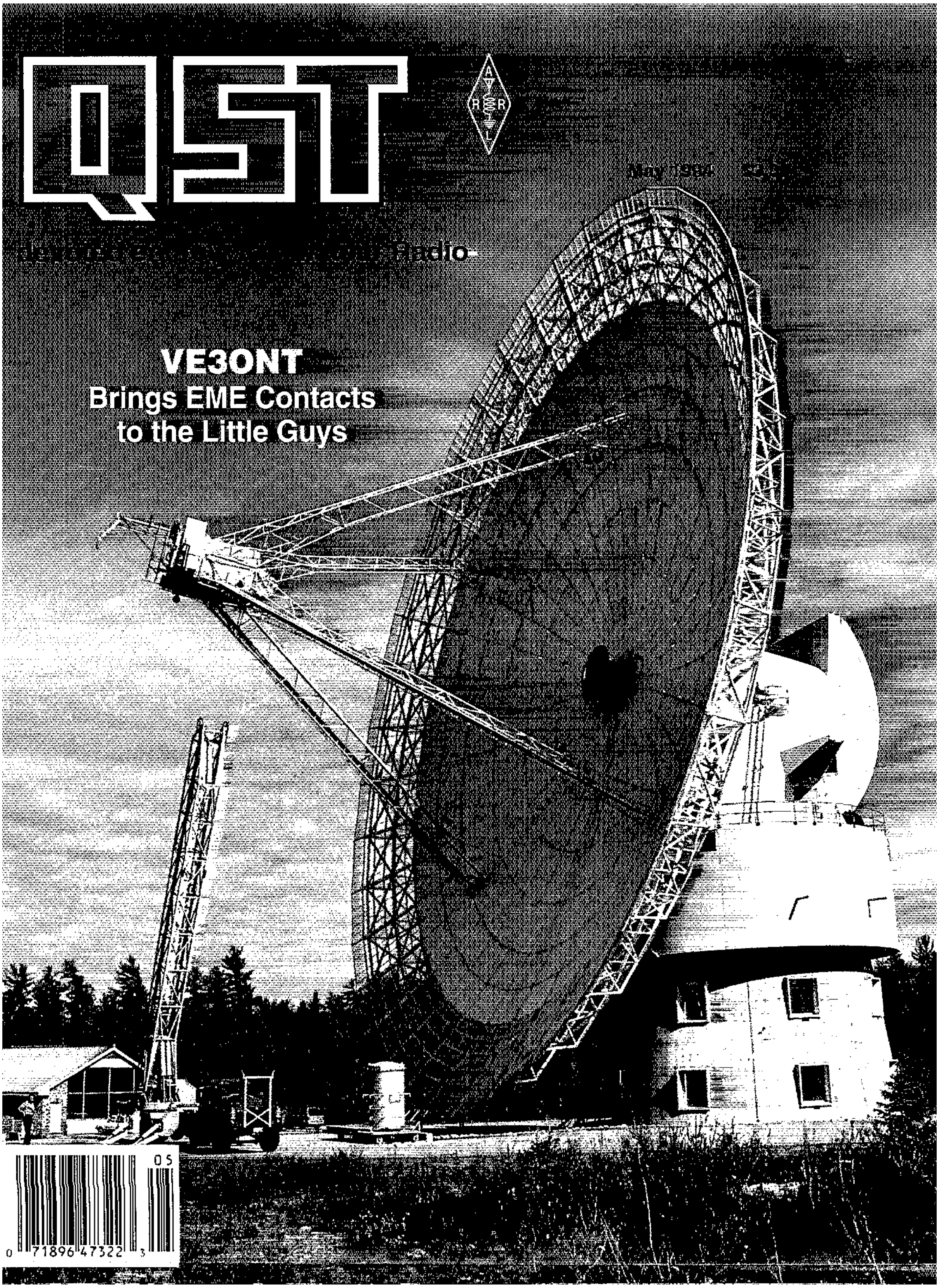

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OUR COVER

This 150-foot dish at the Algonquin Radio Observatory provided enough punch to give many hams their first two-way moonbounce contacts. A party of enthusiastic Canadian amateurs of the Toronto VHF Society brought new meaning to the term "big gun" when they operated VE3ONT in the Seventeenth ARRL International EME Competition last October and November. More details about the station appeared in the World Above 50 MHz column in October 1993 and January 1994 *QST*; this month, there are more pictures in Up Front in *QST*, and the complete contest results are on page 117. *(photo by Peter Shilton, VE3VD)*

CONTENTS

May 1994
Volume LXXVIII Number 5

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

Key Components of Modern Receiver Design—*Part 1*

Today's Amateur Radio receivers routinely achieve performance untouched by earlier generations of ham gear. Can our already excellent equipment be improved still further?

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The fact that many presentations have dealt with receiver capabilities and improvements over the years might seem to imply that all receiving problems have been solved and that the technology is mature. On the other hand, a look at the components actually used in the front ends of current receivers by different manufacturers indicates that just the opposite is true. While improvements have been made in mixers, amplifiers and synthesizers—and from a systems point of view there have been some subtle implementations of individual circuits—there is still work to be done! How else can one explain that different receiver designs—with similar block diagrams, and brought to market with the best of intentions—sound different on the air?

Modern ham equipment seeks to achieve high dynamic range in relatively standard, seemingly well-established ways. Nonetheless, I have discovered that it's useful to reevaluate the definition of dynamic range, and how AGC and gain distribution affect it. As part of this discussion, I will show how to overcome the effects of a form of intermodulation distortion long thought unimportant in ham circles—a species of RF IMD that noticeably degrades reception even in high-end Amateur Radio MF/HF transceivers.

Likewise, we may benefit by revisiting the influence of oscillator phase noise on dynamic range, and the well-known interdependence of fast PLL settling time and phase noise. After discussing my findings, I will propose a hybrid synthesizer arrangement—a DDS-driven PLL system—and show how commercial CAD can now accurately predict the SSB phase noise of oscillators and provide the ability to optimize it.

Dynamic Range Types and Issues

There are several types of dynamic range. The first one, and probably the easiest to understand—"AGC range"—concerns whether a receiver is capable of maintaining a constant audio output level over a large input-signal amplitude range. The traditional school of thought requires AGC action to commence at 1 or 2 μV, leading to a condition where signals that produce an excellent signal-to-noise ratio may show absolutely no S-meter indication—a most undesirable effect. The reason for this is inappropriate receiver gain distribution—generally, a lack of gain at the second IF. Maintaining constant audio output must involve gain control at the receiver's second *and first* IFs, and possibly even at its input. I will address this issue later.

Intermodulation-Distortion Dynamic Range

The output of a linear stage tracks the input signal decibel by decibel, with every 1-dB change in its input signal(s) corresponding to an identical 1-dB output change. This is the stage's *first-order* response.

Because no device is perfectly linear, however, two or more signals applied to it *intermodulate* to some degree, generating sum and difference frequencies. These *intermodulation distortion* (IMD) products occur at frequencies and amplitudes that depend on the order of the IMD response as follows:

- *Second-order* IMD products change 2 dB for every decibel of input-signal change,[1] and appear at frequencies that result from the simple addition and subtraction of input-signal frequencies. For example, assuming that its input bandwidth is sufficient to pass them, an amplifier subjected to signals at 6 and 8 MHz will produce second-order IMD products at 2 MHz (8 – 6) and 14 MHz (8 + 6).
- *Third-order* IMD products change 3 dB for every decibel of input-signal change,[2] and appear at frequencies corresponding to the sums and differences of twice one signal's frequency plus or minus the frequency of another. Assuming that its input bandwidth is sufficient to pass them, an amplifier subjected to signals at 14.02 MHz (f_1) and 14.04 MHz (f_2) produces third-order IMD products at 14.00 ($2f_1 - f_2$), 14.06 ($2f_2 - f_1$), 42.08 ($2f_1 + f_2$) and 42.10 ($2f_2 + f_1$) MHz. The subtractive products (the 14.00 and 14.06-MHz products in this example) are close to the desired signal and can cause significant interference. This is why our receivers' third-order IMD performance is so important.

It can be seen that the IMD order determines how rapidly IMD products change level per unit change of input level. *N*th-order IMD products therefore change by *n* dB for every decibel of input-level change.

IMD products at orders higher than three can and do occur in communication systems, but the second- and third-order products are most important in receiver front ends.

Intercept Point

The second type of dynamic range concerns the receiver's *intercept point*, sometimes simply referred to as *intercept*. Intercept point is typically measured by applying two or three signals to the antenna input, tuning the receiver to count the number of resulting spurious responses, and measuring their level relative to the input signal.

Because a device's IMD products increase more rapidly than its desired output as the input level rises, it might seem that steadily increasing the level of multiple signals applied to an amplifier would eventually result in equal desired-signal and IMD levels at the amplifier output. Real devices are incapable of doing this, however. At some point, every device *overloads*, and changes in its output level no longer equally track changes at its input. The device is then said to be operating in *compression*. Pushing the process to its limit ultimately leads to *saturation*, at which point input-signal increases no longer increase the output level.

The power level at which a device's second-order IMD products equal its first-

[1] Notes appear on page 32.

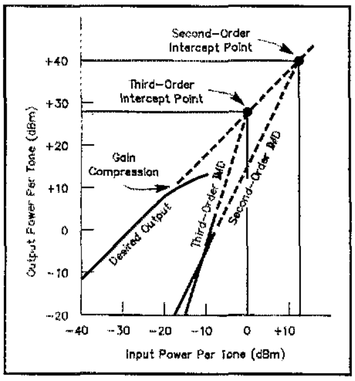

Figure 1—A linear stage's output tracks its input decibel by decibel on a 1:1 slope—its *first-order* response. *Second-order intermodulation distortion (IMD) products* produced by two equal-level input signals ("tones") rise on a 2:1 slope—2 dB for every 1 dB of input increase. *Third-order IMD products* likewise increase 3 dB for every 1 dB of increase in two equal tones. For each IMD order *n*, there is a corresponding *intercept point* IP$_n$ at which the stage's first-order and *n*th order products are equal in amplitude. The first-order output of real amplifiers and mixers falls off (the device overloads and goes into *compression*) before IMD products can intercept it, but intercept point is nonetheless a useful, valid concept for comparing radio system performance. The higher an amplifier or mixer's intercept point, the stronger the input signals it can handle without overloading. The input and output powers shown are for purposes of example; every receiver exhibits its own particular IMD profile. (*After W. Hayward,* Introduction to Radio Frequency Design, *Figure 6.17*)

order output (a point that must be extrapolated because the device is in compression by this point) is its *second-order intercept point.* Likewise, its *third-order intercept point* is the power level at which third-order responses equal the desired signal. Figure 1 graphs these relationships.

Input filtering can improve second-order intercept point; device nonlinearities determine the third, fifth and higher-odd-number intercept points. In preamplifiers, third-order intercept point is directly related to dc input power; in mixers, to the local-oscillator power applied.

Intercept point can be confusing because it can be specified in terms of input or output power. Intercept point should be referred to device *output* because that's where the trouble occurs, but input intercept is commonly given. Therefore, if an amplifier or a mixer has a particular intercept point—let's say +30 dBm at 10 dB gain—and then its gain is increased by an additional 10 dB, its dynamic range *decreases* by the amount of the gain.

A good example of this confusion is the highly acclaimed Plessey SL6440C active mixer. DeMaw and Collins[3] evaluated this device with 200-Ω input and output terminations. Their findings were based on a device gain of 8 dB, maximum, and an output intercept point of +30 dBm. By the time the SL6440C's output impedance is raised to 1.5 kΩ on each collector, all other things being equal, its intercept point deteriorates to about +10 dBm. The additional gain favors the intermod products, degrading the intercept point by 20 dBm and destroying the mixer's dynamic range. It's therefore highly desirable to keep mixing and amplification functions separate—mixer first, amplifier second. A smart way to accomplish this is to put a crystal filter between the two to band-limit the spectrum, and thus the signal power, applied to the post-amplifier. More on this concept later.

Blocking and Desensitization

Because of its noise sidebands, a receiver's synthesizer mixes adjacent-channel signals into the IF chain as noise even though the signals themselves fall outside the IF passband. The result is a degradation of receiver noise floor known as *blocking* or *desensitization.*[4]

In addition, synthesizers are frequently *dirty*—spectrally impure—specifically if complicated mix and divide arrangements are used. Even most modern direct digital synthesizers do not take advantage of configurations that minimize unwanted spurious response. In this article, I will look at a way of drastically improving a synthesizer's output purity while reducing costs at the same time.

Tone Spacing and Measurement Results

Dynamic-range characteristics that involve two or more signals also depend on the signals' frequency spacing. In the standard case, where the desired, low-level signal is bothered by a stronger signal, a receiver's performance drastically varies as a function of the spacing between the two. Most receiver measurements of this type are conveniently done at a test-signal ("tone") spacing of 20 to 25 kHz. Real life is often not so generous! Strong signals may be spaced only a few kilohertz apart, and the receiver's first-IF ("roofing") crystal filter (say, 15 kHz wide) does not protect the second mixer as a much narrower filter, like the 2 to 4-kHz units commonly used for SSB, would do. In a radio using a 15-kHz first-IF filter and a 2.5-kHz second-IF filter, the second mixer must therefore digest the signal energy in 12.5 kHz of spectrum invisible to circuitry after the second-IF filter. The wider roofing filters (Figure 2) used in some amateur equipment makes life even harder for the second mixer.

The standard AGC approach derives its control voltage from second-IF energy and does not even acknowledge the existence of these signals. The current trend to replace IF stages with digital signal processing will aggravate this situation even more because A-to-D converters do not respond gracefully to signals that exceed their dynamic range. AGC attack time can therefore be particularly critical in DSP-equipped receiving systems even if the AGC range is otherwise adequate.

How to Deal with These Issues

What Do We Find

Is the receiver's AGC, selectivity and gain distribution proper? Figure 3 shows the front-end and IF block diagram of a reasonably standard transceiver design—that of the recently introduced Kenwood TS-50 transceiver. The first IF stage of its receiver (TX-RX Unit Q17, a 3SK121 MOSFET) receives AGC, but that this AGC is derived from the output of the radio's *second*-IF chain. For the reasons described earlier, this system cannot protect the TS-50's second mixer from overload caused by signals inside the roofing filter passband and outside the passbands of the switchable filters that head its second IF. A further drawback of this AGC method is that Q17's third-order intercept point worsens as its gain is reduced—a characteristic intrinsic to MOSFETs that are gain-controlled in this way. Using a differential amplifier or PIN-diode attenuator as the gain-controlled stage would avoid this.

Radios constructed along similar lines often suffer from:

- Insufficient AGC range. Listen to a full-carrier, double-sideband AM signal, modulated at least 50%, at a level of 100 mV or more. You will likely hear distortion because many receivers' AGC circuitry run out of control range by the time incoming signals reach this level.
- Overload related to insufficient AGC if a 20-mV signal that appears 5 kHz away from a properly tuned-in CW or SSB signal falls inside the roofing filter passband and outside the second-IF filter.
- Desensitization caused by LO spurious signals and phase noise, which can allow strong nearby carriers to raise the receiver noise floor. Does your all-band transceiver allow you to tune to an input frequency of 5 or 10 kHz instead of 500 kHz or higher? Interesting time and standard-frequency signals can be received below 50 kHz, but designers of receivers and transceivers usually disallow tuning

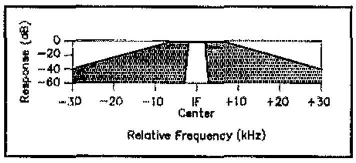

Figure 2—Signals that fall within the passband of a receiver's *roofing filter* (shaded and unshaded zones under curve) but outside the passband of subsequent narrower filters (unshaded zone) can overload the circuitry between them.

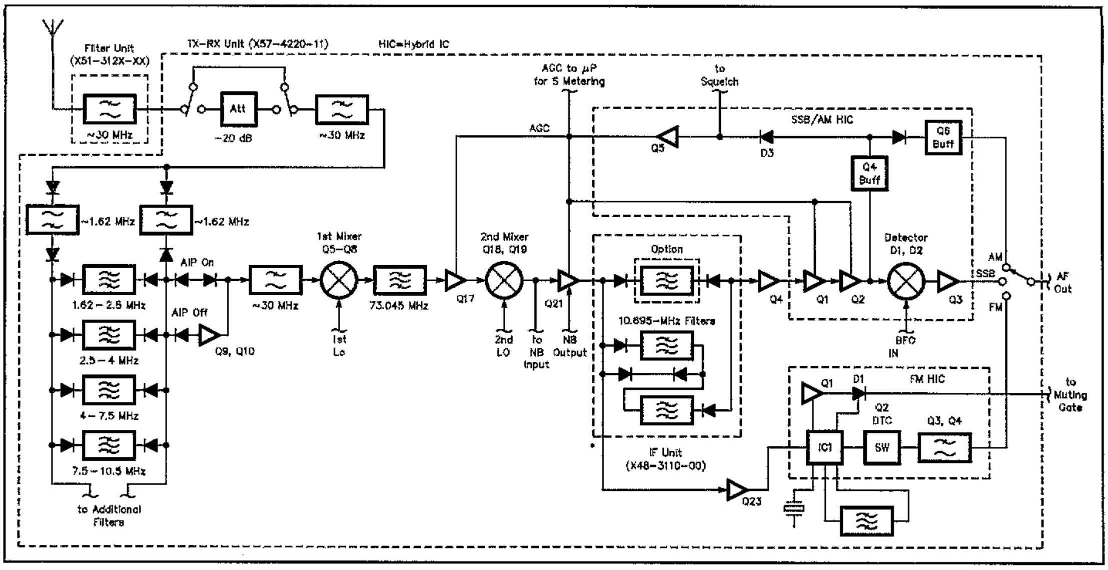

Figure 3—In keeping with general practice in Amateur Radio transceivers, Kenwood's TS-50 applies only second-IF AGC to amplifier stages between its roofing and second-IF filters.

below 100 kHz because of their limited synthesizer purity. Your radio's firmware-set lower tuning limit directly reflects how "good" its synthesizer really is.

The Right Approach

Distributed AGC

Figure 4 shows a partial block diagram of a receiver with distributed AGC that includes second-mixer overload monitoring. One AGC loop operates at the front end and protects the input stage with an electronically controlled attenuator. In most radios, this attenuator is *manually* operated from the front panel and offers a gain-reduction range of perhaps 30 dB in 5, 6 or 10-dB steps. Many operators find its use somewhat confusing because it changes the S-meter indication. Modern microprocessor-controlled radios should be capable of selecting the proper attenuation level and correcting the signal-meter indication appropriately.

Signal-Strength Metering

S-meter calibration, indication and use have long been somewhat emotional issues in Amateur Radio. I strongly recommend that we redesign our signal meters along the lines of those used by our professional colleagues, calibrating them from -20 dBµV (0.1 µV in 50 Ω) to +100 dBµV (100 mV in 50 Ω)—a 120-dB range. Such a system requires that a receiver's AGC system have enough pre-detector gain so that the AGC operates on IF noise alone. Note that, per Table 1, -20 dBµV is close to the extrapolated definition for S0 of about 0.07 µV.[5]

Aside from its objectivity, this arrangement offers the advantage of allowing true comparisons between signals, and accurate system-parameter measurements (such as antenna front-to-back ratio, useful characterization of which is all but impossible with most S-unit-based signal indicators).

In addition to good second-IF AGC and controllable input attenuation, implementing a 120-dB-range signal meter in a multi-conversion receiver requires AGC at the radio's *first* IF. Previous attempts to do this were based on PIN-diode attenuators and have not been followed through by many designers.

The major drawback of the diode-

Table 1
S-Unit Equivalents Based on S9 = 50 µV in 50 Ω

S9	50 µV
S8	25 µV
S7	12.5 µV
S6	6 µV
S5	3 µV
S4	1.3 µV
S3	0.75 µV
S2	0.3 µV
S1	0.15 µV
S0	0.07 µV*

*Extrapolated from S1 value; see text and Note 5.

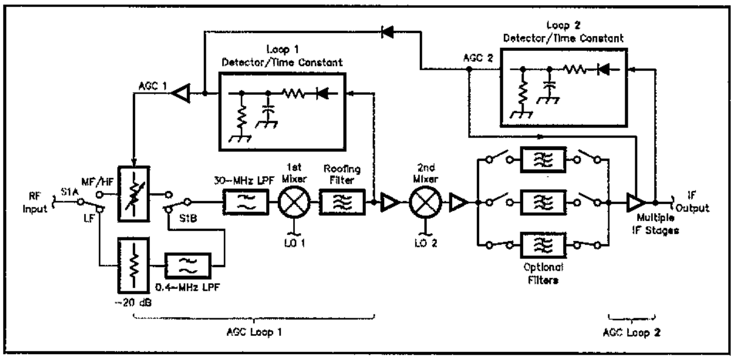

Figure 4—This receiver block diagram includes *two* AGC loops—one driven by first-IF energy that is band-limited by the roofing filter, and another driven by second-IF energy band-limited by the optional second-IF filters. The first loop controls a PIN-diode pi attenuator ahead of the first mixer; the second loop controls second-IF amplifier stages. A microprocessor adjusts the time constants of both loops so that time delays introduced by the filters do not cause AGC oscillation. For clarity, this block diagram omits the input filtering (to the left of RF Input) necessary for a practical system.

attenuator approach, besides its cost, is that an attenuator's noise figure equals its insertion loss. Assuming that such a system's AGC comes from a single AGC detector (usually at the end of the second IF in multiconversion receivers), sophisticated control circuitry is needed to adjust the system's AGC time constant as different IF filters are selected. Otherwise, the varying time delays contributed by different IF filters may cause AGC instability.

Independent First-IF AGC

The correct solution to this problem is an additional AGC loop that operates entirely at the first IF. A monitoring stage samples the second mixer's input level and applies AGC to gently attenuate signals that would otherwise overdrive the second mixer. Although this may limit the maximum signal-to-noise ratio achievable by wanted signals, it also cleans up intermodulation problems that would otherwise occur. Properly implemented and operated in conjunction with good second-IF AGC, independent first-IF AGC like this can largely free us of the receiver-generated intermod products we all too often experience on the air.

Some Comments on Switching Diodes

The receiver sections of amateur MF/HF transceivers generally use diode-switched front-end filtering. The switching diodes used have low junction capacitance and can typically handle medium dc levels (10 to 100 mA). These characteristics are important because we want these diodes to contribute minimal loss when turned on and leak very little RF when turned off.

The two-tone, third-order IMD dynamic-range testing routinely done to amateur transceivers seems to point up no weakness in these switching diodes. In real life, however, a huge number of signals simultaneously appear at a transceiver's antenna connector. Periodically, their voltages all sum in phase, producing, for short durations, enough voltage to change the bias of the diode at the input of the filter in use. This causes intermodulation distortion—generally, second-order IMD. This is ironic for two reasons: First, this diode-generated IMD generates exactly the interference the filters switched by the diodes are supposed to prevent! Second, Amateur Radio equipment reviews have long let second-order front-end IMD go unmeasured because we have long assumed that our radios' front-end filtering reduces this IMD to a nonproblem. Later in this article, I will present measurement results that prove that second-order IMD is a very real problem today.

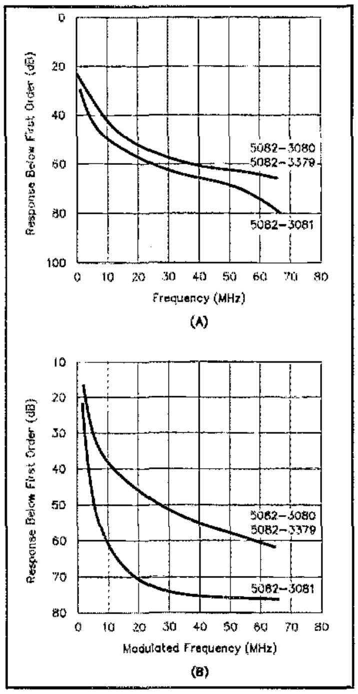

Figure 5—Typical second-order IMD (A) and cross-modulation distortion (B) responses of RF PIN diodes. Cross-modulation distortion, an effect also long untested in most Amateur Radio equipment reviews, is particularly important in these days of superpower shortwave broadcasters operating in and adjacent to Amateur Radio bands. Cross-modulation causes one signal's modulation to appear on other signals passing through the distorting device. (Conditions for A: 10-dB bridged-T attenuator; 40 dBmV output levels; one input frequency fixed at 100 MHz. Conditions for B: 10-dB bridged attenuator; 40 dBmV output levels; unmodulated frequency fixed at 100 MHz; variable frequency 100% modulated by 15-kHz audio.)

The best way to avoid switching-diode IMD is to switch the filters with relays instead of diodes, and military and commercial gear generally take this approach. Relays are costly, however. A less expensive work-around that's acceptably good for Amateur Radio equipment is to use diodes—PIN diodes—designed for this application. The two best-known US manufacturers of PIN diodes for this type of low-frequency application are Hewlett-Packard and Alpha Industries. The best diode for the shortwave range is the HP 8052-3081. (A similar Alpha diode with a minority carrier lifetime of 4 µs is also available. The Alpha diode can handle higher power.)

A hot-carrier diode's lowest frequency of operation is determined by the inverse of the lifetime of its minority carriers. For the HP 8052-3081, which has a minority carrier lifetime of 4 µs, the lowest frequency is therefore 250 kHz (1÷0.000004). Figure 5 shows the intermodulation distortion properties of several HP hot-carrier diodes, including the 3081, as a function of frequency and bias.

Notes

[1] This figure assumes that the IMD comes from *equal-level* input signals.

[2] This figure also assumes equal-level input signals.

[3] D. DeMaw and G. Collins, "Modern Receiver Mixers for High Dynamic Range," *QST*, Jan 1981, pp 19-23.

[4] [This characterization of blocking differs from what we call *blocking* in ARRL Lab receiver testing for *QST* Product Reviews. We define a receiver's blocking dynamic range as the difference, in decibels, between the signal power that produces a 3-dB signal-plus-noise to noise ratio (in other words, the receiver's *minimum discernible signal* [MDS]) and the power of an out-of-passband signal that reduces the audio output produced by a desired signal (at a level 10 dB below the radio's 1-dB compression point for radios with their AGC turned off, or equal to 20 dB above MDS for radios with AGC that cannot be turned off) by 1 dB. This measurement indirectly reflects the purity of the oscillators involved if they are so noisy that reciprocal mixing raises the receiver noise floor enough to mask the onset of blocking. When this occurs, we characterize that measurement as *noise limited* and denote it as such in the Product Review. —*Ed.*]

[5] [Purists considering "S units" in terms of the RST signal-reporting system (in which the lowest S number is 1) may insist that there is no such thing as "S0," but we have not exhausted the communication possibilities afforded by today's real-world radio links by the time we've worked downward to S1 on a 6-dB-per-S-unit basis from the decades-old amateur "standard" of S9 = 50 µV in 50 Ω. Solid communication can often be established and maintained at signal levels too weak to move our S meters.—*Ed.*]

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QST

Coming in Part 2:
Building these concepts into receivers.

You Can Build: A Compact Loop Antenna for 30 through 12 Meters

Are you looking for a low-profile, compact antenna? With a coat of camouflage paint added to it, you could park this one just about anywhere!

By Robert Capon, WA3ULH
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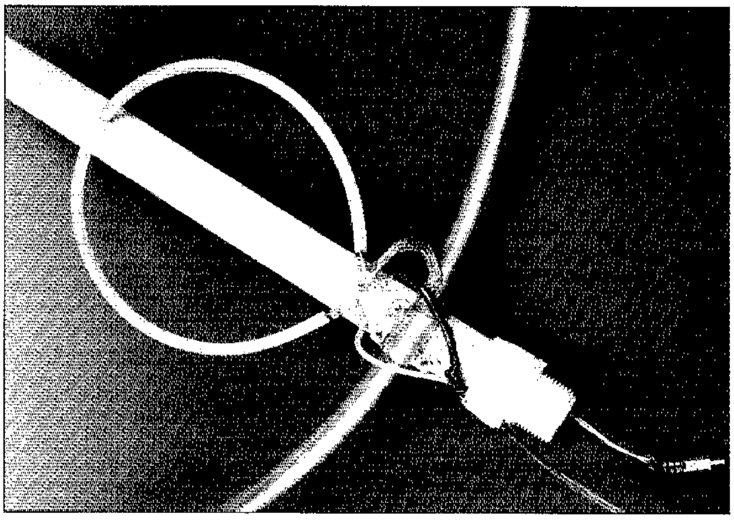


Figure 1—A close-up of the coupling-loop attachment.

I was intrigued by Franz (DL9RBT) Freller's miniature loop antenna.[1] It looked like the perfect portable-antenna solution for my QRP station! I wrote to Franz for information and received a prompt reply. Armed with construction information, I can now tell you how to build a multiband miniature loop antenna that can be set up in less than five minutes. This antenna is compact and performs on a par with a commercial multiband vertical antenna and it's inexpensive—about $35.

A Little Background

This loop is a physically small antenna—only 1/8 of a wavelength in circumference on 20 meters, increasing to 1/4 wavelength on 12 meters. There's extremely low resistance within the loop. It's tuned with a single-section variable capacitor and has a very high Q. As a result, the antenna exhibits a narrow bandwidth on 20 meters (10 to 20 kHz between the 2:1 SWR points), so the capacitor must be adjusted to retune the loop as you move across the band. On the higher-frequency bands, however, the loop has a progressively lower Q and a broader bandwidth (40 to 50 kHz). In fact, the narrow bandwidth and need for frequent retuning on the lower bands is the antenna's only drawback.

The antenna's outer copper-tubing loop (see Figure 1) is inductively coupled to the feed line by means of a small coax loop. This might appear to be a short-circuit because the small loop is attached to the feed-line's center conductor and shield. Actually, the small loop is not a short-circuit at all, but a one-turn inductor coupling to the large loop. The ground (braid) side of the small loop is attached to the large loop. This braid connection does not feed the signal to the large loop; it eliminates capacitive coupling between the two loops.

Mounted in the vertical plane, the antenna is directional; the nulls are perpendicular to the loop's axis and the antenna can be mounted quite close to the ground. If the loop is physically horizontal, it exhibits an omnidirectional pattern. In the horizontal plane, the loop should be at least 1/2 wavelength (about 33 feet on 20 meters) above the ground to work effectively. Installed horizontally, the antenna has a low angle of radiation—excellent for working DX.

Antenna efficiency depends on keeping the loop's surface resistance at an absolute minimum. *Don't* use small-diameter wire to connect the capacitor to the loop. Also, make the connections between the capacitor and the loop conductor as short as possible to eliminate unwanted resistance.

The loop develops a large voltage across the capacitor, and a minimum plate spacing of 3 mm is required for a transmitter output of 100 W. Because I used a small, single-section, air-variable capacitor (available from Ten-Tec[2]) the antenna handles a maximum applied power of approximately 7 W. If you use a different variable capacitor, make sure that it has a value of 2 to 100 pF. It's best to use a tuning capacitor equipped with low-resistance wiper contacts. You may be able to find a suitable tuning capacitor in a friend's junkbox, at a hamfest or a surplus parts outlet. You'll also need an enclosure in which you can house the capacitor; I used a plastic box.

Antenna Construction

Table 1 provides a complete parts list for the antenna. For the large copper loop, I bought 8 feet, 3 inches of 7/8-inch-diameter coiled copper plumbing pipe at a local hardware store for $1.09 per foot. The coiled copper pipe is easy to shape into a loop. The copper pipe dents easily, so handle it carefully. Uncoil the pipe and gently work it into a loop a little at a time. You may find it easiest to work the pipe while it's flat on a carpeted floor or work mat.

[1] Notes appear on page 36.

Table 1
Loop Antenna Parts List

Basic Antenna: 8¼-foot length of ⅝-inch-diameter coiled copper tubing; Ten-Tec 2- to 100-pF variable capacitor (Ten-Tec part no. 23227, available from Ten-Tec, 1185 Dolly Parton Pkwy, Sevierville, TN 37862, tel 615-453-7172, fax 615-428-4483); 20-inch length of RG-8 center conductor and dielectric; Radio Shack enclosure (270-231); Radio Shack knob (274-415), 2-inch-long #8 bolts and nuts; electrical eyelets.

Mast: Two 10-foot lengths of 1-inch-diameter PVC plumbing pipe; three T joints; four male thread-on caps; one female thread-on cap; five end-caps.

Motorized Drive Option: Radio Shack SWR Meter (21-524); two Radio Shack momentary DPDT toggle switches (275-637); Radio Shack female panel-mount phono jack (274-346); Radio Shack battery holder (270-382); 1-rpm high-torque dc motor, such as Edmund Scientific K41860 (12 V dc) or K41327 (3 V dc); 2½-inch-diameter hose clamp.

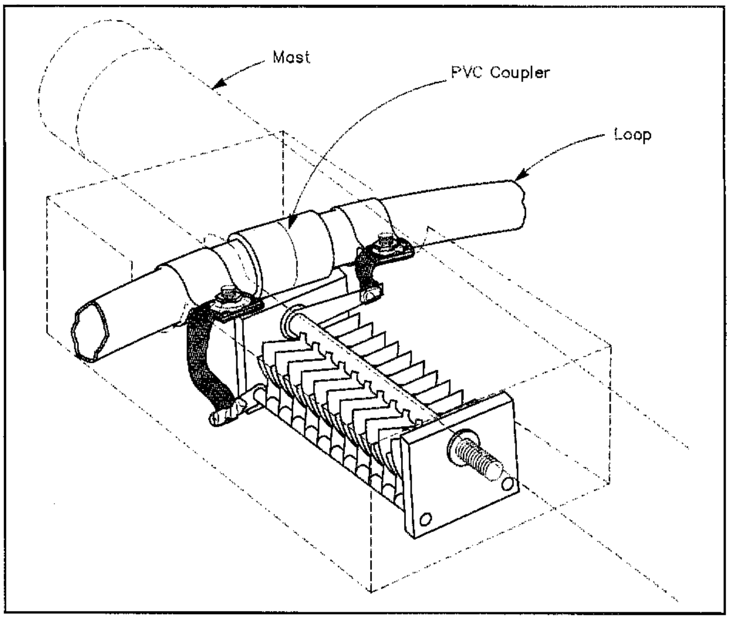


Figure 2—The tuning capacitor is housed in a plastic enclosure that is taped to the mast. Heavy copper braid is used to connect the capacitor rotor and stator sections to the copper-tubing loop.

There are two ways you can mount the tuning capacitor. One way is to use a punch (or nail and hammer) to dent the copper pipe approximately ⅜-inch in from each of the open ends of the loop to provide a drill-bit guide. Drill two ⅛-inch-diameter holes in the pipe. At the connection points, clean the pipe with no. 0000 steel wool and tin the copper loop and capacitor tabs. Bend the capacitor tabs into the small holes drilled into the pipe. Solder the capacitor tabs in place with a 150-W soldering iron or a small soldering torch. Ensure that the loop is adequately heated so that the solder flows into the connection.

A second, and better method (see Figure 2) is to secure the capacitor to the loop by fabricating two short mounting straps, soldering those straps to the ends of the loop, then using copper braid to obtain a low-loss connection to the straps and the tuning capacitor tabs. With this method, use a ½-inch PVC coupler as an end insulator to separate the loop halves.

Cut a 4-foot length of 1-inch OD PVC pipe and cement a threaded male PVC fitting to the bottom using PVC pipe cement (available at most hardware and plumbing supply stores). PVC cuts and drills easily and its joints weld solidly together with the PVC cement. *Be certain to use the cement only in a well-ventilated area!* Read and observe the precautions on the label.

Cut a ½-inch-square notch in the 4-foot pipe approximately 3 inches from the top. Place the loop assembly inside the notch, and fasten the loop to the pipe with the capacitor enclosure as a cover. You'll have to cut a ⅝-inch-diameter hole in the enclosure's end to pass the capacitor's shaft. I strapped the enclosure to the mast with plastic tie wraps. (See Figure 3.)

Fabricate an aluminum bracket to secure the loop's bottom to the PVC pipe. If you use the Radio Shack enclosure, simply cut off a 1-inch strip from either side of the aluminum cover (resulting in a 1-inch sheet metal strap with two holes in the ends), bend the sheet metal over a ½-inch-OD form (such as the handle of a socket wrench), and form ½-inch tabs using pliers. Fasten the loop to the mast with the homemade bracket using a pair of 2-inch #8 bolts.

The small coupling loop is made from a 20-inch length of RG-8 coax center conductor and dielectric. (Save the ground braid for later use.) To facilitate mounting the small loop, I used electrician's eyelets crimped and soldered to the ends.

The coupling loop is fastened to the PVC pipe with a pair of 2-inch-long #8 bolts. Fasten the coax feed line to the same bolts that hold the loop to the mast. Attach the center conductor to one side of the small loop and the shield to the other side. Run a 2-inch length of RG-8 braid from the ground side of the small loop to the metal mounting strap of the large loop.

Put the finishing touches on the antenna by cleaning the entire loop with no. 0000 steel wool and applying two coats of polyurethane varnish. In addition to making the copper shine, this prevents the copper from tarnishing (allowing a build-up of surface resistance) which lowers the loop's radiation efficiency.

A 20- to 40-m loop can be fabricated by using different values: a loop diameter of 1.7 meters and a small coupling loop diameter of 0.34 meter. Such a configuration will have a bandwidth of only 5 kHz on 40 meters, but should exhibit excellent efficiency and a broader bandwidth (20 to 40 kHz) on 20 meters.

Building the Mast

The antenna is so light that you can use almost anything to support it. You could plant it in the ground with a length of aluminum mast, or even hang the antenna with nylon rope for really portable operation.

To support the loop, I modified a design of a PVC base and mast developed by Bruce Auld, NZ5G. This free-standing PVC support structure can be assembled in just a few minutes. The structure has five components: an **H**-shaped base measuring 3 feet on each side, a 4-foot vertical PVC pipe mast, and the 4-foot PVC structure supporting the loop (see Figure 4).

To fabricate each side of the **H** base, cut two 18-inch lengths of PVC pipe and cement one 18-inch section into each end of a PVC **T** connector. Cement PVC end-caps onto the ends of the 18-inch sections. The cross piece of the **H** base is identical to the side pieces, except the ends are finished with male thread-caps instead of end-caps. For the vertical mast, cut a 4-foot length of PVC. Cement a male thread cap to the bottom of the mast (which is threaded to the **H** base), and a female thread cap to the top of the mast (which is threaded to the antenna).

Motorizing the Loop

Franz Freller's photograph in *QST* shows

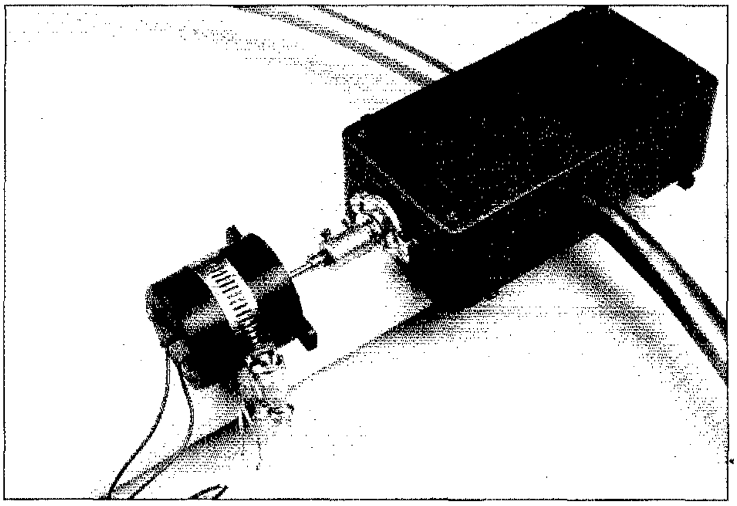


Figure 3—A small dc motor drives the capacitor. In this version, a reduction drive has been added to further decrease the tuning rate (see the sidebar "The Capon Loop and the ARRL Lab"). A shaft coupling links the motor to the capacitor. A 2½-inch-diameter hose clamp fastens the dc motor to the PVC mast. The motor control cable is inside the PVC mast and equipped with a phono plug for quick connection.

a manual lever he used to tune the capacitor. This works fine for many installations, but if you intend to operate the loop remotely, consider building a motor drive for the capacitor.

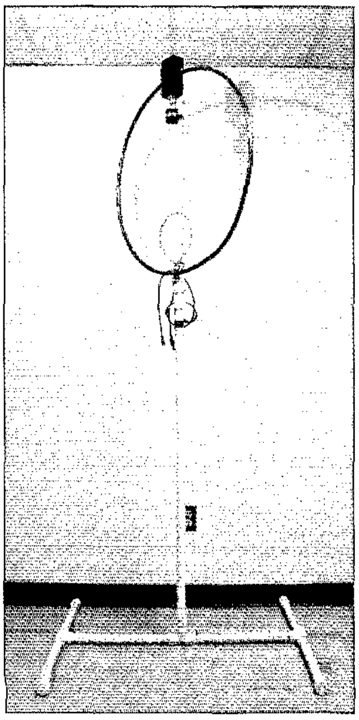


Figure 4—The antenna is supported by an H frame made of PVC pipe.

So I could tune the antenna's capacitor and measure SWR using a single control box, I built a motor-drive controller inside a Radio Shack SWR meter (21-524). I cut two holes in the top of the meter enclosure and installed two momentary contact DPDT toggle switches. I wired the switches (see Figure 5) so that they send positive and negative voltages to the motor drive to turn the motor in either direction.

I found a scrapped high-torque dc motor. The motor turns at about 1 rpm when powered by 2 AA cells. If you can't find the right motor in your scrapbox, Edmund Scientific[2] sells a high-torque 1-rpm dc motor for $22.50. There are two motor models: a 12-V motor (that can be driven by your transceiver's power supply) and a 3-V motor that can be powered by AA cells.

A shaft coupling links the motor to the capacitor. A 2½-inch-diameter hose clamp fastens the dc motor to the PVC mast. I ran the motor control cable inside the PVC mast and put a phono plug on the end for quick connection.

Safety Notes

The loop produces significant RF output, so please follow these precautions. Locate the antenna as far as possible away from people while it's in use and use the minimum power output necessary to maintain communication. Don't touch the antenna or the capacitor when transmitting! You can get an RF burn. For a thorough discussion of RF radiation safety, see Chapter 1 of *The ARRL Antenna Book*, or Chapter 36 of *The ARRL Handbook*.

The Capon Loop and the ARRL Lab

When the loop antenna was brought to the Lab for evaluation, I was excited about testing it because I'd done some antenna modeling of small loops using *ELNEC*.* I'd learned that a small loop at low heights above ground slightly outperforms a half-wave dipole at low angles of radiation (those best for DX). I was intrigued to see how a home-brew version of the popular small loop would work.

We first asked ARRL Laboratory Engineer Zack Lau, KH6CP, to do some testing for us. He set up the antenna in the large open space just south of the Headquarters building. Zack found that with the supplied capacitor, a good SWR could be obtained from 10 to 25 MHz, covering the 30- through 12-meter ham bands. (A capacitor with a lower minimum value of capacitance would allow coverage of the 10-meter band.)

I took the antenna home to give it a try on the air. Because the weather was cold and icy, I set up the antenna in my kitchen, about 20 feet from my shack. It tuned up nicely! I did find that the 1-rpm motor had a bit of overshoot, but it didn't take much practice to tune the antenna to nearly a 1:1 SWR. Although I was using a 9-V transistor battery to power the 12-V motor, the motor had more than enough torque to do the job, even at the lower voltage.

As Rob says, transmitter output powers greater than 7 W were too much for the tuning capacitor employed. But, I'm an avid QRPer, so I throttled the rig back to 5 W and called CQ on 14.060 MHz. Much to my surprise, my first CQ was answered by *two* stations! This antenna played! The band was fading fast, but I received a 559 signal report. A few other contacts proved that the antenna did indeed work. I returned the antenna to Headquarters the next day and bragged a bit about my QRP accomplishments.

Later, we decided to experiment and further decrease the tuning rate, so we added a 6:1 reduction drive salvaged from a dial drive. This made tune-up even smoother.

Reluctantly, I returned the loop to the author. I'd been thinking about a small loop antenna for portable work and HF mobile on my pickup truck. Playing with Rob's loop antenna for a couple of days convinced me: This antenna is for me!—*Ed Hare, KA1CV, ARRL Lab Supervisor*

*Available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.

Operating Results

I set up my antenna indoors vertically on its short PVC mast. When I tuned up the antenna, I found that incoming signals were on par with my full-size multiband vertical antenna that's mounted outside on a 20-foot mast.

The loop has a narrow bandwidth of about 20 kHz on 20 meters between the 2:1 SWR points and a progressively broader bandwidth on the upper bands. As expected, the antenna is quite directional, so I can null out interfering stations by simply rotating the antenna.

I use the antenna with my little MFJ-9020 transceiver, powered by a solar-charged gel cell, running about 3.5 W. My first two contacts using the loop on 20 meters were Z32RC in FYR Macedonia and I8WWV in Italy.

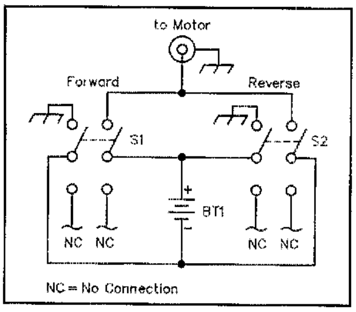


Figure 5—Schematic of the motor drive circuit. A pair of DPDT momentary contact switches supply positive and negative voltages to turn the motor forward and backward.

Summary

For amateurs who are restricted from using outdoor antennas, this miniature loop antenna is an excellent alternative. Because it takes up very little space and sets up in minutes, it's great for Field Day and other portable use. The antenna works well on 12 through 30 meters and I'm sure a 20- through 40-meter version could be built.

Acknowledgment

My special thanks to Franz Freller, DL9RBT, for inspiring this project and for providing me with the necessary background information on how to assemble this miniature loop antenna.

Also, thanks to Bill Hutchins, KM4UO, who assisted me with the construction of the antenna.

Notes

[1] F. Freller, DL9RBT, "Up Front," *QST*, Dec 1992, p 11.

[2] Edmund Scientific Co, 101 E Gloucester Pk, Barrington, NJ 08007-1380; order tel 609-547-8880, fax 609-573-6295; customer service tel 609-573-6260.

New Books

SPACE SATELLITE HANDBOOK

By Anthony "Tony" Curtis, K3RXK

Gulf Publishing Company, PO Box 2608, Houston, TX 77252-2608; tel 713-529-4301. Third Edition, 1994, 346 pp; B&W diagrams, illustrations, tables; 8½x11 inches, hardcover. Retail $39.

Reviewed By Steve Ford, WB8IMY
Assistant Technical Editor

The third edition of the *Space Satellite Handbook* is a rare and pleasant discovery. It's one of those few references that won't put you to sleep five minutes after you open the cover. On the contrary, Tony Curtis keeps you going with fascinating tidbits of knowledge and a conversational narrative.

For example, he doesn't simply tell you that there are more than 100,000 manmade pieces of space junk orbiting our planet. To add spice to such potentially dry information, Tony describes what happened when some larger pieces took the big plunge homeward. (Australia seems to be a favorite target for orbital bombardment.) He also discusses how the junk got there in the first place. (Like the screwdriver that got away from a Russian cosmonaut a few years ago.)

The *Space Satellite Handbook* makes the job of understanding satellites easier by separating them into groups with corresponding chapters: communication, search-and-rescue, weather, earth-observing, navigation, military science, manned and extraterrestrial. Amateur Radio satellites are found in the communication-satellite chapter. Tony devotes 22 pages to past, present and future ham satellites. Because this book is intended for a less technical audience, you won't find great detail about power systems, transponders, ground-station requirements and so forth. Instead, he focuses on brief biographies of each bird. The information in this section is accurate, except where he speaks of a few current satellites in the future tense—as though they weren't in orbit yet. The information concerning Phase 3D is out of date because of recent design changes. These errors are understandable when you consider the rapid progress the Amateur Radio satellite program has enjoyed during the past several years. Keeping up with the ever-changing world of ham satellites is any author's nightmare.

As you read each chapter, you can't help but pause and say, "Hey, I didn't know that!" Did you know that the Russians plan to orbit a replacement for the *Mir* space station in 1997? I didn't until I read the manned-satellite chapter. Until I browsed through the Satellite Scorecard on page 82, I was unaware that Luxembourg has three payloads in orbit (launched by other nations). You've probably heard of the NAVSTAR global positioning satellites (GPS), but do you know how much they've revolutionized the world of mapmaking? A sidebar on page 101 tells the story of how GPS has caused cartographers to revise maps they once thought were highly accurate. In one example, the position of a flagpole on a topographical map of Honolulu had to be moved 1480 feet to the southeast!

It's unfortunate that Tony's attention to NAVSTAR, the current DoD satellite navigation system, caused him to shortchange the earlier, but excellent, NAVSAT system, which he addresses mainly at the individual satellite level, rather than at the system operational level. Even though it was the very first satellite navigation system, NAVSAT was so good that it remained in operational use by DoD for 30 years, it was used over a similar period by the US Coast and Geodetic Survey and others throughout the world for precision cartography, and it continues to serve commercial and recreational users to the present day. NAVSAT surveys determined mapping errors such as the misplacement by cartographers of the Australian subcontinent by a few hundred meters. Depending on the receiving system used and the time duration over which observations are made, NAVSAT can determine position to accuracies of a few centimeters, which is quite a contrast to Tony's quote of best accuracy of 4.9 feet. [NAVSAT was conceived, developed, and managed for the US Navy by the Applied Physics Laboratory of The Johns Hopkins University, which is very near Tony's home.—*K3KMO*]

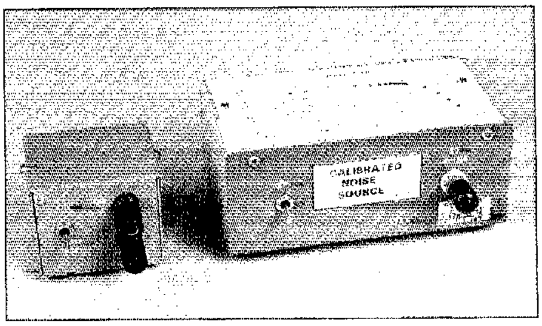
More than 140 pages are set aside for a master list of all satellites—those presently in orbit and those that have met their fate in the atmosphere. Each satellite is cataloged by the year it was launched, its international designation, its name, its country of origin and its launch date. In most cases, Tony includes basic information about each bird's orbit (period, inclination, apogee altitude and perigee altitude). If a satellite is in orbit now (or ever was in orbit), you'll find it in this list.

Tony Curtis has done a stellar job (forgive the pun) with the *Space Satellite Handbook*. This is a must-have reference for anyone with an interest in space technology. It makes a great coffee table or bedside book, and I suspect it would be invaluable for high school and college students. The most serious flaw I could find was the lack of photography. The history of satellites is full of exciting photos and I'm sure a few would add to the atmosphere of the book. Even so, Tony's tight writing style fills the gaps. Your imagination can do the rest.

A Calibrated Noise Source for Amateur Radio

Calibrated and stable noise sources are expensive —but not this one! Here's a reliable unit *you can build* at a quite reasonable cost.

By William E. Sabin, W0IYH
1400 Harold Dr, SE
Cedar Rapids, IA 52403



W0IYH's calibrated noise sources. The smaller 1 MHz to 2.5 GHz unit is to the left of the 0.5 to 500 MHz noise source.

Most hams know about the noise sources included in RF bridges that are used to measure impedances and adjust antenna tuners. A somewhat different device—an *accurately calibrated and stable* noise source—is also useful. If you combine a broadband RF noise source of known power output and known output impedance with a true-RMS voltmeter, you have an excellent instrument for making interesting and revealing measurements on a variety of circuits hams commonly use. (Later on, I'll identify some examples.) The true-RMS voltmeter can be an RF voltmeter, a spectrum analyzer, or an AF voltmeter[1] at the output of a linear receiver.

Calibrated noise generators and noise-figure meters are available at medium to *astronomical* prices. Here, I'll describe a *low-cost* approach you can use with reasonable confidence for many amateur applications where accuracy to tenths of a decibel is not needed, but where precision (repeatability) and comparative measurements are much more important. PC boards are available for this project.[2]

Semiconductor Noise Diodes

Any Zener[3] diode can be used as a source of noise. If, however, the source is to be calibrated and used for reliable measurements, avalanche diodes specially designed for this purpose are preferable by far. A good noise diode generates its noise through a carefully controlled *bulk avalanche*[4] mechanism, which exists *throughout* the PN junction, not merely at the junction surfaces where unstable and unreliable surface effects due to local breakdown and impurity effects predominate. A true noise diode has a very low *flicker noise* (1/f) effect and tends to create a uniform level of truly *Gaussian noise*[5] over a wide band. In order to maximize its bandwidth, the diode also has very low junction capacitance and lead inductance.

For this project, I used the NOISE/COM NC302L diode. It's in a glass, axial-leaded DO-35 package and rated for use from 10 Hz to 3 GHz, if appropriate construction methods are followed. Prior to sale, the diodes are factory-aged for 168 hours and are well stabilized. NOISE/COM[6] has kindly agreed to make these diodes available to amateur experimenters for the special price of $10 each, as compared to the usual low-quantity price of about $25.

Noise-Source Design

The noise source presents two kinds of available output power. One is the thermal noise (−174 dBm/Hz at room temperature) when the diode is turned off; call this N_{OFF}. The other is the sum of this same thermal noise and an "excess" noise, N_E, which is created by the diode when turned on; call this N_{ON} (equivalent to $N_{OFF} + N_E$). For accurate measurements, the output impedance of the test apparatus must be the same whether it is on or off, so that the device under test (DUT) always sees the same generator impedance. In Amateur Radio work, this impedance is usually 50 Ω, resistive. The circuit design must guarantee this condition.

For maximum frequency coverage, a PC-board layout and coax connector suitable for use at microwaves are needed. For lower-frequency usage, a less-stringent approach can be employed. Two noise sources are presented here. One is for the 0.5 to 500 MHz region and uses conventional components that many amateurs already have. The other is for the 1 MHz to 2.5 GHz range; it uses chip components and an SMA connector.

Circuit Diagram and Construction

Figures 1A and 1B are the simple schematics of the two noise sources. In series with the diode is a 46.4-Ω resistor, which, when combined with the dynamic resistance of the diode in the avalanche noise-generator mode (about 4 Ω), totals about 50 Ω. When the applied voltage polarity is reversed, the diode is forward conducting and its dynamic resistance is still about 4 Ω, but the avalanche noise is now turned off. As a result, the noise-source output impedance is always about 50 Ω. The 5-dB pad reduces the effect of any small impedance differences, so that the output impedance is nearly constant from the *on* to the *off* condition and the SWR is less than 2:1.

We must consider the noise situation of the noise diode when it is forward conducting. The resistance of the forward-biased PN junction is a *dynamic* resistance. This dynamic resistance is *not* a source of thermal noise, since it is not an actual physical resistance, such as in a resistor or lossy network. However, the 0.6-V forward drop across the PN junction does produce a shot-noise effect. The mathematics of this shot noise shows that the noise power associated with this effect is only about 50% of the thermal noise power that would be available from a physical resistor having the same value as the dynamic resistance. Therefore, the forward-biased junction does *not* add excess noise to the system.[7] There is a 1/f noise effect associated with this shot noise in the NC302L diode, but its corner frequency is at about 100 kHz and of no importance at higher frequencies. Also, the small amount of bulk resistance contributes a little thermal noise.

In order to maximize the unit's flatness and frequency response bandwidth, noise-source construction methods should aim for

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[1] Notes appear on page 40.

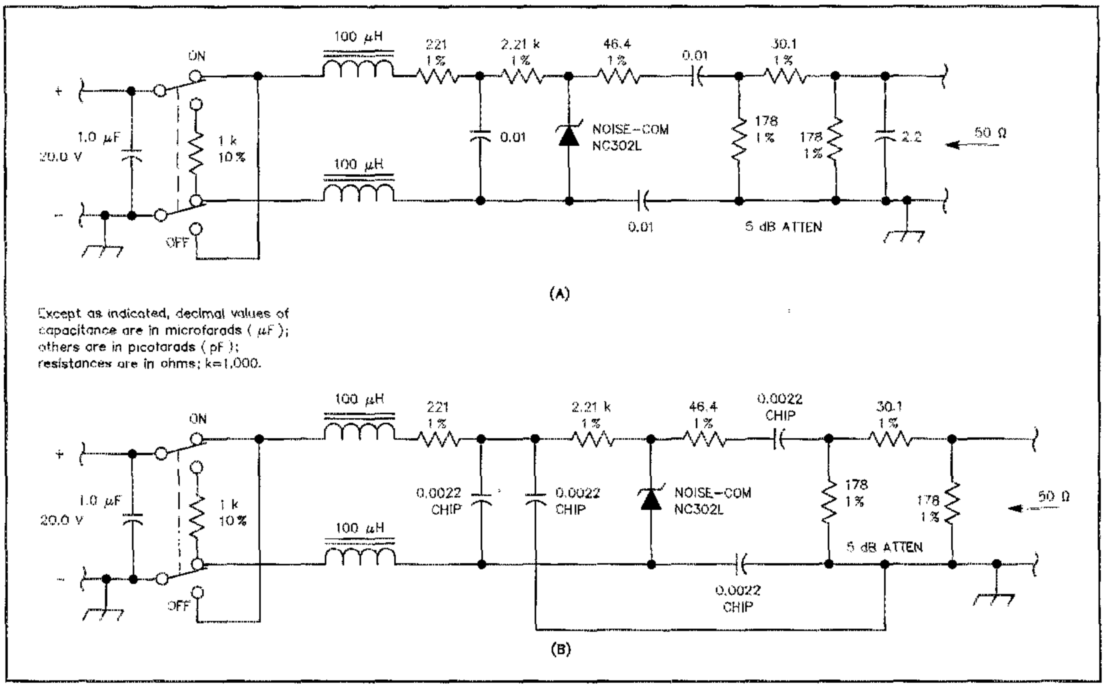

Figure 1—Schematics of the two noise sources. At A, the 0.5 to 500 MHz unit. Resistors are 1/8-W, 1%-tolerance metal-film units. The 1 MHz to 2.5 GHz unit at B uses 1% tolerance, 0.1-W chip resistors and chip capacitors.

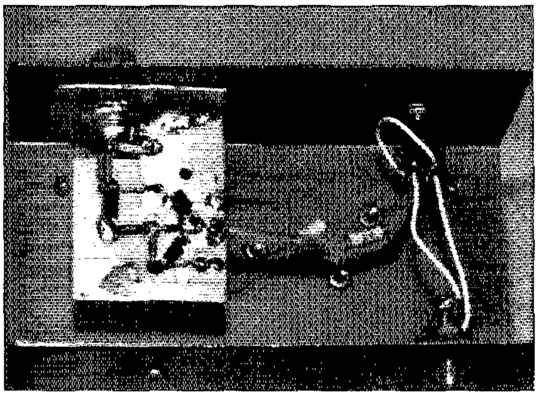

Figure 2—An inside view of the 0.5 to 500 MHz noise source.

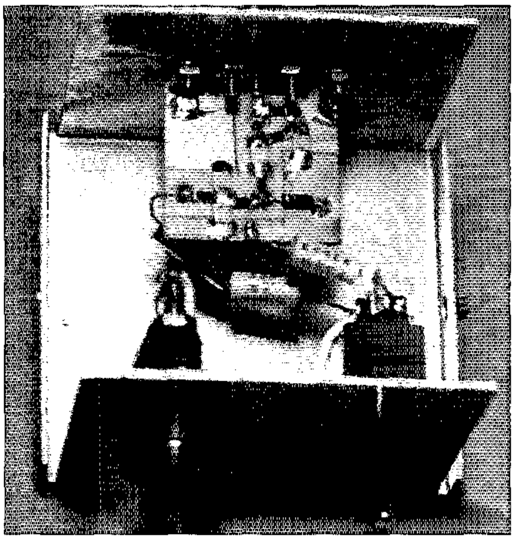

Figure 4—View of the inside of the 1 MHz to 2.5 GHz noise source.

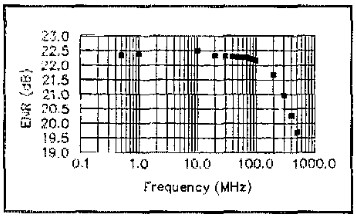

Figure 3—Sample calibration chart of excess noise ratio (ENR) versus frequency for the 0.5 to 500 MHz noise source.

RF circuit lead lengths as close to zero as possible, and minimum inductance in the ground path and the coupling capacitors. The power-supply voltage must be clean, well bypassed and set accurately. Figure 2 shows my 0.5 to 500 MHz unit. This construction method satisfies quite well the electrical requirements I wanted for this model. At 500 MHz, the return loss with respect to 50 Ω at the output jack decreased to 10 dB and I didn't try to extend the calibration beyond that frequency. A calibration chart (Figure 3) is attached to the unit's top for easy reference. Figure 4 shows the

Frequency (MHz)	0.5 to 500 MHz Unit ENR (dB)	SWR	1.0 to 2500 MHz Unit NR (dB)	SWR
0.5	22.33	1.03		
1	22.38	1.03	21.38	1.03
10	22.45	1.04	21.46	1.03
20	22.35	1.06		
30	22.32	1.06		
40	22.32	1.09		
50	22.30	1.11		
60	22.29	1.12		
70	22.25	1.15		
80	22.22	1.17		
90	22.20	1.20		
100	22.15	1.23	21.80	1.07
200	21.65	1.42		
300	20.96	1.62		
400	20.25	1.70		
500	19.60	1.90	20.71	1.44
1000			20.12	1.86
1500			20.00	2.06
2000			20.70	2.14
2500			21.51	1.88

Figure 5—NOISE/COM calibration for both of my noise sources. The data is, of course, not universal; it varies from unit to unit.

inside of the 1 MHz to 2.5 GHz noise source.

Calibrating the Noise Source

If the construction is solid, the calibration should last for a long time. There are two ways to calibrate the noise source. *If the unit has been carefully constructed and its correct operation verified,* NOISE/COM will calibrate home-built units over the desired frequency range for $25 plus return shipping charges. Note that one factory-calibrated unit can be used as a reference for many home-calibrated units. Figure 5 shows the NOISE/COM calibration data for both models of *my* noise sources, including SWR data. The noise data is strictly valid only at room temperature, so it's necessary to avoid extreme temperature environments.

The second calibration method requires a signal generator with known output levels at the various desired calibration frequencies. One approach is to build a tunable weak-signal oscillator[8] that can be compared to some accessible high-quality signal generator, using a sensitive receiver as a detector. The level of the signal source in dBm is needed.

Access to a multistage attenuator[9] is also desirable. If you build the attenuator, use the nearest 1% values of metal-film resistors so that systematic errors are minimized. A total attenuation of 25 dB in 0.1-dB steps is desirable. Attenuator construction must be appropriate for use at the intended frequency range. In some cases, a high-frequency correction chart may be needed.

With the calibrated signal source and the attenuator feeding your receiver in an SSB or CW mode, use the techniques discussed in the referent of Note 1 to determine the excess noise (N_E) of the noise source and the noise bandwidth (B_N) of the receiver.

Excess Noise Ratio

A few words about excess noise ratio (ENR) are needed. It is defined as the ratio of excess noise to thermal noise. That is,

$$ENR = \frac{N_{ON} - N_{OFF}}{N_{OFF}} = \frac{N_E}{N_{OFF}} \qquad \text{(Eq 1)}$$

When the noise source is turned on, its output is $N_{OFF} + N_E$. The ratio of N_{ON} to N_{OFF} is then

$$\frac{N_{ON}}{N_{OFF}} = \frac{N_{OFF} + N_E}{N_{OFF}} = 1 + \frac{N_E}{N_{OFF}}$$

$$= 1 + ENR \qquad \text{(Eq 2)}$$

Therefore, ENR is a measure of how much the noise increases, and the noise generator can be calibrated in terms of its ENR.

Normalizing ENR to a 1-Hz bandwidth and converting to decibels, this is

$$ENR\ (dB) =$$

$$174\ (dBm/Hz) + \frac{N_E\ (dBm)}{B_N\ (Hz)} \qquad \text{(Eq 3)}$$

Prepare a calibration chart and attach it to the top of the unit (see Figure 3). If you decide to have the unit factory calibrated, first perform the calibration procedure so you're fairly sure everything is working properly. Remember, a factory calibrated unit can be used as a reference for other home calibrated units, once you've worked out the calibration-transfer procedure according to your lab capabilities. This requires some careful thinking and proper techniques. If you have any doubts, a NOISE/COM calibration is the best choice.

Noise-Figure Measurement

The thermal noise power available from the attenuator remains constant for any value of attenuator setting. But the excess noise, and therefore the ENR (in dB) due to the noise diode, is equal to the calibration point of the source minus the setting (in dB) of the attenuator.

The noise-figure measurement of a device under test (DUT) uses the Y method and the setup in Figure 6. If the DUT has a noise-generator input and a true-RMS noise-measuring instrument at the output, then the total output noise (including the contribution of the measuring instrument) with the noise generator turned off is

$$N_{OFF(TOT)} = kTB_N\, F_{TOT}\, G_{DUT}\, G_{NMI} \qquad \text{(Eq 4)}$$

where kTB_N is thermal noise, G_{DUT} is the gain of the DUT, G_{NMI} is the gain of the noise-measuring instrument and F_{TOT} the noise factor of the combination of the DUT and the noise-measuring instrument. When the noise generator is turned on, the output noise is

$$N_{ON(TOT)} = kTB_N\, F_{TOT}\, G_{DUT}\, G_{NMI} +$$

$$(ENR)kTB_N\, G_{DUT}\, G_{NMI} \qquad \text{(Eq 5)}$$

where the last term is the contribution of excess noise by the noise generator. Note that none of these values is in dB or dBm.

If we divide Eq 5 by Eq 4 and say that the ratio

$N_{ON(TOT)} \div N_{OFF(TOT)} = Y$,

then

$$\frac{N_{ON(TOT)}}{N_{OFF(TOT)}} = Y = \frac{F_{TOT} + ENR}{F_{TOT}}$$

$$= 1 = + \frac{ENR}{F_{TOT}} \qquad \text{(Eq 6)}$$

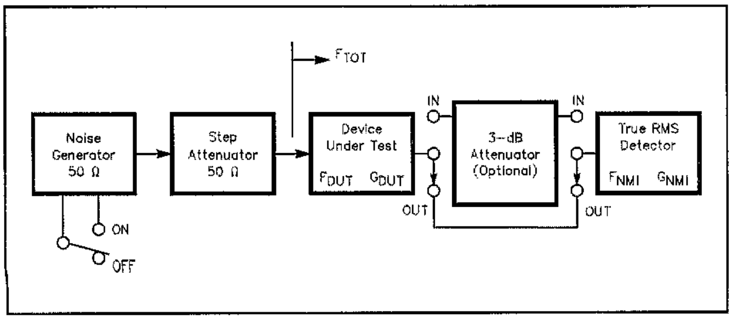

Figure 6—Setup for measuring the noise figure of a device under test (DUT).

Note that kTB_N, G_{DUT} and G_{NMI} disappear, so that these quantities need not be known to measure noise factor. If we solve Eq 6 for F_{TOT}, we get the noise factor

$$F_{TOT} = \frac{ENR}{Y - 1} \qquad \text{(Eq 7)}$$

If the noise output doubles (increases by 3 dB) when we turn on the noise source, then Y = 2, and the noise factor is numerically equal to the excess noise ratio (ENR). If the attenuator steps are not fine enough, or if the attenuator is not reliable over the entire frequency range, use Eq 7 to get a better answer. (It's much simpler to use a good fine-step attenuator.) The value of F_{TOT} is that of the DUT in cascade with the noise-measuring instrument. To find F_{DUT}, we must know the noise factor F_{NMI} of the noise-measuring instrument and G_{DUT} and then use the Friis formula, unless G_{DUT} is very large (as it would be if the DUT were a high-gain receiver [see Note 1]).

$$F_{DUT} = F_{TOT} - \frac{F_{NMI}}{G_{DUT}} \qquad \text{(Eq 8)}$$

The validity of Eq 8 (if we need to use it) requires that the noise bandwidth of the noise-measuring instrument be less than the noise bandwidth of the DUT (see the referent of Note 1). Verify this before proceeding.

There's another advantage to using the power-doubling method. If the 3-dB attenuator of Figure 6 is used to maintain a constant noise level into the following stages and the RMS meter, this means that the noise factor, using the calibration scale and the input attenuator (without using Eq 7), is

$$F_{DUT} = ENR + \frac{1}{G_{DUT}} \qquad \text{(Eq 9)}$$

If G_{DUT} is large, then the last term can be neglected. If G_{DUT} is small, we need to know its value. However, we do not need to know the noise factor F_{NMI} of the circuitry after the DUT, as we did in the previous discussion.

The 3-dB attenuator method also removes all restrictions regarding the type of noise-measuring instrument, since the meter reading is now used only as a reference point. This last statement applies only when two noise (or two signal generator) inputs are being compared.

Frequency Response Measurements

The noise generator, in conjunction with a spectrum analyzer, is an excellent tool for measuring the frequency response of a DUT, if the noise source is much stronger than the internal noise of the DUT and that of the spectrum analyzer. Many spectrum analyzers are not equipped with tracking generators, which can be quite expensive for an amateur's budget.

The spectrum analyzer needs to be calibrated for a noise input, if accurate amplitude measurements are needed, because it responds differently to noise signals than to sine-wave signals. The envelope detection of noise, combined with the logarithmic amplification of the spectrum analyzer, creates an error of about 2.5 dB for a noise signal (the noise is that much greater than the instrument indicates). Also, the noise bandwidth of the IF filter is different from its resolution bandwidth. Some modern spectrum analyzers have internal DSP algorithms that make the corrections so that external noise sources and also carrier-to-noise ratios, normalized to some noise bandwidth like 1.0 Hz, can be measured with fair accuracy if the input noise is a few decibels above thermal. One example is my Tektronix Model 2712. If only relative response readings are needed, then these corrections are not needed.

Also, the noise source itself can be used to establish an accurate reference level (in dBm) on the screen. An accurate, absolute measurement with the DUT in place will then be this reference level (in dBm), plus the increment in decibels produced by the DUT.

The noise-generator output can be viewed as a collection of sine waves separated by, say, 1 Hz. Each separated frequency "bin" has its own Gaussian amplitude and random phase with respect to all the others. So the DUT is simultaneously looking at a collection, or "ensemble," of input signals. As the spectrum analyzer frequency sweeps, it looks simultaneously at all of the DUT frequencies that fall within the spectrum analyzer's IF noise bandwidth. The spectrum display is thus the "convolution" of the IF filter frequency response and the DUT frequency response. If the DUT is a narrow filter, a very narrow resolution and a slow sweep are needed in the spectrum analyzer. In addition, the analyzer's video, or post-detection, filter has a narrow bandwidth and also requires some settling time to get an accurate reading. So, some experience and judgment are required to use a spectrum analyzer this way.

Using Your Station Receiver

Your station receiver can also be used as a spectrum analyzer. Place a variable attenuator between the DUT and the receiver. As you tune your receiver, in a narrow CW mode, adjust the attenuator for a constant reference level receiver output. The attenuator values are inversely related to the frequency response.

A calibrated noise source with an adjustable attenuator that can be easily switched into a receiver antenna jack is an excellent tool for measuring antenna noise level or incoming weak signal level (in dBm), or for establishing correct receiver operation.

The noise source can also be combined with a locally generated data-mode waveform of a known dBm value to get an approximate check on modem performance or to make adjustments that might assure correct operation of the system. The rigorous evaluation of system performance requires special equipment and techniques that may be unavailable at most amateur stations. Or, you could evaluate the intelligibility improvement of your SSB transmitter's speech processor in a noise background.

Summary

The calibrated, flat-spectrum noise generator described in this article is quite a useful instrument for amateur experimenters. Its simplicity and low cost make it especially attractive. Getting a good calibration is the main challenge, but once it is achieved, the calibration lasts a long time—if the right diode is used. The ENR of the units described here is in the range of 20 dB. You may want to use a high-quality, external, 10-dB attenuator barrel to get into the range of 10-dB ENR. If you send the unit to NOISE/COM, send the attenuator also and ask that it be included in the calibration. That attenuator then "belongs" to your noise source and should be so tagged. If the attenuator is of high quality, the output SWR will also be improved. Ask for calibrations with and without the attenuator, if you like. NOISE/COM suggests periodic recalibration, at your discretion.

Acknowledgments

I appreciate the review and suggestions, and also the collaboration with respect to diode purchase and calibration service, of Gary Simonyan and Bent Hessen-Schmidt at NOISE/COM.

Notes

[1] W. Sabin, "Measuring SSB/CW Receiver Sensitivity," *QST*, Oct 1992, pp 30-34. See also Technical Correspondence, *QST*, Apr 1993, pp 73-75.

[2] PC boards are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price, $3.50 plus $1.50 shipping. A PC-board template package is available free from the ARRL. Address your request for the SABIN NOISE SOURCE TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

[3] The term *Zener diode* is commonly used to denote a diode that takes advantage of avalanche effect, even though the Zener effect and the avalanche effect are not exactly the same thing at the device-physics level.

[4] The term *bulk avalanche* refers to the avalanche multiplication effect in a PN junction. A carrier (electron or hole) with sufficient energy collides with atoms and causes more carriers to be knocked loose. This effect "avalanches" and it occurs throughout the volume of the PN junction. This mechanism is responsible for the high-quality noise generation in a true noise diode.

[5] *Gaussian noise* refers to the instantaneous values that the noise voltage has. These values conform to the Gaussian probability density function of statistics.

[6] NOISE/COM Co, East 49 Midland Ave, Paramus, NJ 07652. Contact Gary Simonyan at 201-261-8797.

[7] Motchenbacher and Fitchen, *Low Noise Electronic Design*, (New York: Wiley & Sons, 1973), p 22.

[8] W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur*, (Newington: ARRL, 1986).

[9] *The ARRL Handbook for Radio Amateurs*, (Newington: ARRL, 1994), pp 25-37 to 25-39.

QST

Amateur Use of Telescoping Masts

Do you need an inexpensive skyhook for your antenna? Here's one candidate you may have overlooked.

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One of the many pieces of equipment designed for the TV industry is the telescoping mast used to support TV antennas. In some areas more than about 20 miles from a TV transmitter (and not served by a cable TV company), these masts are used frequently.

TV telescoping masts aren't commonly used by hams. Observation and inquiry show two reasons for this: (1) a lack of knowledge of the capabilities and the limitations imposed by telescoping-mast design; (2) having had (or heard of) a bad experience with such a mast. The latter seem to stem totally from mishandling during installation, leading to mast and antenna damage and—occasionally—injuries to people. The mishandling is directly related to a lack of knowledge regarding proper installation techniques.

Mast Design Limitations

It's important to know the limitations imposed by telescoping TV mast design. One major TV mast manufacturer's flier emphasizes this by stating "Telescoping masts are not recommended for commercial or ham installations" in the specification list. The "not recommended" ignores the fact that many amateur antennas are *smaller* than some TV antennas used in rural areas and that the mast alone can be used as a vertical antenna.

None of the ads, fliers or instructions packed with a telescoping mast provide adequate information about the capabilities and limitations of its design. The most I've seen is a short note to the effect that the mast should not be used with antennas having a wind load area of more than 2 square feet. In TV antenna ads, you're told the number of elements a particular antenna has, but not necessarily its wind load area or its weight. In most amateur antenna ads, this vital information is given. Telescoping-mast fliers and instructions also omit correct installation procedures and safety procedures. About the only precaution I've seen mentioned is to "stay away from power lines."

Here, I'll explore the capabilities of telescoping masts for amateur use. I'll provide data to allow you to determine correct use in your application, up to the limit of mast capability. Proper installation and use are also covered.

Telescoping-Mast Design Principles

A telescoping TV mast is much more than a few pieces of steel tubing. It includes a number of design features to make use simple and safe. These feature areas are shown in Figure 1, starting at the top of a mast section and working toward the bottom.

There's size reduction in the top section to bring its ID to a size slightly greater than the next smaller section, which is also the next higher section when the mast is extended. Partly, this is for strength, to distribute the load from the upper section. It's also an antirattle feature, to limit noise caused by wind moving the mast.

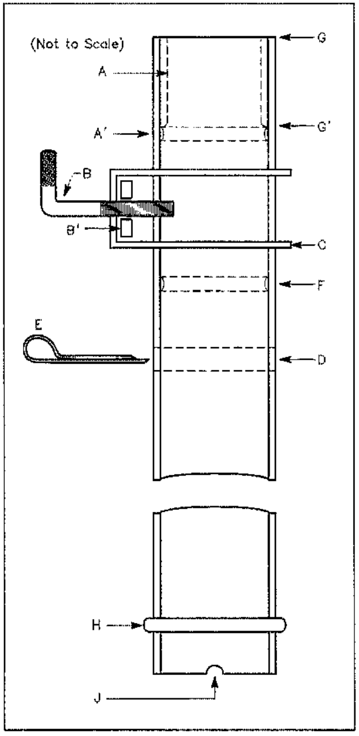


Figure 1—Major features of a single section of a typical TV telescoping mast section. See text for explanation of elements A through J.

In some designs, the top 4 to 6 inches of a section is roll-swaged to the smaller diameter as at A. In others, the swage may extend only $^1/_4$ to $^3/_8$ inch, as shown in Figure 1 at A'. In these, the swage is typically 1 to 2 inches below the top of the section.

Just below the swaged area are two holes in the section. The top one (B) is typically $^1/_4$ inch in diameter, and penetrates only *one side* of the tubing. This hole is for a clamping screw. The screw's primary use is to hold the smaller, inner-mast section in place while a new grip on this section is taken during mast erection. Secondly, the screw keeps the two sections tightly together to prevent wind noise. This screw is *not intended* to be the permanent—and only—mechanism to prevent slippage and mast collapse (more on that later).

The clamping screw (B) has some type of a support fixture. One style uses a simple strap, with a floating square nut for the mating screw. The nut is held in place by metal tabs. A second type (C) uses a flat, **U**-shaped metal sheet, with a hole on each of the **U**-shaped sides just larger than the tubing diameter. There's a captive nut (B') at the bottom of the **U**. Either type may have a small, internally threaded stud not shown here. This stud is used to mount an insulated standoff for carrying twin lead or coax, holding it away from the mast to reduce signal loss and wind-induced slapping noise.

The second, lower hole (D) is typically $^5/_{16}$ inch in diameter, and penetrates the tube completely, along the diameter. This hole accepts a large cotter pin (E), which is the support for the next-smaller mast section. (The best designs equip the cotter pin with a short chain to prevent the pin from being lost.) This pin is intended to carry *all of the weight above it:* tubing sections, rotator, antenna and guys. It's a safety feature: With overload, the pin should shear and the mast should safely telescope downward. As we'll see, this is the weak point in the design.

In designs I've seen, there's an additional, smaller-diameter swaging (F) between the two holes. The contact of this and the upper swage create a force couple, which transfers bending forces from the smaller section to the larger. It's also part of the antirattle design. Additionally, it's a safety feature (working with the next feature to be described) to prevent sliding the small section completely out of the larger one while raising the mast.

Rings are employed for attaching guy wires. Some designs make the hole diameter in a ring just greater than the diameter of the next-smaller mast section. The rings are used at the very top of a section, at G. Another design makes the hole in the ring just larger than the topmost swage outside diameter. This slides over the section, and rests on the shoulder of the swage, at G'. Another design makes the ring hole larger than the tubing. This ring is kept from sliding down the mast by a weld bead around the mast at A'. The ring located at G is the least desirable since the ring tends to jam if the mast is telescoped downward. The ring size varies with the section diameter.

Glossary of Terms Used in the Tables

ALLOW—allowable load.
AREA—Section area exposed to wind.
A WND LD—allowable wind load.
BOLT LD—maximum shear load on the bolt.
COTR LD—maximum shear load on the cotter pin.
DIAM—outside diameter of a section.
GUY LEN—guy length.
GUYS—length of three guys.
GUY COMP—compressive load on a mast due to guy tension.
GUY LD—guy load.
GUY STR—guy strength.
H WND LD—horizontal wind load.
HGT—overall height.
LAT AREA—cross section of tubing.
LEN—length.
LEN 3 GUYS—length of three guys.
LEN 4 GUYS—length of four guys.
LIMT LD—section load at failure limit.
MAX AREA—maximum area of antenna and rotator.
MAST WND—mast wind load.
MAX V LD—maximum vertical load on section.
NOM WT—nominal weight of a section.
PANEL—a section of mast.
RAD GYR—radius of gyration of section.
SAFE LD—maximum safe load (with a safety factor of 4).
STAT LD—static load.
THICK—thickness of the section wall.
TRY—a trial value.
V WND LD—vertical wind load.
WIND—wind load on projected area.
WND LD—dynamic wind load.
WT—weight.
WT MAST—weight of the mast.
WT GUYS—weight of the guys.

There are two types of guy rings. One is a simple, flat plate, typically 1/8 inch thick. The other is thinner, with a lipped inner hole, and with the outer edge roll-crimped. It's claimed that the rolled edge eliminates the need for a guy thimble to keep the guy from chafing through. Prudence indicates the use of thimbles, which are *not* supplied with any masts I've seen.

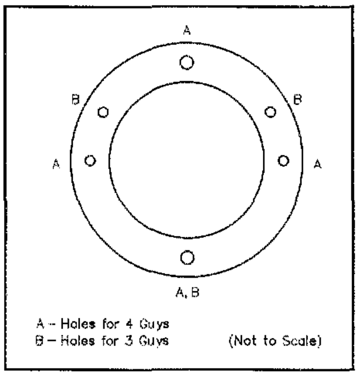


Figure 2—Top view of the guy ring plate found at the top of each section of a telescoping mast. The hole arrangement allows use of three or four sets of guys.

These guy rings have holes for six guy wires, spaced as shown in Figure 2. They allow use of three sets of guys (commonly); four sets are rarely used.

Just above the bottom of the mast section, at H in Figure 1, is a weld ring. The size of this weld ring is controlled so that its outside diameter is just smaller than the inside diameter of the next-larger section. This acts with the swage of A or A' to take the bending moment when extended. It also prevents the inner tube from extending past the swage at F, preventing overextension.

At J is a pair of notches in the tube bottom. These are half circles, 5/16 or 3/8 inch across. They're intended to receive the cotter pin and prevent the mast section from rotating. The notches also distribute the shear load presented by the weight of the upper parts of the mast system.

After the sections are finished—with all holes punched or drilled and all welds made—they are hot-dipped galvanized. (Or at least, they should be!) I've seen one mast of unknown manufacture that was zinc-plated rather than hot-dip galvanized. I've been told paint-dipped masts exist. Such treatments are *not recommended*, as rust makes for a short mast life.

Mast Specifications

Telescoping masts are manufactured in three weights. The light-duty mast is formed from 18-gauge steel, which has a nominal

Table 1
Telescoping Masts

Quantity	*Unit of Measure*	*Section 1*	*2*	*3*	*4*	*5*
LEN	ft	10	10	10	10	10
HGT	ft	10	19	28	37	46
PANEL	ft	9	8	8	8	9
THICK	in.	0.05	0.05	0.05	0.05	0.05
DIAM.	in.	2.25	2.00	1.75	1.50	1.25
AREA	in.^{2}	0.35	0.31	0.27	0.23	0.19
GUY LEN	ft	25.08	29.83	36.24	43.57	51.43
LEN 3 GUYS						558.43
LEN 4 GUYS						744.57
LAT AREA	ft^2	1.88	1.67	1.46	1.25	1.04
NOM WT	lb	15	13	12	10	8
WIND	lb	37.50	33.30	29.20	25.00	20.80
RAD GYR	in.	0.98	0.87	0.76	0.66	0.55
MAX V LD	lb	4070.26	3659.60	2903.26	2176.73	1281.30
With Cotter Pins						
COTR LD	lb	745.14	745.14	745.14	745.14	745.14
LIMT LD	lb	745.14	732.14	720.14	710.14	702.14
SAFE LD	lb	186.28	183.03	180.03	177.53	175.53
With Stainless-Steel Bolts						
BOLT LD	lb	2943.75	2943.75	2903.26	2176.73	1281.30
LIMT LD	lb	2943.75	2930.75	2891.26	2166.73	1273.30
SAFE LD	lb	735.94	732.69	722.82	541.68	318.32

Characteristics of typical mast sections. At the top of the table are the dimensions and weights of the five sections of a typical 50-foot mast. Derived quantities of wind area, section radius of gyration and section strength considered as a column are shown. The bottom parts show the shear strength and safe vertical load on the section for two types of retaining pins.

thickness of 0.049 inch. The heavy-duty mast is 16 gauge, with a nominal thickness of 0.063 inch. There is also an intermediate-weight mast, which uses 16-gauge steel for the top section, and 18-gauge for all other sections. This reduces the weight and cost of the complete assembly; it does not reduce the carrying capacity of a complete "as supplied" mast appreciably, but it does reduce the safety factor.

Because it seems likely that most amateur installations need the best possible mast, only the tallest, heavy-duty type is considered here.

Section specifications from the catalog of one large manufacturer are shown in Table 1. (See the "Glossary of Terms" for an explanation of the abbreviations used.) This 5-section design is often called a "50-foot" mast, but has a maximum height of 46 feet. One foot of height is lost in the overlap between each two sections. The table gives the overall height, the section length exposed to wind, and the panel length, the distance where there is no added strength from overlap with higher or lower sections. The weight of each section and its radius of gyration is shown, as well as the projected area and the wind loading for a 20 lb/ft^2 wind. This corresponds to a wind speed of 70.7 mph. This is selected as adequate for short-term use, such as a Field Day installation. The wind loading should be increased for permanent installations. Acceptable design values are:

SE Florida, Cape Hatteras	50 lb/ft^2
SE USA coasts, some other areas	40 lb/ft^2
Rest of the contiguous states	30 lb/ft^2

Load capability data shown later is based on the 20 lb/ft^2 value.

A Standard Installation

The telescoping mast is designed to have a set of guy wires reaching from the ground to the top of each mast section. There are, of course, many mast heights, arrangements of these guys, as well as many ways of mounting antennas on the mast, and many antenna sizes. To reduce the analysis to reasonable size, a standard antenna mast installation is assumed.

We'll use a five-section mast with guys extending from the top of each section to a common point, as shown in Figure 3. Assume the common point to be located one-half of the total mast height from the base of the mast. All antenna and rotator weight is assumed to be located at the top of the mast: There are no intermediate antennas.

The standard guys are seven-strand galvanized wire common in TV installations. These guys are specified to have a tensional breaking strength of 910 lb and weigh 31.8 lb per 1000 feet. However, guy strength was found to be one of the limiting factors, so other material was also tried, as described later. Guys were assumed to be nonelastic: not changing length under load. I also assumed a negligible pre-load on the guys. Very closely, this corresponds to the precept that the guys should look tight and feel loose.

The mast is assumed to be as described in Figure 3, and to be in new condition. Lacking other information, common material strength values are used.

Basis for Analysis

The foregoing assumptions allow analysis of each mast section as if it is totally independent of the others. The basic force diagram for the top section is shown in Figure 4. The horizontal force to the left is the wind load on the antenna and rotator and on the mast section. At the right is the slanting guy tension under load. This load is resolved into horizontal and vertical components.

Additional vertical loads on the mast are the weight of the antenna and rotator, and the weight of the mast section. The wind load on the mast is resolved into two components, one-half at the upper-guy attachment point, the rest being at the next guy, and not affecting this section. The wind load on the guy is neglected. The horizontal component of guy tension is equal to the sum of the wind loads at the upper guy. The vertical component then appears as a compressive load on the mast, adding to the weight of antenna, rotator and mast section, and guy. The relation between these components are:

Compression/Wind load = Guy height ÷ Guy base. For the top section and the installation assumed, the compression is just twice the total wind load. Guy tension = SCORE (Compression × Compression +

Table 2
Telescoping Mast 2

Quantity	*Unit of Measure*	*Weight, Antenna and Rotator, lb* 10	*20*	*30*	*40*	*50*	*100*
WT MAST	lb	9	9	9	9	9	9
WT GUYS	lb	0.5	0.5	0.5	0.5	0.5	0.5
STAT LD	lb	19.5	29.5	39.5	49.5	59.5	109.5
MAST WND	lb	5.2	5.2	5.2	5.2	5.2	5.2
1/8-inch Steel Guy and Cotter Pin							
ALLOW	lb	175.5	175.5	175.5	175.5	175.5	175.5
GUY COMP	lb	156.0	146.0	136.0	126.0	116.0	66.0
TRY GUY	lb	174.4	163.2	152.1	140.9	129.7	73.8
GUY STR	lb	227.5	227.5	227.5	227.5	227.5	227.5
GUY LD	lb	175.5	163.2	152.1	140.9	129.7	73.8
V WND LD	lb	157.0	146.0	136.0	126.0	116.0	66.0
H WND LD	lb	78.5	73.0	68.0	63.0	58.0	33.0
A WND LD	lb	73.3	67.8	62.8	57.8	52.8	27.8
MAX AREA	ft^2	3.7	3.4	3.1	2.9	2.6	1.4
3/16-inch Dacron Line and Bolt							
ALLOW	lb	318.3	318.3	318.3	318.3	318.3	318.3
GUY COMP	lb	298.8	288.8	278.8	268.8	258.8	208.8
TRY GUY	lb	334.1	322.9	311.7	300.5	289.3	233.4
GUY STR	lb	687.5	687.5	687.5	687.5	687.5	687.5
GUY LD	lb	334.1	322.9	311.7	300.5	289.3	233.4
V WND LD	lb	298.8	288.8	278.8	268.8	258.8	208.8
H WND LD	lb	149.4	144.4	139.4	134.4	129.4	104.4
A WND LD	lb	144.2	139.2	134.2	129.2	124.2	99.2
MAX AREA	ft^2	7.2	7.0	6.7	6.5	6.2	5.0

Load and wind area spreadsheet calculation of allowable antenna area for various antenna plus rotator weights. The top part of the table shows the dead weights and mast wind loads. The lower parts combine these vectorially to obtain the allowable wind force and the antenna area, for the two types of support pins used. Note the large difference in antenna weight in the last two columns.

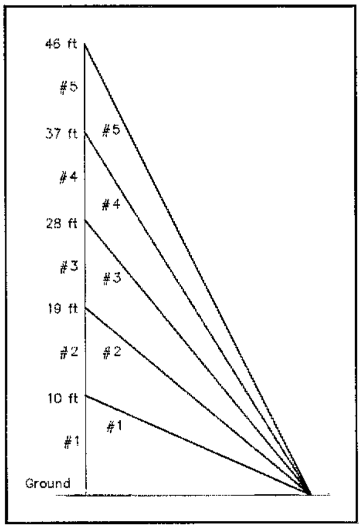


Figure 3—Assumed mast and guy installation. See text for assumption as to guy location.

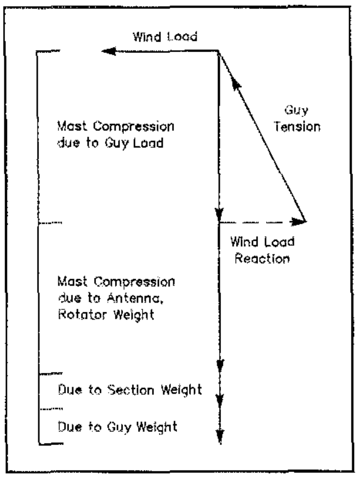


Figure 4—Load vector diagram of top mast section. Lateral wind load is opposed by horizontal component of guy tension. Its vertical component plus dead weights form the compressive load on the mast section. See text for relations.

Wind load × Wind load). For the top section, with the dimensions assumed, the guy tension is 1.118 times the total wind load.

We also need some strength values. The mast section is acting as a column under compression. Since the length/diameter ratio is 9 × 12 ÷ 1.25, or 86.4, the column is a short column. The compression strength is:

Max load = 17000 − 0.485 × length × length ÷ (RadG × RadG) for typical steels. Length is the length of the panel or part of the mast that has no overlap from another section. RadG is the radius of gyration of the section about its long axis. For the thin-wall tubing of these masts:

$$\text{RadG} = (\text{OD} - \text{thickness}) \div 2 \qquad \text{(Eq 1)}$$

where OD is the outer diameter of the section.

The other relation we need is the shear strength of the cotter pin. This is:

$$\text{Max pin load} = 15000 \times \text{Pin area} \qquad \text{(Eq 2)}$$

For a 3/16-inch diameter cotter pin, the maximum load is 745 lb (there are two bearing points on the cotter pin, each in single shear).

The allowable working load is equal to the maximum load divided by the safety factor desired. The required safety factor is partly a matter of laws of strength loss due to deterioration, the fact that the largest load encountered grows with time, and engineering judgment. For simple structures where safety is not a direct concern, a safety factor of 4 is often used.

Allowable Section Loads

The relations just given were easily transferred to a spreadsheet for calculation (see Table 1). The last row of the top part is the maximum allowable vertical load on the section, considered as a column in compression.

The second part of this calculation sheet is for safe vertical load with cotter pins. The top row is the allowable shear load on a 3/16-inch-diameter cotter pin. Just below this is the limit load, in all these cases that of the cotter pin. The safe compression load with a cotter pin is shown next. Note that any pin must support the weight of the sections above it.

Because the cotter pin is so limiting, I calculated the maximum shear load of a 1/4-inch stainless-steel bolt as shown in the bottom three lines of the table. This assumes 30,000 lb/in^2 as the bolt strength, so be certain the bolts are *stainless steel*. This allows a load increase factor of about 4. The limit load now is the compressive strength of the top two sections, then the bolt shear strength for the lower ones.

The analysis method assumes that the wind load on the mast itself is transferred to the points of guy attachment. Under conditions of very high winds—or if an intermediate antenna is installed on the mast—bending loads may be appreciable. Analysis of the affected mast section can be made using standard pinned-and-bending relations.

Allowable Antenna Weight and Area

From Table 1, the safe total load on the mast is established by the top section: 175 lb if cotter pins are used and 318 lb with stainless-steel bolts. This safe load must be apportioned into the static and dynamic load. The static load is that due to the weight of antenna, rotator, fittings and guys. The dynamic load is the changeable part, here due only to wind load. This is a way of saying that the allowable area of the antenna goes down as the installed weight goes up.

The easy way to do the apportionment is to assume a static load, then calculate the wind load that uses up the rest of the safe load allowance, using the equations presented earlier. This is easily done by a new spreadsheet, as shown in Table 2. Here, the columns are the values that result from the assumed antenna plus rotator weight, as listed at the top of the numerical column. The top section adds the weight of the mast section and guy, giving the total static load. For convenience, the mast wind load is also tabulated here.

The next group is for the standard installation. The limit strength of the top section is given first (cotter-pin shear). This minus the static load is the allowable compression due to wind. This is converted to a trial guy load, and compared to the guy strength. The smaller is taken as the guy load. This is resolved into the vertical (compression) and horizontal (wind) components. Half the total wind load on the mast section is subtracted to give the allowable antenna wind load, which is then converted to allowable antenna plus rotator wind area.

Table 3
Sample Antenna Types, Weights and Wind Load Areas

Antenna Type	*Weight (lb)*	*Area (ft^2)*
24 el, 2 m	4.5	2.3
3 el, 10 m	9.9	2.0
3 el, 10 m	18.0	3.1
2x24 el, 2 m	21.5	4.8
6 el, 6 m	26.0	4.8
3 el, tribander	27.0	4.4
3 el, 20 m	30.0	5.5
2 el, 40 m	44.0	6.4
4 el, 20 m	55.0	8.1

For small, light, antenna/rotator combinations, the guy strength dominates the dynamic load, limiting the allowable area to 3.7 ft^2 At 50 lb of dead weight, the allowable wind area decreases to 1.4 ft^2.

Remember that the assumed wind load is only 20 lb/ft^2, two-thirds the recommended permanent installation value for most of the country. It is easy to see why these masts are limited to small antennas for long-term use.

The lower part of Table 2 examines the result of using a stainless-steel bolt, and using a guy with a breaking strength of 2150 lb. This is the value for 3/16-inch Dacron line. The strength of 1/4-inch-diameter, seven-strand wire rope with a hemp core is essentially the same. With these changes, the mast can now be used for heavier antennas of greater wind area. The capacity of the standard condition has been doubled, or nearly so.

Several points are important if this extra capacity is used. The safety factor of the entire system has been reduced from normal use factors; in particular, the mast may collapse by guy failure or bending, rather than telescoping together. Additionally, the erection becomes more difficult as size and weight of antenna increase. Extra care is needed.

To get an idea of the range of usefulness of these telescoping masts, we need to look at antenna catalogs. Table 3 presents some figures from several well-known manufacturers. To these should be added the rotator weight: 5 to 10 lb for a small unit.

For temporary use, the small antennas for 10 meters and higher bands are nicely within limits of the "out of the box" mast. But even a lightweight tribander on a beefed-up telescoping mast is pushing the safety limit. You can probably get a full-sized 20-meter Yagi, or a loaded 40-meter Yagi up for a Field Day installation, but you're likely to lose the mast and antenna if a thunderstorm comes up. For permanent installations, a lightweight 10-meter beam should have extra guy and pin strength. Smaller antennas for the higher bands are okay with the standard mast. Be careful: If you must use a large antenna, use a larger mast.

The TV mast itself can be used as a vertical antenna with no additions. Add a set of three to six wires parallel to the mast, and up to a foot or so from it, to form a cage for low loss and improved bandwidth. A bottom insulator of fiberglass cloth and resin can be used, or the cage wires can be fed as a folded dipole. A 3, 6, or even 10-foot capacity-hat is within the design capability of these masts. The usual rules about good grounds apply.

A simple installation for Field Day use replaces the top guys with a pair of 40- and 80-meter dipoles, with end cord added to keep the antenna ends as far above ground as the site permits. With a lightweight tribander at the top, scores will depend more on operators than on installation limits.

Telescoping-Mast Installation

The telescoping mast is designed to be extended in only one way. *The mast must be vertical and the lower section properly guyed before extending the upper sections.* The bad reputation of the telescoping mast is almost completely due to attempts to ignore this simple rule. If you extend the sections, mount the antenna and then attempt to raise the assembly from a horizontal position to the vertical, you'll bend the mast. You may also damage the antenna and injure someone. In fact, the mast may bend if you try to raise it to the vertical while extended, just from its own weight—even if there's nothing mounted on the mast.

The recommended safe procedure is:

- Get a firm footing. For a typical Field Day installation on dirt, this means using a steel plate, with a spike on the bottom to penetrate the earth, and a pin on the top to keep the mast from slipping sideways. Base and tripod mounts for other surfaces are available.
- Mount the collapsed mast, plumb it to the vertical, and guy it with the bottom section permanent guys. A little pre-tension on these guys is a good idea.
- Tie one or two stepladders to the mast, for further work.
- Mount the rotator and antenna, and attach cables and the guys for the upper sections. Be sure that all fastenings are properly made and that the cables and guys won't tangle.
- If the antenna is small and light and there is no wind, an experienced person can get the antenna to a height of 30 feet. For greater heights, or with wind present, two to seven people are needed. Four handle the guys, two on ladders push up the sections and one person handles the clamping screws and the cotter pin or bolt. The guy handlers should not place strain on the guys, but should be prepared to keep the mast vertical. Practicing with the mast only, with no antenna in place, is a good idea.
- Starting with the top section, push it up a foot or two and tighten the clamp screw just enough to keep the section from slipping when released. Take another grip, and raise again, repeating until the stop is reached. Slip the cotter pin or bolt in place. Carefully lower the section until it rests on the pin or bolt. Turn the section until the slots engage. Then, tighten the clamp screw sufficiently to prevent rattling (wear a pair of good leather gloves). Don't use pliers—the screw will dent and deform the inner tube enough so that it will be impossible to lower the mast later.
- Repeat for the next lower section, and so on. If the mast starts to leave the vertical, or the wind picks up, stop, tie the guys temporarily and get help.
- Don't strain, and don't take chances. Even a few pounds falling from 10 or 20 feet can be deadly.

In principle, taking down a mast is easy: Just reverse the erection steps. If the mast is badly rusted, or the sections have been deformed by bending or by excessive clamping, however, this can be a chore—even impossible. If a little work doesn't get the antenna down to ladder height, consider bringing in a crane or ladder truck.

Some Precautions to Take

Always check local building codes when considering the installation of a large antenna or high tower or mast structure. Much trouble and expense can be avoided by staying within imposed limits.

If you're going to use these telescoping masts above their rated TV antenna size and load limits, I recommend running a load analysis with the weights and dimensions involved. When doing this, include such factors as the distance from the top of the mast to the antenna mounting point, and the weight and wind area of the rotator. I recommend you *measure* the size and thickness of the mast sections. Errors in filling orders have occurred.

Remember: Safety First!

Bob Haviland was first licensed as W9CAK in 1931. He obtained his BSEE degree at Missouri School of Mines in 1939, and his Professional Engineering license in New York.

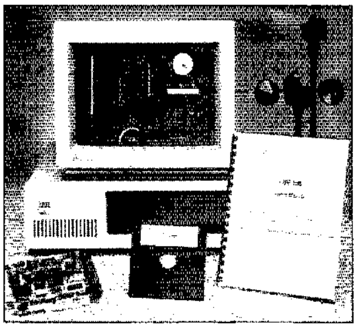
Bob was project engineer for the first radio transmission from beyond the ionosphere, in 1949, at White Sands, New Mexico, and for the first missile launching from Cape Canaveral, in 1951. He developed ablation (the use of material which goes from solid to gas state) for space-vehicle reentry protection, and initiated programs for recovery of equipment and data from space.

Bob's worked extensively on communication and broadcast satellite concepts, and played a founding role in commercial satellite communication. Between 1959 and 1972, Bob was a member of US delegations to the ITU and CCIR. He served as Chairman, subcommittee for 27 to 1215 MHz, FCC WARC Advisory Committee for Amateur Radio, 1976.

Bob is a Fellow of the Institute of Electrical and Electronic Engineers, the American Astronautical Society and the British Interplanetary Society. He holds eight US patents, and is author of many articles and 14 books, including The Quad Antenna, *CQ Publishing, 1993.*

New Products

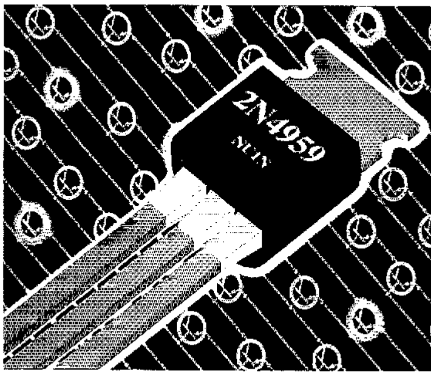
WEATHER AND DATA MONITOR INSTRUMENTS



◊ Weather observers, repeater control ops, packet operators and other hams interested in experimentation can check and record atmospheric conditions or other environmental factors remotely by using a PC to operate a remote Environmental Monitoring System. The ENV-100 directly connects to standard 4-20 mA or 0-1 V sensors for data-acquisition applications. It accepts as many as 200 modules linked to one PC serial port. The ENV-100 is a plug-in board for IBM-compatible personal computers and can be easily customized from the keyboard, with no external power supply or other interface device. The onboard clock time-stamps data for later evaluation. It provides a standard RS-232 (EIA-232)/485 multidrop sensor-to-computer interface; programmable setpoints; uses a complete ASCII command set; supports up to six analog-to-digital channels per module; holds 16k of nonvolatile memory for data logging and independent control (32k with $49 upgrade); and comes in a compact, weatherproof case for harsh environments. The ENV-100 can operate alarms, relays and controls via TTL outputs. A four-wire cable delivers power and communications. It operates on 7-30 V dc at 10 mA. An interface kit with *SYS50* software to graphically display, store and manipulate data on a PC, an ac power pack and 60 feet of cable is available for $59.

Remote modules gather data to log information on temperature, humidity, solar radiation, rainfall and barometric pressure (ENV-50-HUM); wind direction and velocity, rainfall and temperature (ENV-50-WDT); and electrical signals of 0-10 V, 0-1 V, 0-2.5 V, 0-100 mV, -5 to +5 V, 0-100 mA and 4-20 mA (ENV-50-VOL). Retail prices: ENV-50-WDT $395, ENV-50-HUM $395, ENV-50-VOL $379, rain gauge with 60 feet of cable $85. Don Preston, SensorMetrics Inc, PO Box 1049, Lakeville, MA 02347; tel 508-946-4904.

Under the Hood V: *Solid State Devices*



In this, the fifth installment in the Under the Hood series, we take a practical look at the diodes, transistors, integrated circuits, and other solid-state devices that make modern, compact, feature-laden communications gear possible.

By Bryan Bergeron, NU1N
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In our quest for a practical understanding of what lies under the hood of our feature-laden, microprocessor-controlled trans-ceivers, we've examined the basic passive components—resistors, capacitors, and inductors—in terms of their form, function, and application to radio. Although passive components are essential to any modern communications circuitry, it is the evolution of solid-state technology that rapidly redefined both amateur and commercial radio.

Semiconductors have come a long way since the "cat's-whisker" galena crystal detectors in use over half a century ago. Affordable, ultra-compact, battery operated, lightweight, and rugged communications units would be impossible without semiconductor diodes, transistors, ICs, and other solid-state components. Solid-state components offer a number of advantages when compared to vacuum tubes, switches, relays, and other mechanical assemblies that they replace. For example, solid-state components are generally:

1. Smaller, allowing more compact circuit designs.
2. More efficient. Transistors, for example, do not require the added energy to power the heater elements essential to the operation of vacuum tubes;
3. Cooler running. With no filaments to heat, solid-state components run cooler, resulting in longer component life;
4. Safer to work with. Although there are exceptions, most solid-state components are operated at relatively low voltages, commonly no more than 24 V.
5. More reliable. Although all components have a finite life span, most solid-state components outlive their vacuum tube or mechanical equivalents.
6. Less expensive. Single-IC receiver systems can be purchased for less than the price of a generic vacuum tube.

It's often difficult to appreciate the wide variety of solid-state components and assemblies available today. Perhaps more confusing, a given component may be available in a variety of packages. For example, a given transistor design may be leaded or surface-mount, bolt-on or solder-mount, encapsulated in an epoxy or metallic housing. Identification of small surface-mount components is especially difficult. With a little patience, a digital multimeter, and a good magnifying glass, however, most of the simpler surface-mount components can be identified.

Although we can't hope to cover the universe of solid-state components here, we can take a broad look at the most prominent stars, namely diodes, transistors, and ICs. Below is a brief introduction to some basic solid-state concepts, followed by an overview of diodes, transistors, and ICs.

Solid-State Concepts

Whatever the external appearance or intended function of solid-state components, they all share a common characteristic—their operation depends on some semiconductor material. Semiconductors are so named because they have a conductivity somewhere between that of insulators and conductors. Semiconductor conductivity and operating characteristics are purposely modified by adding substances or "impurities," such as phosphorous or boron, through a process known as doping, to otherwise pure semiconductor materials. Doping affects the relative abundance or absence of electrons or holes in a semiconductor, which in turn defines the electrical properties of the material.

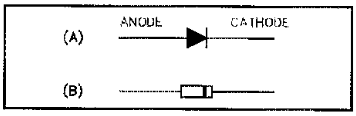

Figure 1—The simplest semiconductor device is the diode. At A, the schematic symbol for a diode; at B, the drawing of a typical diode package shows the band used to indicate the *cathode* terminal.

Some Useful Generalizations

Most solid-state components are based on the same semiconductor materials, allowing a few useful generalizations to be made as to their operating characteristics. Compared to their vacuum tube and mechanical equivalents, solid-state components are relatively fragile in terms of withstanding even brief periods of high voltages and temperatures, but relatively resistant to shock, vibration, and humidity. Whereas a vacuum tube or mechanical relay might recover completely from a voltage spike, an unprotected transistor or IC would likely be destroyed when exposed to a simple electrostatic discharge (which may be induced by careless handling!). Thus, although an IC might survive the mechanical stress of being dropped onto your shop floor, it might be destroyed by the electrostatic discharge generated as you pick it up! High temperatures and solid-state devices are simply not compatible. For this reason, high-power devices are normally attached to the chassis or other large heat-sink.

Solid-State Components

Diodes

Semiconductor diodes, the simplest and one of the most common solid-state devices, are used as signal and power switches, power supply gates, voltage references, capacitors, clippers and clamps, frequency multipliers, and as rectifiers in ac-to-dc power supplies (see Figure 1).

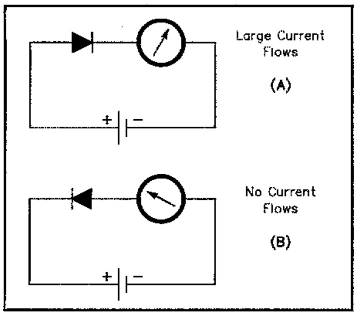

Figure 2—The basic characteristic of a diode is that it allows current flow in only one direction. At A, a positive voltage is applied to the anode, and a negative voltage to the cathode. The meter indicates a large current flow. At B, the diode is reversed, so the voltage on the anode is negative and the voltage on the cathode is positive. While a slight *leakage current* flows, it is too small to be seen with the meter.

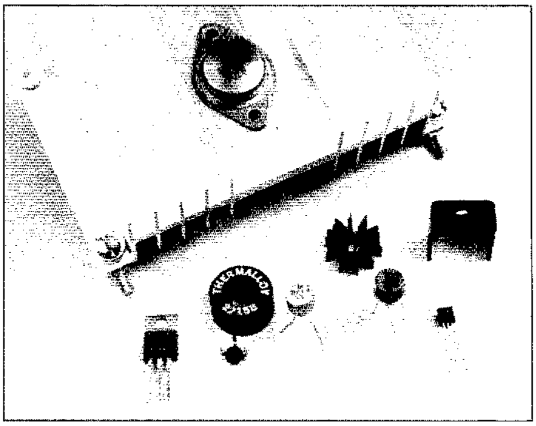

Figure 3—Transistors are built into a variety of metal and plastic packages. The actual transistor junction will be destroyed by excessive heat. Heat sinks are sometimes used to conduct away the excess heat. Heat sink size is determined by how much heat is generated by the transistor.

Diodes are used in electronics because they offer a greater resistance to the flow of current in one direction than in the opposite direction (this feature also makes diode troubleshooting relatively easy with the aid of a sensitive ohmmeter). See Figure 2.

Diodes are commonly classified by power and voltage handling capabilities, reverse leakage current and other factors, depending on the intended application. For example, a relevant rating for silicon power diodes to be used in power supply circuits is the *peak inverse voltage*—the maximum reverse voltage that can be applied before the avalanche point is reached; as well as the diode's current-handling capabilities.

Diodes can be packaged in a variety of materials, including epoxy, metal and glass. Silicon junction diodes have cornered the power diode market, mainly because of their wide temperature and voltage operating range, high power ratings, and small size.

Germanium diodes have a greater reverse leakage current than silicon diodes, but smaller forward voltage drop. If you're replacing a defective diode, use a diode with the same basic construction and rating. Germanium signal diodes don't generally fare well as replacements for high-power silicon junction diodes, and vice versa.

High-speed switching diodes have a typical reverse recovery time (the time required for the diode to return to a normal state after being switched from a reversed bias condition) of about 2 nanoseconds, compared to about 30 microseconds for a general purpose silicon power diode. Schottky-barrier or hot-carrier diodes have a low forward voltage drop, good frequency response, and low noise, and make ideal detectors. Zener diodes provide relatively stable reference voltages. Voltage-variable capacitors or Varicaps, are used in variable-frequency oscillator circuits. PIN diodes (made by defusing *P* and *N* material onto an almost *I*ntrinsically pure silicon layer) are used in RF switching, phase-shifter, and attenuator applications. Impatt, Trapatt, and Gunn diodes have applications at microwave frequencies, where they are used to generate RF directly from dc.

Perhaps the most intriguing class of diodes is light-emitting diodes or LEDs, which produce light when the P-N junction is forward biased. Compared to conventional filament lighting, LEDs are less expensive, have a greater life span, are more efficient, cooler in operation, more rugged, water resistant, and flameproof.

Photodiodes, in some respects the inverse of LEDs, convert photons or light into electric current.

Transistors

Stepping up in complexity from the 2-terminal diode is the transistor, a 3-terminal device that provides current amplification (Figure 3). Transistors have applications ranging from AF and RF amplifiers and oscillators to frequency converters, mixers, and high-speed switches. Like diodes, transistors are generally classified in terms of their internal construction, which in turn greatly influences their electrical characteristics.

The oldest, and perhaps best-known transistor designs are the bipolar NPN and PNP transistors, with their familiar emitter, base, and collector leads (Figure 4). The bipolar transistor and its successors have been largely responsible for the slow demise of the vacuum tube, at least for small-signal applications.

New designs and manufacturing techniques have produced transistors with less noise, greater power-handling capability, higher voltage ratings, higher gain, and other enhanced operating characteristics. A notable departure from the bipolar design is the field-effect transistor or FET. Unlike bipolar transistors, which operate on current flow, FETs work like vacuum tubes, in that the current flow in the output elements (from the source to the drain) is controlled by the voltage applied to one or more control elements (the gates). Like vacuum tubes, FETs provide both high input impedance, good dynamic range, and good noise characteristics. FETs are commonly used in preamplifier circuits at both audio and radio frequencies.

The two basic types of FETs differ mainly in whether the controlling gate electrode(s) is insulated from the source and drain elements. The gate in the junction FET or JFET is not insulated, whereas the gate in a metal oxide semiconductor FET or MOSFET has a layer of metal oxide between the gate(s) and the source and drain elements. Power FETs, known as vertical FETs, MOSPOWER FETs, and VMOS FETs, are available for RF amplifier applications, with moderate power output ratings through VHF.

To the average amateur, the most important transistor electrical characteristics are power handling capability (small signal versus power), operating frequency range (AF versus RF), and current amplification factor. Power transistors differ from small-signal designs mainly in internal construction and packaging. Most power designs are intended to be bolted onto a chassis or other heat sink, while most small-signal types are meant to be mounted in air. Transistors intended for small-signal applications typically generate less noise than those intended for power applications.

Most transistors will work at audio frequencies, but all transistors have some cutoff frequency where they can no longer provide amplification. Modern transistors are available with cutoff frequencies well into the UHF range and above.

Troubleshooting a suspected transistor is a bit more involved than checking a junction diode with an ohmmeter. Although an ohmmeter can verify absence or presence of gross opens and shorts, it's difficult to assess other parameters without more advanced test equipment. Fortunately, many

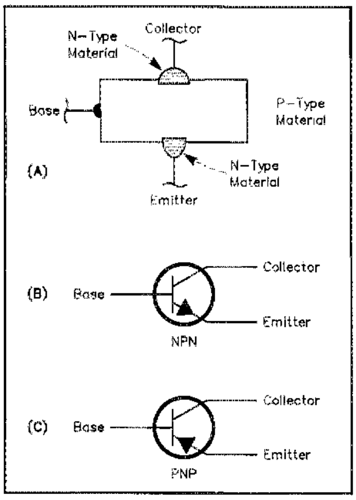

Figure 4—A junction transistor is made by *doping* N-type material onto P-type material (A). At B, the schematic diagram for an NPN transistor; at C, the schematic diagram for a PNP transistor.

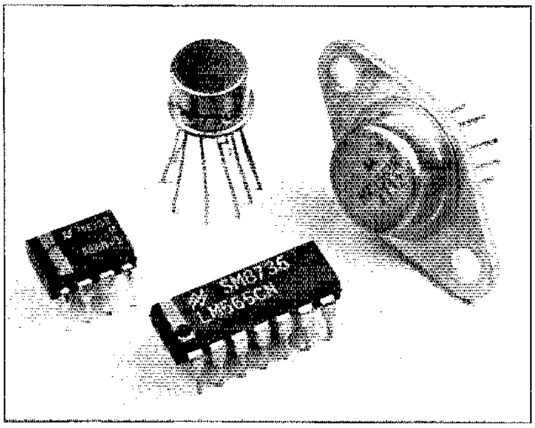

Figure 5—Like transistors, integrated circuits are built into a variety of packages, some of which are shown here. The two "bugs" in the foreground are *dual-inline package* or *DIP* ICs. Most digital ICs are supplied in some form of DIP, while analog ICs may be commonly found in any of the packages shown.

modern hand-held digital multimeters offer transistor-gain-test function.

If you look through any good electronics parts catalog, you'll find transistors listed in terms of their construction, suggested applications, operating current and voltage, minimum gain (hfe), and package type. For example, a typical listing might read: "NPN Power AF, 120 V, 250 mA, hfe 40, TO-39" for an NPN transistor designed for audio-frequency power applications, with maximum ratings of 120 V and 250 mA, a minimum hfe of 40, and in a TO-39 package (see the *ARRL Handbook* for examples of transistor packaging). If you're replacing a defective junction transistor, you'll be mainly concerned with the NPN or PNP classification, operating voltage and current, hfe, and package type, typically in that order. You may be able to use a transistor with a somewhat lower hfe or different package type (within physical limits, of course), but NPN and PNP transistors can't be interchanged. Similarly, a bipolar transistor can't be directly substituted for a MOSFET.

Integrated Circuits

Compared to the relatively limited world of diodes and transistors, ICs can be thought of as a catch-all for all solid-state components (see Figure 5). ICs often contain hundreds or thousands of diodes, transistors, and passive components in intricate circuits designed to perform functions ranging from simple digital logic switching, phase-locked-loop frequency control and direct digital synthesis to microprocessor control. ICs are relatively inexpensive, plentiful, and extremely functional, given their size. By reducing component count, both engineering time and equipment production cost are often reduced. In addition, because of the manufacturing process, it's generally easier to obtain better transistor matching within ICs than with discrete components.

Despite the extremely wide range of possible functions, ICs are generally classified as either digital, analog, or mixed. Digital ICs deal with input and output signals that are either on or off, as in the digital logic circuitry used in your PC. In contrast, analog or linear ICs work with time-varying signals, such as audio and some forms of video. Mixed ICs, a hot area in IC design circles, combine both digital and analog functions on a single chip. Analog-to-digital and digital-to-analog converter ICs are often implemented in a single, mixed-IC design.

Perhaps the most notable contribution of solid-state components to communications has been inexpensive digital ICs, especially microprocessors. Plug-in boards populated with digital ICs can transform any PC into a multi-mode communications package capable of handling everything from packet and SSTV to satellite tracking. Similarly, many contemporary transceivers rely on digital circuitry to manage RF switching, frequency synthesis, and most often used frequencies.

Families of ICs have similar characteristics, as defined by their common construction.[1] Examples of digital IC families include: TTL (transistor-transistor logic) and CMOS (complementary metal oxide semiconductor). Digital logic families vary considerably in input/output potentials, speed, and power consumption. For example, CMOS designs are used where power consumption must be minimized. More importantly, most families cannot be connected directly to other families because of differences in input and output levels. The one exception is the CMOS family, which can interface directly with most other digital IC families. In addition to the differences between digital IC families, there is also considerable variation in the characteristics of ICs within a family. Despite variations in characteristics though, all TTL ICs adhere to the same input and output level standards.

Unlike digital ICs, analog or linear ICs are designed to operate on a variety of signal input and output levels. Examples of analog ICs include the popular *operational amplifiers*, which are high-gain, dc-coupled amplifiers. Analog ICs are often designed to replace a number of discrete components, from those required to create an entire receiver, multimeter, or RF source, to video amplifiers and voltage regulators.

ICs are usually identified by their unique packaging. Compared to other devices, however, IC packaging provides little insight into the function of an IC. It's almost impossible to determine the nature of a particular IC without referring to the alphanumeric labels on the component package.

Unfortunately for most amateurs, the move toward increased use of ICs has made troubleshooting all but the simplest analog or digital ICs a formidable task. If you're lucky enough to be working with plug-in ICs, then try substitution—just like in the old vacuum-tube days. In many cases, repair requires shipping the entire transceiver back to the manufacturer. Even so, most of us would argue that the added functionality is well worth the troubleshooting hassles—especially given the functionality we have come to expect from our equipment. It would simply be impossible, were it not for modern solid-state devices.

For Additional Reading

R. Schetgen, ed, *The ARRL Handbook for Radio Amateurs* (Newington: ARRL, 1994), Ch. 4 "Solid-State Basics."

L. Wolfgang, *Understanding Basic Electronics* (Newington: ARRL, 1992).

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[1] G. Tharalson, "Which Logic Family Is Best For You?" *Electronic Products*, 31(16): 53-7, 1989.

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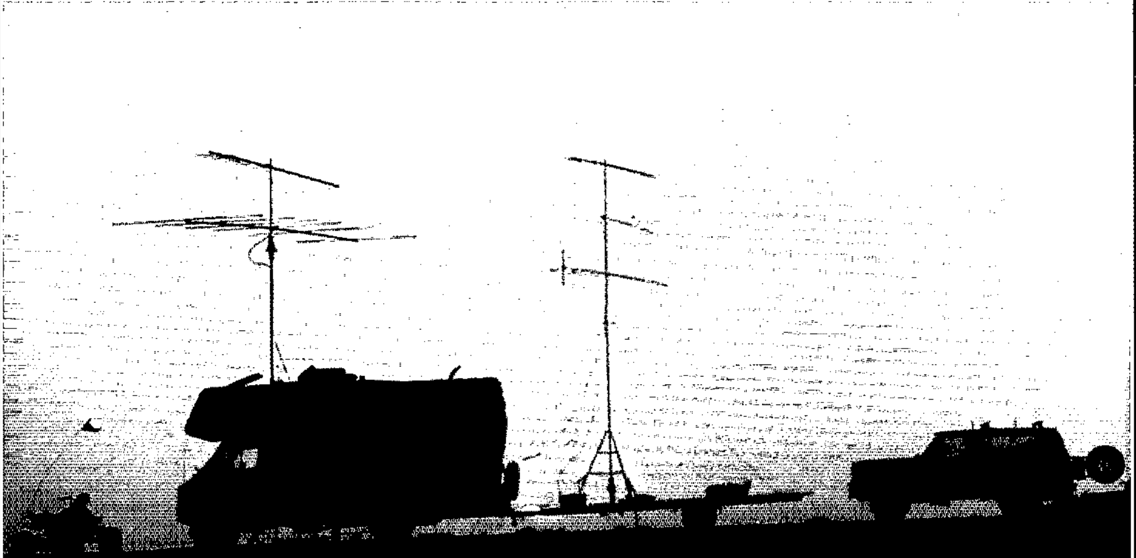

Outdoors and Amateur Radio!

Field Day and the June VHF QSO Party

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OUR COVER

From 160 meters through microwaves, the bands will be full of signals this month. US amateurs will set up simulated emergency stations in the wilds and before the eyes of the public for Field Day 1994 on the weekend of June 25 and 26. And hams will pour RF above 30 MHz into the skies for the June VHF QSO Party, June 4 and 5. (See the complete rules for both events in May *QST*.)
—*Courtesy of WB6DTA and W8DF*

CONTENTS

June 1994
Volume LXXVIII Number 6

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

Key Components Of Modern Receiver Design—*Part 2*

First-IF AGC, a high-dynamic-range mixer, a low-noise VCO—here are circuits you may see "pretty soon now" in commercial gear. Or maybe you'll just build them yourself!

By Dr. Ulrich L. Rohde, KA2WEU
52 Hillcrest Dr
Upper Saddle River, NJ 07548

Part 1 of this article* conveyed dynamic range, AGC and synthesizer concepts important in high-performance MF/HF reception. This month, we get down to putting these ideas into practice.

Building These Concepts Into a Receiver

Achieving electronically controllable front-end attenuation is the first important step, and a conventional PIN-diode attenuator can provide it. The standard constant-impedance T PIN attenuator exhibits intermodulation that increases with attenuation from 0 to –6 dB and then diminishes. To maintain a third-order intercept point of +30 dBm at frequencies above 1.5 MHz, a slightly different biasing scheme must be provided. Figure 6 shows a schematic of a suitable PIN-diode attenuator. These PIN diodes are not immune to second-order IMD on large signals, and therefore a preselector should be used in front of this AGC stage.

In the IF, a differential amplifier with RF feedback as previously published in *QST*[6] should be used. Figure 7 shows this stage with an added overload-detection stage and first-IF AGC. The gain-controlled stage includes an AND function for local (AGC1) and global (AGC2) AGC. Figure 8 graphs the computer-simulated performance of a combination Norton feedback amplifier and differential amplifier as implemented by the BRF96-BFR96-BFT66 trio in Figure 7. It provides a noise figure in the vicinity of 2 dB and a high intercept point with very little reverse feedback. By contrast, the standard feedback amplifier often used in the post-mixer position suffers from unacceptably high feedback and is therefore frequently used with a small attenuator (commonly 6 dB) between its output and the crystal filter that follows it.

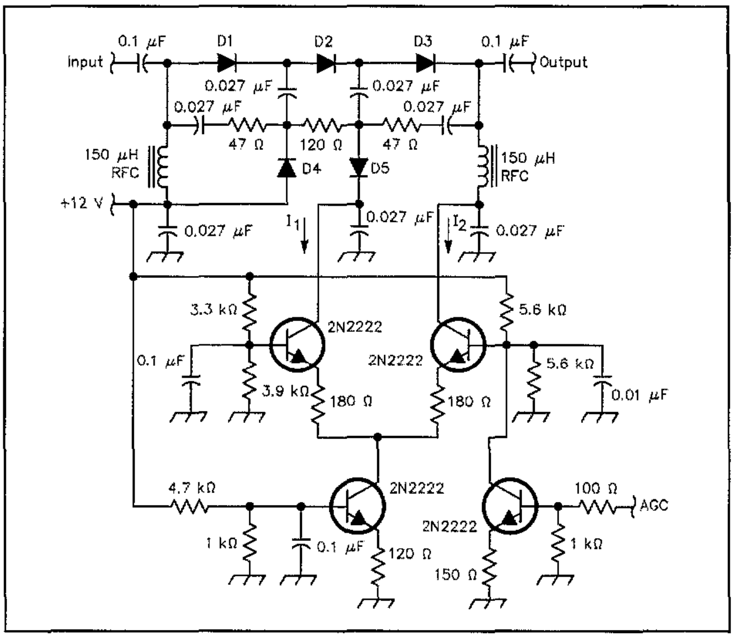

Figure 6—PIN-diode attenuator with dc amplifier for use in a high-performance communications receiver. D1 through D5 are Hewlett-Packard HP8052-3081 PIN diodes. The circuit's attenuation increases with increasing positive voltage at AGC.

The dual-AGC circuit acts as follows: The overload-control loop (AGC1) is normally inactive, and the differential amplifier acts as part of the regular (AGC2) AGC loop. When a strong signal falls into the window outside the second-IF filter and inside the roofing filter, the AGC2 loop does not respond to it. Instead, the AGC1 loop applies a control voltage to the local AGC1-controlled stage (Q1 in Figure 7) and reduces its gain to prevent second-mixer overload.

The AGC1 loop protects the second mixer, but what about the input of the AGC1-controlled amplifier itself? We can keep this circuit from being overdriven by activating the system's continuously adjustable front-end attenuator, which is included in the loop.

This three-loop AGC control system is currently the best possible way to deal with the real-world effects of finite intercept point in the various components in the chain. This approach is more costly than simpler systems, however, and thus is infrequently incorporated even in high-end military and commercial gear.

A key issue in the multiple-AGC-loop approach is the *selection of the right AGC* distribution and time constants. The best paper I've seen on this subject was pub-

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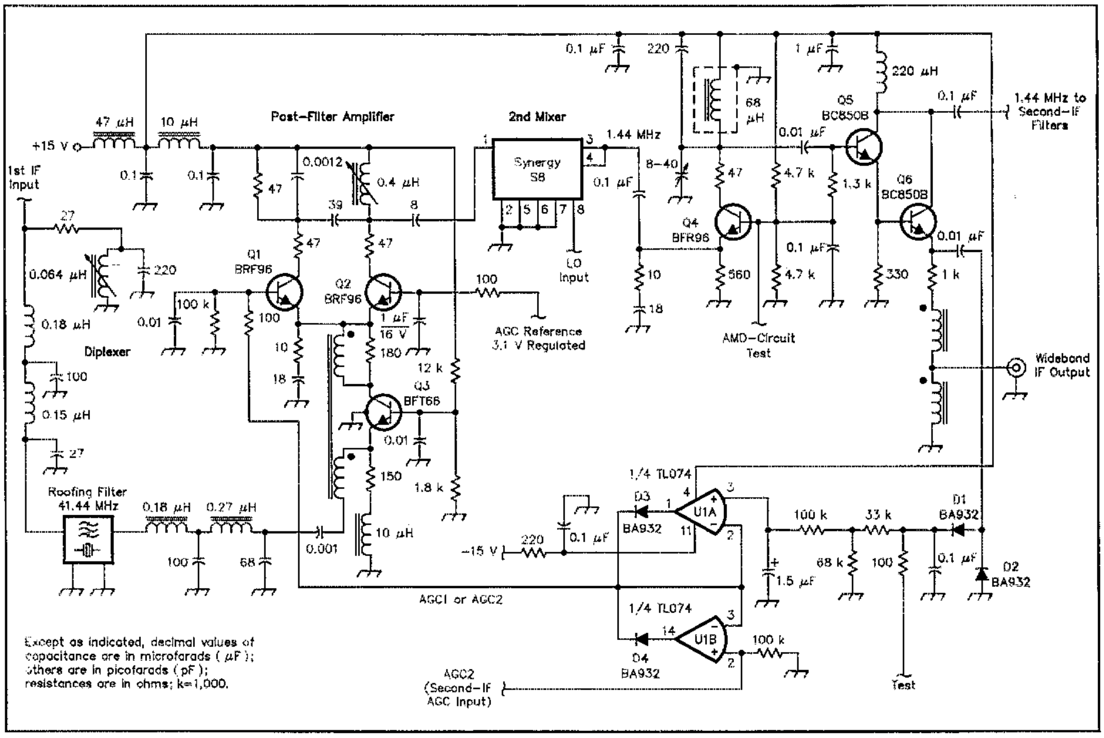
*Part 1 of this article appeared on pages 29-33 of May 1994 *QST*. Note: The microvolt equivalent shown for S4 in Table 1 of Part 1 should read 1.5 instead 1.3.

[6]Notes appear on page 31.



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

Figure 7—AGC amplifier and overload monitor schematic for a high-performance shortwave receiver. Q1 to Q3 serve as a differential, gain-controlled post-roofing-filter amplifier. Q4 acts as a buffer amplifier, Q5 drives the receiver's second IF stages (1.44 MHz), and Q6 drives the AGC1 voltage-doubler detector (D1 and D2) and a wideband IF output. Q1 receives control voltage via U1A (AGC1, the control voltage generated by detector/voltage doubler D1 and D2) and U1B (AGC2, the control signal generated by the receiver's second-IF AGC system).

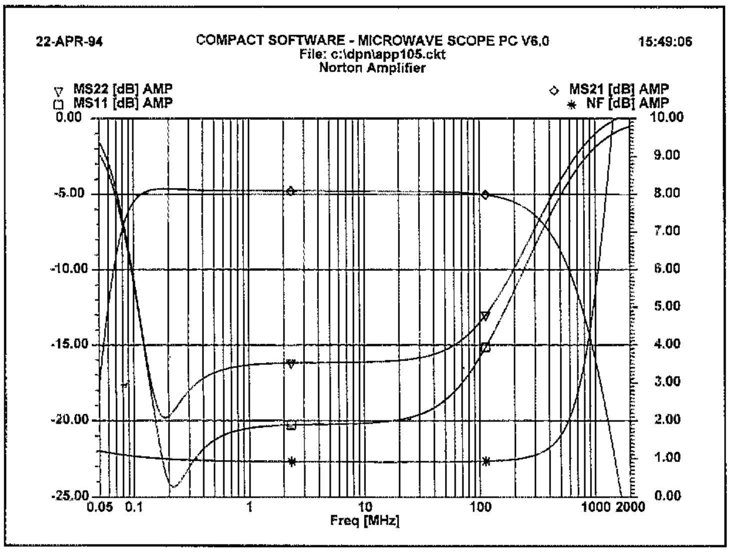


Figure 8—Performance of a combination differential amplifier and Norton noiseless-feedback amplifier as simulated by Compact Software's *SCOPE 6.0.*

lished by J. Porter,[7] but the sources cited in the references (see Part 1) should also be investigated.

Where to Put the Roofing Filter

Selectivity should be as high as possible as close as possible to the antenna, of course, but doing so can involve compromise. The system of which Figure 7 is part positions the first-IF crystal filter directly after the first mixer (see the block diagram in Figure 4). This is a compromise that causes two problems. One is that a crystal filter's input and output impedances are usually profoundly reactive within and adjacent to its passband, and mixers react violently to reactive terminations. The second problem is that filters are lossy, and this loss directly degrades the system's noise figure if an intermediate amplifier is not present to overcome it.

Traditional monolithic crystal filters exhibiting reasonable bandwidths of 15 kHz or less have input and output impedances from 600 to 1500 Ω. By the time two or three such filters are cascaded to obtain a reasonable shape factor, their combined loss can

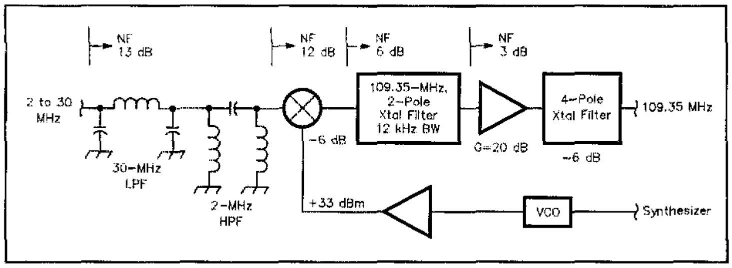


Figure 9—A high-performance receiver front end as outlined by Sabin (see Note 8). This approach uses two crystal filters to achieve an acceptable shape factor.

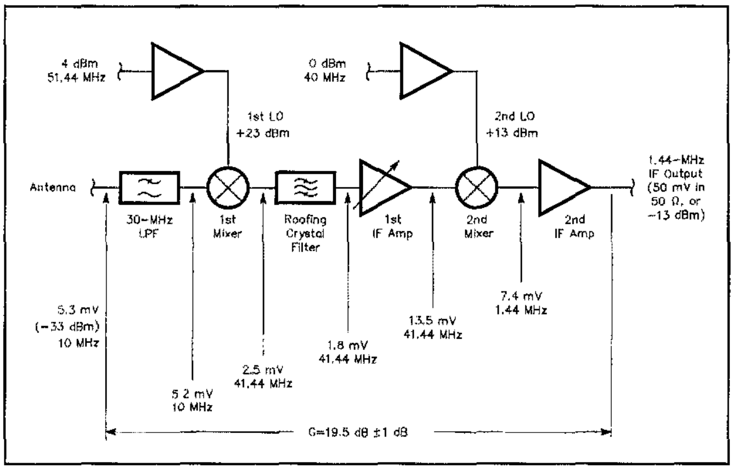


Figure 10—Level block diagram of the input stages of a modern, high-performance shortwave receiver. Note that the first IF amplifier in this system receives AGC, while the first IF amplifier in Sabin's paper (see Note 8) operates at constant gain. A practical system requires the addition of an RF preselector, which can be a set of sub-octave filters, a motor-driven multistage circuit or a binary-coded input filter.

reach 6 to 8 dB. Adding the mixer loss (6 dB), the roofing-filter loss (8 dB) and the 4 or 5-dB noise figure of the traditional first-IF input stage results in an unacceptably high noise figure of 18 to 20 dB! The only recourse for a designer is to insert a post-mixer preamplifier, which violates the concept of keeping high selectivity as close as possible to the antenna.

To avoid the necessity for a post-mixer amplifier, one must invest somewhat more money in a discrete-crystal roofing filter with an insertion loss under 5 dB (typically 3.5 dB) and follow it with a post-filter amplifier with a noise figure of 2 dB or less. This approach can provide a total NF of less than 12 dB—equivalent to 0.3 µV for 10 dB (S+N)/N with a bandwidth of 2.4 kHz. I implemented this post-mixer filter arrangement in the 1973-vintage SE 6861/11 ManPack by AEG Telefunken. Sabin independently published an excellent paper on the concept in 1981.[8]

Figure 9 shows a high-performance receiver front end as outlined by Sabin. This approach resorts to two crystal filters, a two-pole and a four-pole, because its four-pole filter alone would be insufficiently selective. The amplifier between the filters is necessary because the four-pole filter has an insertion loss of 6 dB. Figure 10 shows a more modern approach that combines the techniques previously outlined. It uses one crystal filter that achieves the desired shape factor with less than 4 dB of insertion loss.

How About Mixers?

The area of mixers, specifically switching mixers, has already received ample coverage. Figure 11 shows the result of a two-tone test of a mixer with 5-dBm tones applied to the mixer input. Figure 12 shows the results of a three-tone test using the same tone levels. To protect the spectrum ana-

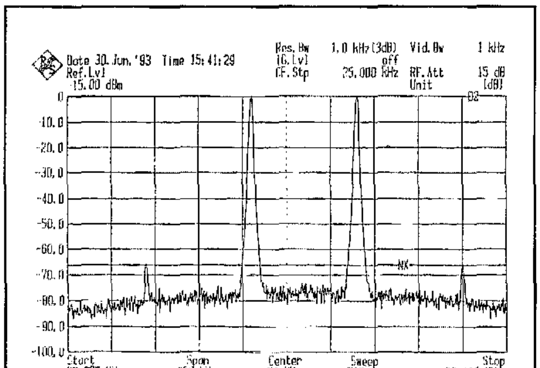


Figure 11—Result of a two-tone mixer test with 5-dBm tones at the mixer input. The two smaller pips are third-order IMD products.

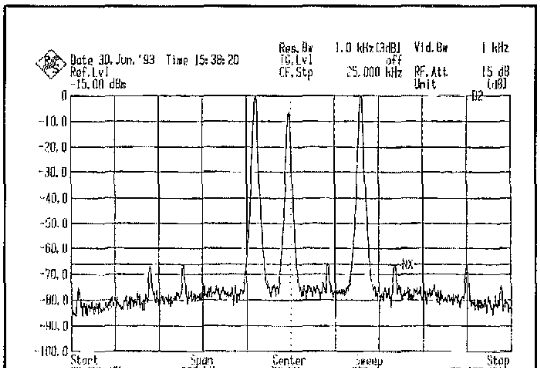


Figure 12—Result of a three-tone mixer test with 5-dBm tones at the mixer input. Additional third-order IMD products have appeared, along with the two-tone products shown in Figure 11.

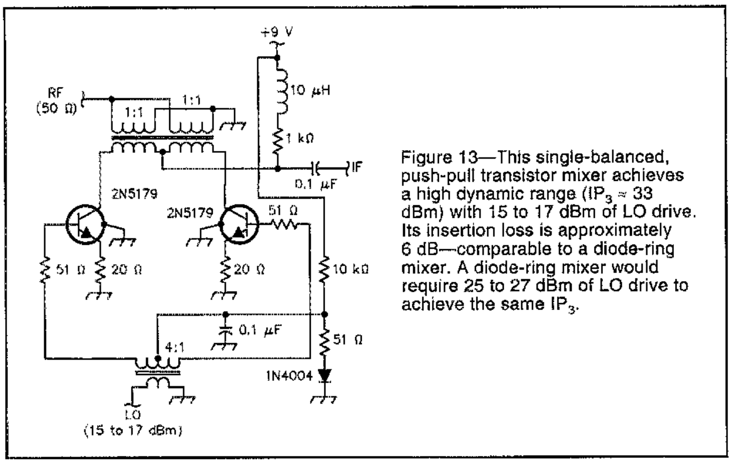


Figure 13—This single-balanced, push-pull transistor mixer achieves a high dynamic range ($IP_3 \approx 33$ dBm) with 15 to 17 dBm of LO drive. Its insertion loss is approximately 6 dB—comparable to a diode-ring mixer. A diode-ring mixer would require 25 to 27 dBm of LO drive to achieve the same IP_3.

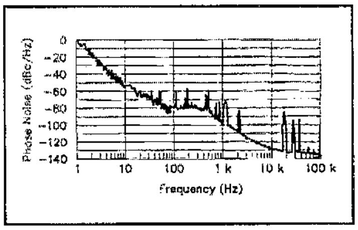


Figure 15—Measured phase noise for a practical system based on the DDS-driven-PLL technique of Figure 14. The test equipment used for this measurement and that of Figure 18 was a Hewlett-Packard HP3048A.

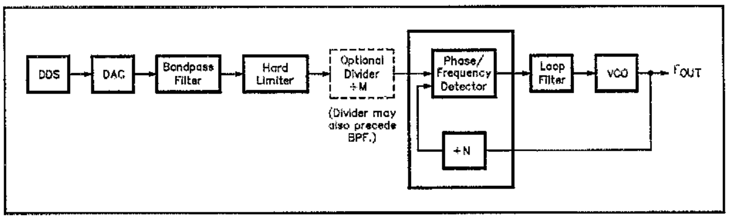


Figure 14—Block diagram of a DDS-driven PLL. This approach simultaneously achieves fine steps and fast settling time by using a direct digital synthesizer as a variable reference for a conventional PLL.

lyzer, an external pad was added. Notice that the IMD product levels are pretty much the same in both cases, but that the number of spurious products significantly increases in the three-tone test.

Three-tone testing is a worthwhile extension of the two-tone testing with which radio amateurs are already familiar.[9] In on-the-air operation, the appearance of another strong signal between two strong carriers can be particularly annoying because of the significantly higher number of spurs generated. Three-tone testing can shed more light on a VHF or UHF radio's tendency to generate spurs in response to the many high-level VHF and UHF signals that exist in major metropolitan areas. Three-tone testing is also important because it can indirectly show a mixer's tendency toward cross-modulation. The transfer of modulation from a strong modulated carrier to weak unmodulated carriers is still observed on ham bands adjacent to or containing strong broadcast signals.

Recent mixer studies have lead to the development of a medium-frequency mixer that is well-suited to shortwave applications and home-built projects. Shown in Figure 13, it consists of two transistors in a push-pull configuration that is singly rather than doubly balanced. Because of the degenerative feedback introduced by the 20-Ω emitter resistors, the two transistors do not have to be matched. This mixer's real advantage lies in its achievement of a 33-dBm output intercept with only 17 dBm of LO drive. Typically, a diode ring with the same intermod performance requires 25 to 27 dBm of LO drive.

Tests indicate that the upper frequency limit of this mixer lies in the 500-MHz region. The circuit's lower frequency limit depends on the transformer inductances and the ferrites used for the transformer cores.

Finally, the Synthesizer!

Synthesizers have been around for a long time, in phase-locked loop (PLL) and other forms.[10] Recent synthesizers have taken advantage of direct digital synthesis (DDS). DDS devices are available with 32-bit resolution, but ham equipment using DDS has so far tended to use less-expensive 16-bit parts. Such equipment has so far used DDS in a number of mix and divide schemes, but none have used the technique in what should be called a *DDS-driven PLL.*

Figure 14 block diagrams such a system: a PLL frequency synthesizer with a reference frequency generated by a DDS. The advantages of this approach are:

- Broad output span
- Extremely fine frequency resolution with constant steps
- Fast switching time
- Fast settling time
- Low phase noise
- Low spurious noise (the design uses no mixers)
- Relatively simple circuitry

A Qualcomm application note[11] provides a detailed analysis of this type of synthesizer. Using this approach, I and others have constructed a microprocessor-controlled frequency synthesizer that covers 75 to 105 MHz based on 32-bit resolution. Its output steps are essentially 0.07 Hz in size. A microprocessor drives the DDS and the PLL.

The system uses a Harris HSP45102 DDS and a Phillips TDA8702 digital-to-analog converter to generate a variable reference. Its output, centered around 10.7 MHz, is fed to a Motorola MC145170 single-loop PLL after being cleaned up by an inexpensive 10.7-MHz crystal filter.

The system's lock-up time depends solely on the time constant of the loop filter associated with the Motorola chip, and this approximation synthesizer gives a maximum error of 0.1 Hz relative to the output. The reference frequency chosen is 100 kHz, and the loop bandwidth is adjustable from 100 Hz to 1 kHz by selecting the filter components.

Figure 15 shows the SSB phase noise of the total system, which also uses a novel VCO (Figure 16). Figure 17 shows the predicted phase noise of this oscillator alone. Figure 18 shows the measured phase noise of the Kenwood TS-50's multi-loop synthesizer, a sound design, as a comparison. Comparing Figures 15 and 18 reveals the surprising fact that the simpler, wide-range single-VCO system is less phase-noisy than the TS-50. This is surprising because synthesized MF/HF equipment has long used *multiple* VCOs to cover a 30-MHz span solely because this approach allows

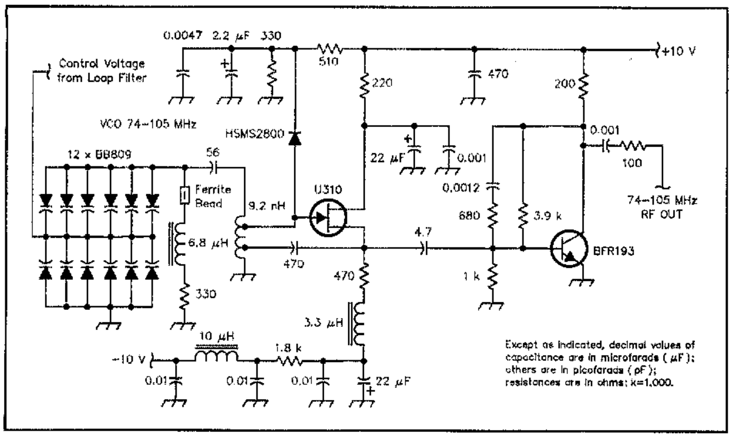


Figure 16—Circuit of the wideband VCO used in the system measured in Figure 15. Multiple tuning diodes are used to reduce the circuit's phase noise. The HSMS2800 diode at the U310's gate is reverse-biased and does not clip positive signal peaks like the gate-to-ground clamping diodes sometimes used in JFET oscillators by amateur builders. See text.

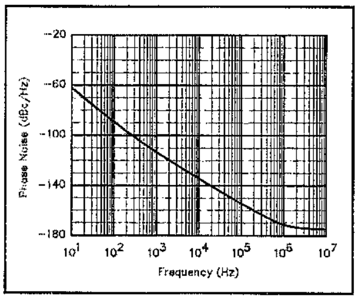


Figure 17—Predicted phase noise of the wide-band VCO alone.

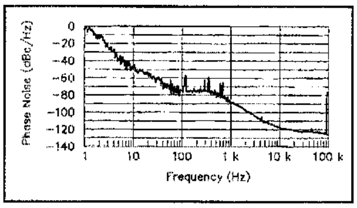


Figure 18—Measured phase noise of the TS-50 multi-loop synthesizer.

VCO optimization for minimum phase noise. Achieving the same noise performance with a *single* VCO such as that shown in Figure 16 should allow us to reduce equipment cost, complexity and size without compromising phase-noise performance.

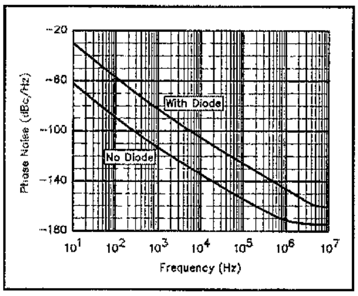


Figure 19—Predicted phase noise of the wide-band VCO with and without the gate-clamping diode long used by amateur builders in JFET VFOs. Depending on the feedback level, a gate-clamping diode can seriously degrade an oscillator's phase-noise performance. Gate bias achieved by means of a source resistor (see Note 12) is a better means of preventing gate-source conduction in a JFET oscillator.

The gate diode in Figure 16 is reverse biased and does not clip the gate signal. In the past, some authors have recommended using clipping diodes to limit positive peaks at the gates of JFET oscillators.[12] If feedback is high in such an oscillator, the external diode can add significant noise to the oscillator output. Figure 19 shows that the noise increase due to a gate-clamping diode can be as high as 30 dB, but this effect is rarely mentioned in the literature. Note, however, that merely including a gate-clamping diode does not *guarantee* higher phase noise. Loose coupling between the JFET and tuned circuit may keep the gate voltage low enough so that the clamping diode does not conduct. A nonconducting gate clamper adds no noise to the oscillator output.

Notes

[6] U. Rohde, "Recent Advances in Modern Shortwave Receiver Design," *Part 1, QST*, Nov 1992, pp 45-55, Fig 11.

[7] J. Porter, "AGC Loop Control Design Using System Theory," *RF Design*, Jun 1980, p 27.

[8] W. Sabin, "Use of Mixers in HF Upconversion Receivers/Exciters," Wescon/81, Professional Program Session Record 24, pp 24/1 to 24/4, Electronic Show and Convention, Sep 15-17, 1981.

[9] Even though two of the most popular Amateur Radio emissions (CW and SSB) are forms of amplitude modulation, AM as a concept and a technique receives relatively little attention from many radio amateurs today mainly because hamdom has largely abandoned full-carrier, double-sideband AM radiotelephony in favor of SSB radiotelephony. Meanwhile, "just across the hall" in commercial practice, high-quality AM transmission and reception are very much alive and up-to-date in the form of television and LF, MF and HF sound broadcasting. Amplifiers and amplifier devices designed for high-quality linear amplification of AM and wideband multiplexed signals, especially those intended for TV, satellite and cable-distribution service, are routinely three-tone-tested to quantify their ability to simultaneously amplify multiple signals with minimal intermodulation and maximum linearity.—*Ed.*

[10] A complete survey of synthesizers for communications and testing equipment can be found in U. Rohde, *Digital PLL Frequency Synthesizers—Theory and Design* (Englewood Cliffs, NJ: Prentice-Hall Inc, 1983). This book is available from Compact Software, 201 McLean Blvd, Paterson, NJ 07504, tel 201-881-1200, fax 201-881-8361.

[11] "Hybrid PLL/DDS Frequency Synthesis," Qualcomm Inc application note CL80-3459-1, Jun 1990.

[12] Experimenters have used the gate-clamping diode in JFET oscillators because (1) it disallows gate-junction conduction and greatly reduces the frequency drift that can indirectly result and (2) it is more experimenter-friendly than using a source bias resistor to accomplish the same thing because wide FET-to-FET characteristics variation may make some copies of a given oscillator circuit fail to start if a particular source resistor value is specified. Two experiments at WJ1Z in converting existing gate-clamped JFET oscillators to source bias suggest, however, that a source resistance of 1 kΩ is a good initial value. (Hint: Good parts generally give better results, so also consider using a narrower-tolerance FET than the ubiquitous and mediocre MPF102 or 2N3819—a 2N4416, J310 or U310, for example.) The WJ1Z oscillators converted in this way (1) exhibited higher output than they did when gate-clamped and (2) *were just as frequency stable* with source bias as they were when gate-clamped.—*Ed.*

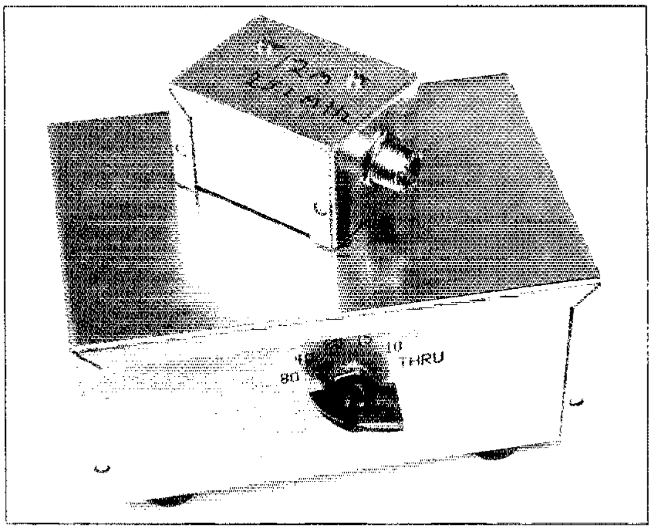
Coming in Part 3:

Testing and improving the strong-signal performance of ham transceivers, and a peek at one of the latest DSP-based commercial radios.

QST

Inexpensive Interference Filters

Here's an inexpensive way to improve the operating conditions at your station. You'll enjoy the simple construction project and its benefits—at home or out on Field Day.



By Alan Bloom, N1AL
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Santa Rosa, CA 95409
alanb@sr.hp.com
Photos by Kirk Kleinschmidt, NT0Z

For most Amateur Radio clubs, the premier operating activity of the year is Field Day. It draws together operators of all levels, from the newest Novice to the most experienced contester, to operate nonstop for 24 hours. But, along with the excitement, fresh air and camaraderie comes an unwelcome intruder: interference.

Anyone who has ever operated a club Field Day station understands the problem: Trying to operate multiple HF transmitters and receivers in close proximity almost guarantees interference. Synthesized radios often make the problem even worse because of synthesizer phase noise.

One way that phase noise shows up is as broadband noise emitted by transmitters. It's true that a modern transmitter's output low-pass filter reduces noise emissions on higher-frequency bands. Unfortunately, however, synthesized receivers also suffer from phase noise, which causes a broadband spurious response. The result is that interference can occur between a transmitter and receiver on any two bands.

The solution is to insert a band-pass filter, tuned to the proper frequency, between each transceiver and its antenna. That way, the filter can prevent both spurious noise emission from the transmitter and spurious response by the receiver. If each filter covers an entire band, no tuning is required.

Just as interference isn't limited to Field Day operations, neither is the use of these filters. One club member built copies of the filters I'll describe to cure some receiver problems he was experiencing from local AM broadcast stations. The filters also helped reduce the 40-meter crud level caused by out-of-band shortwave stations.

The SCRA Field Day Filters

Six filters were built as a club project by the Sonoma County (California) Radio Amateurs, Inc, W6LFJ. Each filter is band switched so it can be used on any of the five most popular HF ham bands. W6LFJ generally runs seven stations for Field Day: two CW, two phone, one Novice CW, one Novice phone, and one packet.

Because each filter section covers an entire ham band, it provides no protection for stations that share a band by simultaneously operating CW and phone. To solve

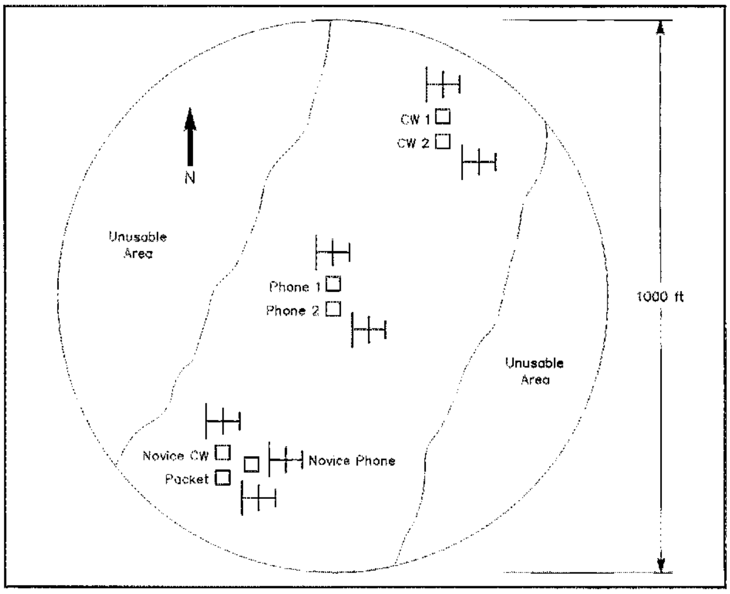
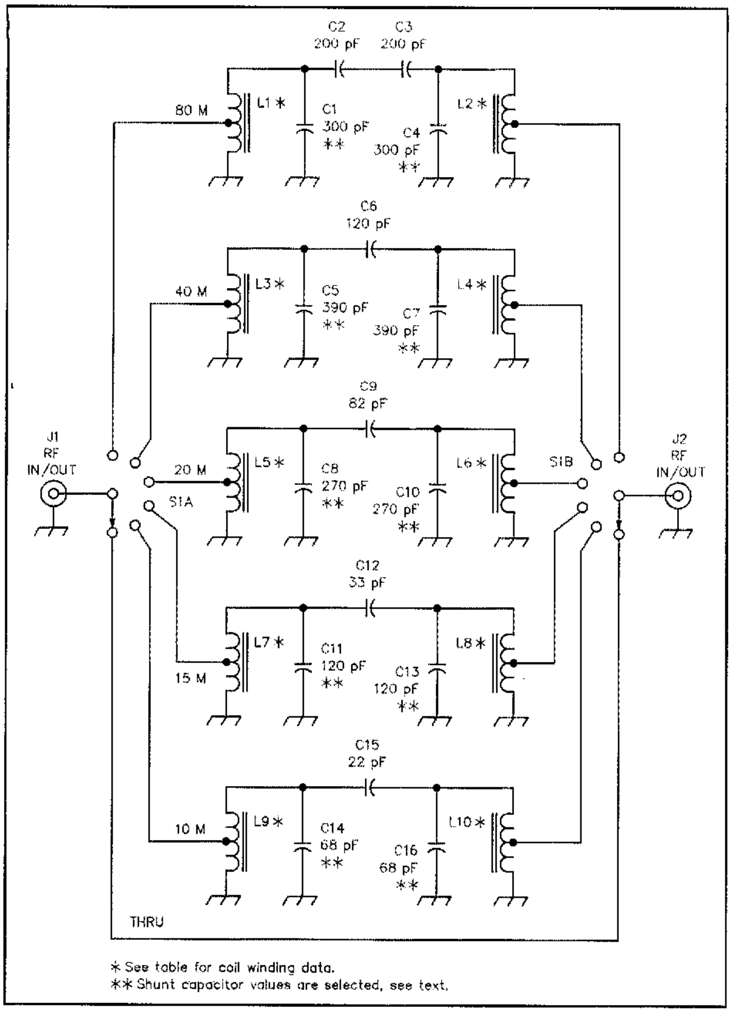


Figure 1—The filters don't reduce interference between stations on the same band. By locating CW, phone and Novice stations as far apart as possible (within the Field Day 1000-foot limit), interference is minimized.



* See table for coil winding data.
** Shunt capacitor values are selected, see text.

Figure 2—Schematic of the 5-band interference filter. Coil data is given in Table 1.

Table 1
Filter Data. All Coils are Wound with #16 Enameled Wire.

Band (meters)	*Core*	*Number of Turns/Strands*†	*Tap at Turn*‡	C_{shunt} *(pF)*	$C_{coupling}$ *(pF)*
160	T-94-6	26/1	13¼	1200	390
80	T-94-6	26/1	9¼	300	100*
40	T-94-10	13/2	7¼	390	120
30	T-94-10	13/2	6¼	200	56
20	T-94-10	8/3	5¼	270	82
17	T-94-10	8/3	4¾	150	47
15	T-94-10	8/3	4¼	120	33
12	T-94-10	8/3	4	82	27
10	T-94-10	8/3	3¾	68	22

*Two 200-pF capacitors in series.
†See the explanation in the text.
‡From ground end of inductor.

that problem, we use three operating positions (CW, phone and Novice/packet) situated as far apart as possible within the permissible 1000-foot circle (see Figure 1). The physical separation allows us to have three stations operating on the same band at the same time. The filters prevent the two stations at each position (operating on different bands) from interfering with each other.

We also constructed single-band filters for 160, 30, 17, 12 and 10 meters. Because the Novice phone station can operate only on 10 meters, there's no need for a band switch on that unit. The Novice CW filter could similarly be simplified to three bands: 80, 40 and 15 meters. A 10-meter filter isn't needed because the station would interfere with the nearby 10-meter Novice phone station anyway. The phone station can occasionally switch to CW if desired.

The Design

Refer to Figure 2. Each filter section uses a pair of coils wound on T-94 powdered-iron toroids connected as a pair of coupled tuned circuits. For low loss, silver-mica resonating and coupling capacitors are used. The calculated filter power rating is about 150 W, which was confirmed last Field Day when someone accidentally connected a 1-kW amplifier to one of the filters! (After replacing one blown capacitor, the unit worked fine again.) For high-power operation, connect the filter between the transceiver and the amplifier. With 150 W applied continuously, the coils get only slightly warm to the touch.

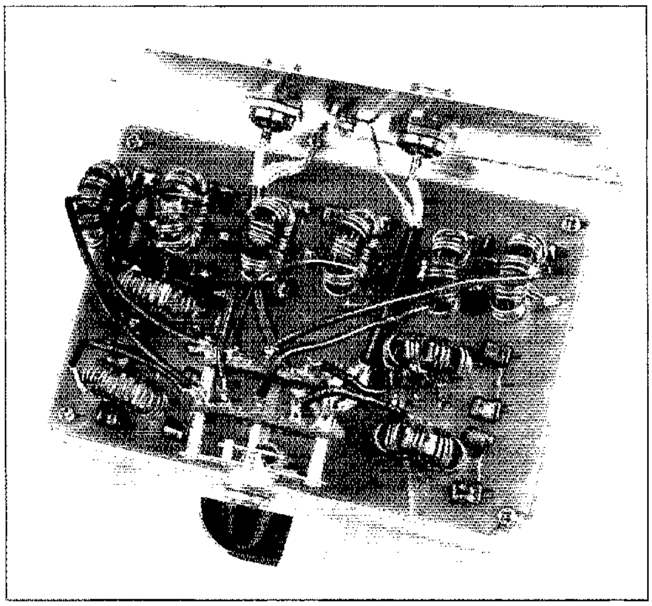
I experimented with various core materials and wire sizes to find the inductance value that provides optimum unloaded coil Q on each band. (Unloaded Q is the quality factor of the coil by itself, with no other components connected. Loaded Q is the net Q of the coil when it is loaded down by the effective parallel resistance of the circuit.) Once I found the optimum inductance value on each band, I calculated the 50-Ω tap point to provide the desired loaded Q and determined the proper coupling capacitor size.

Unloaded Q is very important because it determines the amount of power absorbed by the coil, which affects insertion loss and power-handling capability. The percentage of power absorbed in each coil is 100 × Q_L/Q_U, where Q_L and Q_U are the loaded and unloaded Q, respectively. To reduce loss, you can reduce Q_L by increasing the coupling factor, but that widens the bandwidth, which reduces the rejection of out-of band interference.

Most of the coils used have unloaded Qs of about 250 and loaded Qs of about 3. So the total loss of the two coils is $2 \times 100 \times 3 \div 250 = 2.4\%$, or about 0.1 dB insertion loss at the peak of the response.

Construction

The prototype filters were built on PC



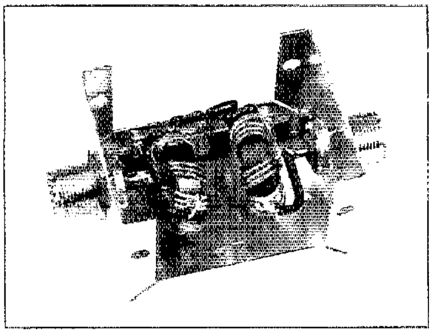


Figure 3—Construction details. If double-sided PC board is used, etch away the copper on the unused side to avoid the lossy capacitance that would otherwise exist between the ground plane and pads.

boards "etched" with an X-acto knife. The unit shown in Figure 3 is constructed on an etched PC board. (A PC-board template package and ready-made PC boards are available.[1]) The input and output connections of each section are flying leads soldered between the band switch and the tap points on the appropriate coils. We used a 2-pole, 6-position band switch to cover the five most popular bands plus a "through," or bypass, position. Additional bands can be added easily by using a band switch with more positions. Table 2 contains a parts list.

No PC boards are used in the single-band units. They are built in small boxes using terminal strips to mount the components.

All the coils are wound with the same size wire, preferably #16, although #18 will suffice (see Table 1). The higher-frequency coils use two or three wire strands in parallel to reduce resistive loss and improve Q. Winding the coils is by far the most time-consuming aspect of construction. Wind the wire on the core as tightly as possible and clean and tin the wire ends and the tap point before installation.

The leads between the coils and the band switch should be installed after the rest of the unit is assembled. Stray lead inductance is significant; in effect, it becomes part of the filter circuit. Keep the leads as short as possible consistent with keeping each pair of leads (going to each band section) the same length. You'll notice in Figure 3 that some of these pairs are intentionally routed with the two wires close to each other. We

Table 2

Parts List for the 5-Band Filter Unit.

Suppliers: Amidon Associates, Inc. PO Box 956, Torrance, CA 90508, tel 310-763-5770, fax 310-763-2250 (powdered-iron cores).

Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76062, tel 800-346-6873, 817-483-4422; fax 817-483-0931.

Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891, tel 800-866-6626, 401-596-3080, fax 401-596-3590. Ocean State also stocks most of the parts listed.

(All capacitors are 500-V units.)
C1, C4—300 pF (Mouser 5982-10-500V300).
C2, C3—200 pF (Mouser 5982-15-500V200).
C5, C7—390 pF (Mouser 5982-15-500V390).
C6, C11, C13—120 pF (Mouser 5982-15-500V120).
C8, C10—270 pF (Mouser 5982-15-500V270).
C9—82 pF (Mouser 5982-15-500V82).
C12—33 pF (Mouser 5982-15-500V33).
C14, C16—68 pF (Mouser 5982-15-500V68).
C15—22 pF (Mouser 5982-15-500V22).
J1, J2—SO-239 chassis-mount connector (Mouser 523-83-878).
L1, L2—Toroid core (Amidon T-94-6).
L3, L4, L5, L6, L7, L8, L9, L10—Toroid core (Amidon T-94-10).
S1—2-pole, 6-position switch (Centralab 1410 or Mouser 10WR027).

Misc: Enclosure for the 5-band filter—3×7×5 inches HWD, (Mouser 537-TF-782).
Enclosure for a single-band filter—1⅝×2¾×2⅛-inch HWD (Mouser 537-TF-770).
Knob—¼-inch shaft (Mouser 512-2300).
4 spacers—#6 × ¼ in. (Mouser 534-405).
5 screws—#6-32 × ½ in. (Mouser 572-01890).
4 washers—#6 lock washer (Mouser 572-00675).
5 nuts—#6-32 hex (Mouser 572-00486).
2 lugs—#6 solder lug (Mouser 534-904).
4 feet—stick-on type (Mouser 517-SJ-5012BK).

Other: approximately 20 feet of #16 enameled wire (Ocean State Electronics MW-16, ¼-lb. spool) or #18 wire (Mouser 501-MW18H-1LB, Ocean State MW-18, ¼-lb. spool) can be used with slightly increased insertion loss; about 10 inches of RG-58 coax; single-sided PC board, 6½×4¼ inches (see Note 1).

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[1] PC boards for the 5-band filter are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price, $12 plus $1.50 shipping.

A template package containing a PC-board pattern, a part overlay and panel labels for a 5-band filter is available from the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please include a business-size SASE and identify your request for SCRA/BLOOM INTERFERENCE FILTERS.

found that this introduces a notch in the stop-band response,which provides additional rejection on an adjacent band.

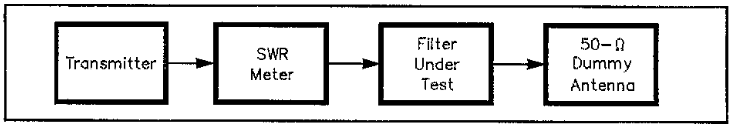


Figure 4—Test setup. With reduced transmitter power, adjust the resonator capacitors for lowest SWR on each band.

The coils' inductance won't always be exact because of variations in core permeability and winding technique. The capacitance values also vary within their allowed tolerance. For both reasons, you'll likely have to try a few capacitor values and select the ones that give best results. If you have a network analyzer, or spectrum analyzer with tracking generator, by all means use it. However, the filters can be tuned up with a low-power transmitter, dummy antenna and SWR meter (see Figure 4). Adjust the transmitter power for the lowest level that gives a full-scale SWR-meter deflection and select the parallel tuning capacitors (C1 and C4, C5 and C7, etc) for lowest SWR at the center of the desired band. Always maintain identical capacitance on each of the two resonators. Then, check the SWR at a number of points across the band. If the match is bad at the *high* end, *reduce* the capacitance; if the match is bad at the *low* end, *increase* the capacitance.

The capacitance values shown in

SCRA Filters Without Crosstalk

The filter-response curves in Figure 5 exhibit deep stopband nulls—nulls that happen to be quite useful because they tend to enhance a given filter's rejection of adjacent ham bands. These nulls may raise eyebrows among readers familiar with the response of "top-coupled" double-tuned circuits—that's what the SCRA filters are—because such filters shouldn't exhibit stopband nulls.

As Alan Bloom says, these nulls result from *crosstalk*—filter-to-filter coupling. And the nulls do occur: Figure 5 reflects measurements made by Alan and crew, and we also saw the nulls in ARRL Lab testing. As Alan implies, you'll need to closely duplicate the construction, mounting and wiring of the multiple filter unit shown if you want your version to exhibit similar nulls.

What if you *don't* build your SCRA filters the same way? What if, say, you build just one filter "point to point" in its own small box or, as some "RF heads" will do almost instinctively, you construct the filters to keep crosstalk to a minimum? Figure A, the result of computer modeling with Compact Software's *SCOPE 6.0*,* shows what you can expect if you build double-tuned-circuit (DTC) filters that actually perform like DTCs.

Don't let Figure A's broader, null-less curves disappoint you. Even if the SCRA filters you build act "only" like the clean, classic DTCs shown in Figure 2, you'll still be glad you brought them out on Field Day. They'll still significantly reduce interstation phase-noise problems, and they'll still greatly reduce the level of possible second-order IMD responses. (When you're operating at 7 MHz, for instance, second-order receiver IMD caused by your site's 21 and 14-MHz transmitters can make receiving difficult. Even if your 7-MHz SCRA filter has no passband nulls, it should reduce that IMD by at least *40 dB* because it attenuates intermodulating 14 and 21-MHz signals by at least 20 dB—and we aren't even considering the additional beneficial effect of antenna selectivity.)

So, however you build your Field Day filters, you'll probably be glad you did—as long as you get them going by FD, that is!—*David Newkirk, WJ1Z*

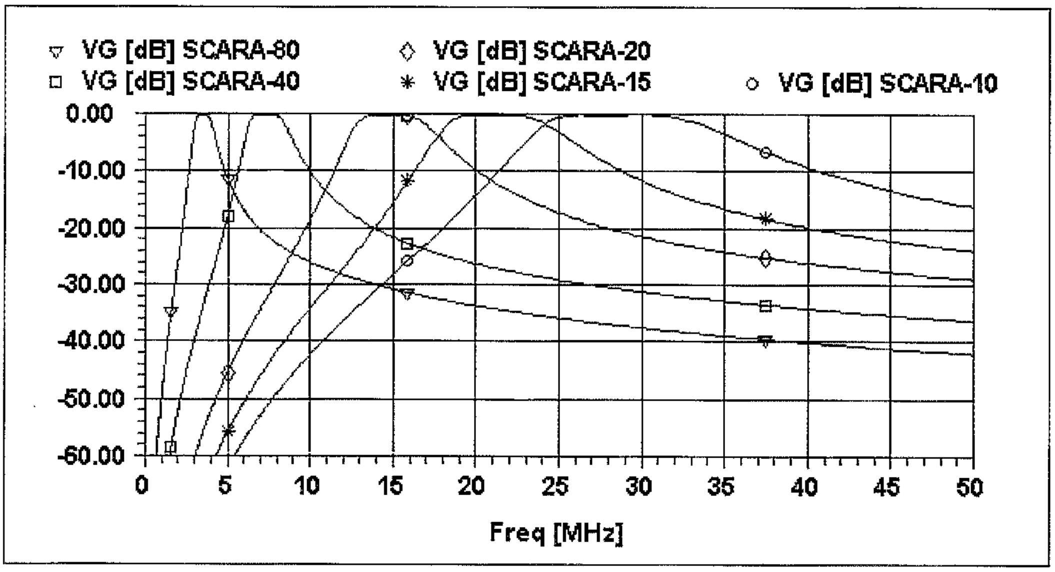


Figure A—If you build your SCRA Field Day filter(s) for no filter-to-filter crosstalk, they'll exhibit responses more like this composite view of two simulations done with Compact Software's *SCOPE 6.0.* The filter circuits shown in Figure 2 cannot produce the sharp stopband nulls of Figure 5 on their own; circuit "strays"—inductance and capacitance—produce the nulls. Even if your versions don't exhibit stopband nulls, they should still prove mighty useful on Field Day. See text.

Table 1 are nominal values that worked with the units we built. To allow for tuning, you may want to purchase capacitors of values one size smaller than indicated and add low-value capacitors in parallel as needed; extra holes have been provided in the PC-board layout for the purpose. There should be no need to trim the coupling capacitors (C2, C3, C6, etc), as those values are not critical.

Results

Figure 5 shows the frequency response plots of a typical unit for each band of operation. Adjacent-frequency rejection runs from 15 to 35 dB, depending on the band (80, 40, 20, 15 and 10 meters).

On-the-air results are satisfactory. During the past two Field Days, we experienced *no interference* between radios operating on different bands from the same operating position.

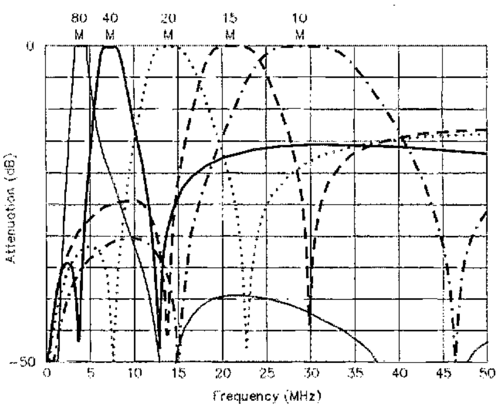


Figure 5—Frequency response of a 5-band interference filter.

The SCRA Field Day filters were very much a club project. I especially want to thank John Breckenridge, WB6FRZ, who rounded up volunteers and ramrodded the construction phase through in time for the 1992 Field Day. Jim Andrews, KC6PJW, had sore hands for days from winding literally dozens of toroids, even with help from Galen Read, KD6HHT; Joe Reid, KC6RKY; Bob Nellor, KC6SOT; and Jim Goodman, KD6JZJ, who also provided the work area and the caloric supplies.

New Products

GROVE SDU 100 SPECTRUM DISPLAY

◊ If you've ever wanted to "see" a real-time computer display of every signal on a given band, but the thought of hooking up an expensive spectrum analyzer to your receiver or antenna system leaves you cold—and broke—consider Grove's new SDU 100 spectrum display unit.

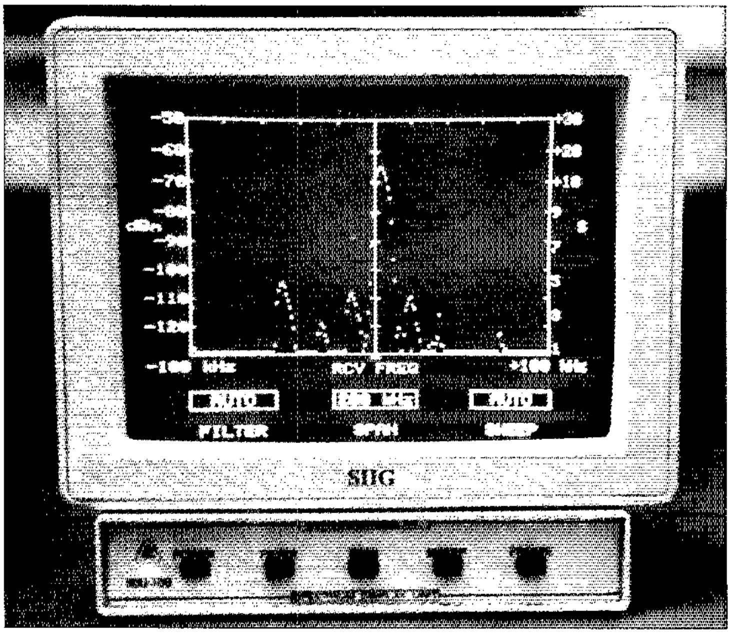
The SDU 100 and its matching 9-inch CRT turn your ICOM R-7000, R-7100 or R-9000 (the unit will work with other shortwave receivers and scanners, some of which require minimal modifications) into a powerful spectrum analyzer. The SDU 100 lets you "see" signals on adjacent frequencies and quickly tune them in. Military and law enforcement agencies take advantage of this technology, and now you can, too!

The SDU 100 works with any receiver that has an IF output of 8.8, 10.7, 21.4, 45 or 70 MHz (10.7 MHz is standard for the SDU). The display's bandwidth can be set from 10 MHz to 0 kHz to fine-tune the readout depending on the frequencies and modes being tuned and the resolution required.

Testing the SDU 100 was a lot of fun. I didn't play around with it for too long because I didn't want to "get used to" using it or I'd be tempted to buy one—at $599.95, it's affordable, as far as spectrum analyzers and spectrum displays go, but $600 is still $600!

Tuning in elusive signals on the VHF public service bands was a snap. Seeing signals above and below my radio's (a loaner ICOM R-7100) tuned frequency appear and disappear added an interesting and useful dimension to VHF/UHF scanning.

The SDU 100 is available from Grove Enterprises, 300 S Hwy 64 W, Brasstown, NC 28902-0098; tel 800-438-8155.—*Kirk Kleinschmidt, NT0Z*



Beginner's Boomers: Two Phased Vertical Arrays for 30 Meters

Try these inexpensive, easy-to-build, low-profile gain antennas!

By Gary D. Borich, W5UDV and Robert S. Logan, NZ5A
1009 Harwood Place
Austin, TX 78704

Among many hams, verticals have a questionable reputation. Some think they radiate equally badly in all directions. Others say they should only be used as emergency fishing poles. Still others regard verticals as necessary evils for apartment dwellers, those with small backyards and close-in neighbors or those wedded to esthetically sensitive spouses.

With over thirty years' amateur experience behind both of us, we feel that, with the work normally given any good antenna system, verticals work very well for both domestic and DX contacts. Indeed, a well-constructed vertical probably gives more bang for the buck than a wire antenna strung through the trees in the backyard, or hung off the side of the average tower.

For the ham who wants to discover the advantages of gain antennas, a phased array of two verticals is a simple way to upgrade the station without a lot of trouble or expense. You can also greet your neighbor with a clear conscience.

This article explains how to build an inexpensive pair of phased verticals for the 30-meter band. Two popular designs of phased vertical are described that will delight the wallet and fill the log with calls and names of interesting friends.

We're both dedicated QRPers. We constructed each design and then conducted tests over several months, using power levels between 250 mW and 2 W. In no case did we encounter special difficulty in making domestic QSOs at these power levels. In fact, we made several DX QSOs with hams in Europe, the Caribbean and the South Pacific.

Designing the Array

Phased arrays vary in design from simple, straightforward 2-element versions set in a straight line to multielement systems arranged in geometric configurations. The benefit of increasing the complexity of a phased array is higher gain and more patterns to choose from. The disadvantage is greater difficulty in delivering correct amounts of RF current to each element, and ensuring that currents to the elements are properly phased. If these two objectives are not achieved, actual patterns for your particular array may not match calculated values. Beamwidth may increase and minor lobes may be introduced behind and around the main beam. In other words, you may not be able to control where your signal goes without considerable physical and engineering effort.

Although these ills can plague simple 2-element arrays as well, techniques for achieving optimum operation of such a

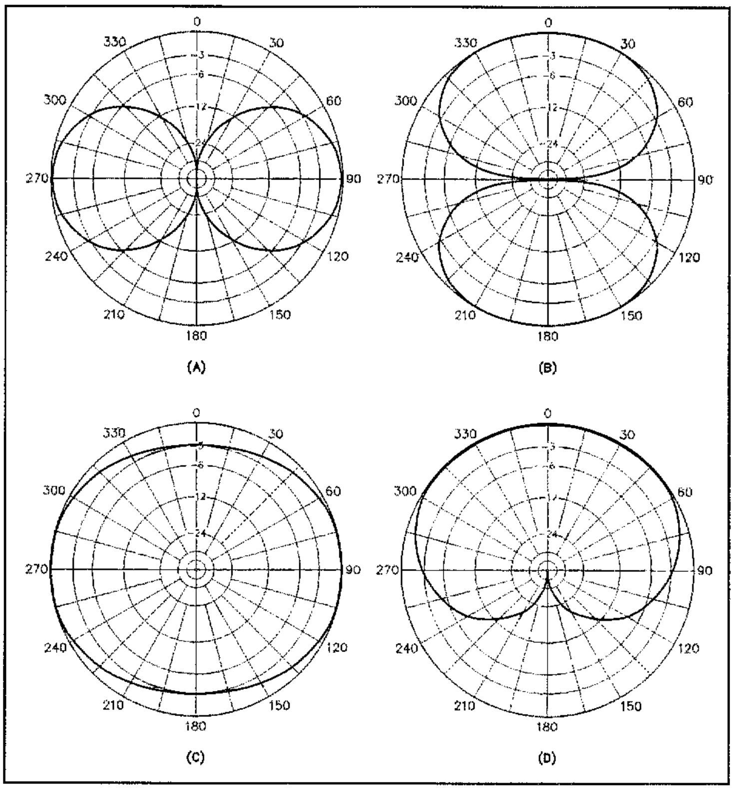


Figure 1—Theoretical radiation patterns for a 2-element vertical phased array. The elements are aligned with the vertical axis, and the uppermost element is the lagging element. At A, λ/2 spacing, broadside; at B, λ/2 spacing, endfire. At C, λ/4 spacing, broadside; at D, λ/4 spacing, cardioid.

system are more easily grasped by the average ham without special test equipment. With two elements properly spaced, for example, you only have to cut the phasing lines carefully and perhaps adjust the length of each element to equalize currents. The benefits of a low-profile beam antenna can be realized for a modest effort. For these reasons, we chose to construct 2-element phased arrays.

The W5UDV design (Figures 1A and 1B) consists of two elements spaced a half wavelength apart and switched to either a broadside or end-fire pattern. When fed in phase with two coax phasing lines of equal length (Figure 1A), the antenna radiates a figure-eight pattern broadside, or perpendicular to the plane of the elements. This arrangement gives approximately 4 dB gain over a single vertical, and a beamwidth of about 60° between half-power points.

When a length of half-wave line is switched into one leg of the antenna, a figure-eight end-fire pattern is then radiated in line with the two elements (Figure 1B). The gain here is about 3 dB over a single vertical. The elements are now referred to as *out of phase*.

At NZ5A, the two elements are spaced a quarter wavelength apart (Figures 1C and 1D). This gives slightly less gain than W5UDV's antenna but creates three possible patterns instead of two and includes a front-to-back ratio similar to a horizontal beam. When fed in phase with two equal lengths of coax phasing lines, a figure-eight pattern again is radiated broadside to the plane of the elements, with a gain of about 1 dB (Figure 1C). In this configuration, although the array acts much like a dipole in terms of directivity, the lower angle of radiation enhances total performance.

When a quarter-wavelength line is switched into either leg of the antenna (*out of phase*), the signal is concentrated in the direction of the lagging leg (Figure 1D). The lagging leg is the antenna into which the extra phasing line is inserted. A unidirectional cardioid pattern results, with gain of 3 dB. Thus, if two elements are placed in a north-south line, the pattern can be switched to cardioid north, cardioid south or broadside east-west.

Building—From the Ground Up

The array consists of two vertical elements, appropriate phasing lines, a relay box and a ground system.

Antenna Elements

The vertical elements are metal poles, 23.5 feet long. Both authors constructed them from commonly available, inexpensive materials. W5UDV used the remnants of an old Hy-Gain 18-AVT vertical, with the original base serving as the bottom 5 feet of one element. To match the two verticals as closely as possible, and maintain equal current flow to each of them, an identical base section for the second vertical element was purchased from Hy-Gain. Another 15 feet of each antenna element consisted of 10-foot and 5-foot sections of 1¼-inch TV mast purchased at Radio Shack. These sections are swaged to fit together, and are connected electrically with a short jumper wire.

Two slip-sleeves, each two feet long, were then cut from pieces of tubing from the old vertical. One sleeve was slipped inside the top of the TV mast as the final extension of the vertical element. The other slip-sleeve was clamped between the base section and the bottom of the TV mast, to allow final vertical height to be adjusted easily from the base of the antenna during tune up. To secure each slip-sleeve, two 3-inch-long crosscuts were made in the tubing. Several 1½-inch hose clamps were then used to squeeze the crosscut sections onto the slip-sleeves. The bottom clamps were loosened to adjust the antenna height and then tightened at the point of lowest SWR.

Each element is mounted on a 1½-inch-diameter thick-wall pipe driven 18 inches into the ground with the pipes spaced 48 feet, 8 inches apart (a half-wavelength at the design frequency). The phasing lines are connected to the coax sockets prefabricated by Hy-Gain into the base sections.

NZ5A used two 20-foot sections of Cyclone rail fencing obtainable at lumber yards. Each main section is extended 3½ feet with a lightweight metal tube that is normally installed in closets to hang clothes on. Sheet-metal screws secure the extender 6 inches into the main section.

The two verticals are mounted on soft-drink bottles that serve as base insulators, spaced 24 feet, 4 inches apart (a quarter wavelength at the design frequency). A sheet-metal screw connects the center conductor of the phasing line to the bottom of each element. Thoroughly scrape the coating from the area on which the center conductor will be connected, to ensure good electrical contact between it and the pipe.

Remember, these simple metal poles have definite beam patterns as a system, so line up the plane of the elements to favor the desired directions. At W5UDV, for example, the two elements were placed beside a stone wall, which provided a north-south end-fire pattern to central Asia and the South Pacific, and an east-west broadside pattern to Africa and the Central and Northern Pacific regions.

At NZ5A, a wood privacy fence on which the verticals are mounted is aligned with the 030° azimuth. Thus, from central Texas, the cardioid north pattern is directed to Europe, the cardioid south to the South Pacific and the east-west broadside to South America and the Caribbean in one direction and Asia and Japan in the other.

Phasing Lines and Relay Box

At both authors' locations, phasing lines for broadside operation are each a half-wavelength long. For end-fire operation at W5UDV, another half-wavelength line is inserted into one phasing line. For cardioid operation at NZ5A, a quarter-wavelength line is switched into one phasing line or the other to beam either north or south.

The electrical length of a phasing line is shorter than its physical length in terms of wavelength because a radio signal travels slower in physical material than in free space. Thus, one cycle at 10.1 MHz travels 30 meters in free space but only 24 meters in RG-58 foam-insulated coax. The difference is due to the *velocity factor* of the material through which the RF travels. The *ARRL Antenna Book* gives the velocity factor for many different types of coax.

The specific electrical length of a half-wavelength phasing line is:

$$L = \frac{492 \times \text{velocity factor}}{\text{design frequency (MHz)}} \qquad \text{(Eq 1)}$$

For quarter-wavelength lines, substitute 246 for 492 in Eq 1.

We learned a hard lesson cutting phasing lines. Although many references state that two elements fed with lines of equal length result in a broadside figure-eight pattern, none mentioned that certain equal lengths are better than others. We found through experimentation that equal lines, each a half-wavelength (or multiples of a half-wavelength) long, resulted in the best match between antenna and feed line for broadside operation. The reason may be because a half-wavelength line repeats source (antenna) impedance at the feed line. Shorter lengths, though equal, caused matching and SWR problems. [Mutual impedance between the elements contributes to the matching problem, especially when switching between two different antenna-phasing schemes.—*Ed.*]

The additional phasing lines for cardioid and end-fire operation are switched in and out of the broadside phasing lines by DPDT relays wired as shown in Figure 2. We both used Radio Shack relays. W5UDV used a miniature PC-mount relay, while NZ5A used two, larger, 8-pin octal-socket relays. The relays are mounted in waterproof boxes at the vertical elements. It isn't necessary to buy a special waterproof container. Bathtub caulk is excellent material for sealing edges and screw holes in aluminum boxes.

Ground System

The ground system at each location was planned for good efficiency and least effort. First, two ground rods were installed at each vertical element. Ground resistance near the antenna drops (and radiation efficiency increases) significantly with the

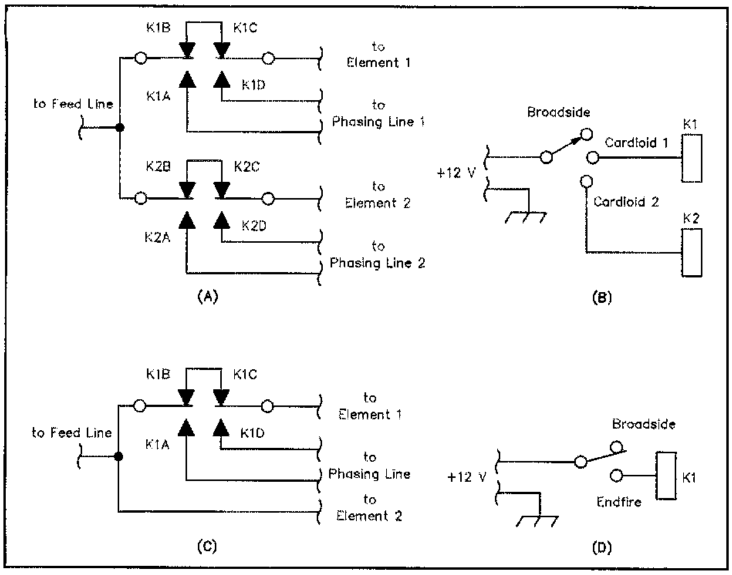


Figure 2—Wiring diagrams for switching phasing lines. A and B show the scheme used by NZ5A. At A, relay box wiring; at B, control box wiring. C and D show the scheme used by W5UDV. At C, relay box wiring; at D, control box wiring.

addition of a second ground rod connected in parallel to the first and independent of the antenna.[1] Think of the ground rods as two 100-Ω resistors in parallel. Separately, each resistor measures 100 Ω, but in parallel the total resistance is 50 Ω. That's what happens with ground resistance at the antenna, although the effect is not as linear with Mother Nature as with manufactured resistors. In practice, using more than two ground rods merely proves the Law of Diminishing Returns.

The ground rods are installed no more than three feet from the vertical element. If the antennas are scaled for other frequencies, the maximum distance *in feet* from antenna element to ground rod should be no more than 0.1 wavelength at the frequency where the antenna will be used. Heavy wire, such as #8 or #6, or even triple strands of #12 wire, connects the ground rods together and to the grounded shield of the coax feed line.

Second, an extensive ground screen was installed beneath and around the vertical elements. A ground screen greatly reduces ground loss because screen has a higher pack density than individual, linear radial wires laid out every few degrees or tens of degrees. With materials available today, you no longer have to resort to thin-gauge "chicken-wire" (hardware cloth) mesh that creeps up every Saturday morning to exactly the height of the lawn-mower blade.

W5UDV laid down 4×4-inch pattern-wire fencing as the ground screen. Three sections of fence measuring 4×8 feet were laid in a 180° arc around each antenna, and two additional sections were laid directly between the two verticals. The latter two sections are important for ground return during broadside operation, since maximum radiation occurs midway between the two elements. All fence sections of ground screen were bolted together or connected by large-gauge wire braid. In this manner, 256 square feet of metal ground were placed beneath the antenna in relatively short order.

NZ5A used the welded steel reinforcing mesh that is laid down in driveway or patio forms before concrete is poured. This so-called "re-mesh" was obtained free of charge from houses being built in the neighborhood. Builders were only too glad to have the scrap material carted off at no cost to them. (But ask first!) After a few weeks of collecting some rather nice bundles of wire mesh, 540 square feet of nicely welded, heavy-gauge steel wire were rolled down in the yard in about an hour and bolted together. Be sure to scrape the area under the bolts to bare metal and spray paint the tightened connection. These precautions ensure good electrical connections for years to come.

So the goal of "least effort" was set. How about efficiency? We think our DX results speak for themselves!

Tuning the System

The key to an effective 2-element phased vertical array is equal current division between the vertical elements and low and similar SWR on both elements. If these are achieved and you have a good ground system, actual patterns will conform closely to published patterns. In nontechnical language, if you aim it, they will come.

The RF current between the vertical elements should be divided equally. If your power is 1 W at the junction, your wattmeter should register ½ W at each element, less the feed-line loss. When this is not the case, the typical reason is because SWRs differ between the elements, due to three different types of impedance that occur in phased arrays. Self-impedance is the impedance at the base of each antenna with the other antenna disconnected. Mutual impedance is the impedance at the base of each antenna with the other antenna connected. Finally, system impedance is the impedance presented by both elements at the transmitter or junction of feed line and phasing lines with both antennas connected.

Because all three impedances influence a phased vertical system, different SWRs can be obtained at each point in the system. The goal of tuning the system is to obtain a low and equal SWR at the base of the elements and a low SWR at the junction of feed line and phasing lines, or at the station. With phasing lines cut to proper length and the ground system in place and connected, tune for best SWR as follows:

1. Starting with one vertical element at a time and with the other disconnected at the junction between the phasing line and the feed line, feed the active element with sufficient power to obtain a good reading on the SWR bridge and adjust the element length for lowest SWR. Repeat this process with the other element, remembering to disconnect the element just tuned.
2. With both elements now connected, feed power to the junction of feed line and phasing lines. Again adjust the length of each element in turn for lowest SWR *at the base of the element*. Repeat this step as necessary until the SWR is about equal across both elements and as low as possible.
3. With both elements connected in the "in phase" (broadside) pattern and being fed power, measure the SWR at the junction and the shack. The two readings should be the same.
4. If the SWR readings are substantially unequal, you may have miscalculated the length of your feed line. It should be a half-wavelength or multiple.
5. If the SWR readings are about equal but above 1.5:1, carefully adjust the length of each element in small steps.
6. If the SWR readings are about equal and measure 1.5:1 or less, quit fooling around and get on the air!

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[1] Notes appear on page 40.

Differences of ½ SWR unit or less across the elements are of little consequence, since that represents less than 0.2 dB difference in the effective radiated power between the elements.[2] When the difference significantly exceeds this level, the array will still work, but not as planned. Operational results will include unexpected patterns and minimal gain. If a low and equal SWR is difficult to obtain, look for physical causes on and around the two elements. Maybe the two elements were built slightly differently, or of unequal heights. One element may be obstructed in the near field while the other is more in the clear. Perhaps the ground system under one element is laid out differently. If the problem is not readily visible, you can build an inexpensive antenna tuner and place it at the base of the element to tune out SWR.

Great Performances

In four months of casual operation using 2 W or less from central Texas, the authors worked a total of 23 states and 7 countries on most continents. After about half the contacts were made, we reduced power to our normal operating power of 1 W. We continued to make domestic and Canadian contacts with ease, but truly long-haul DX contacts were more difficult to complete at that power level. Nevertheless, only a QSO with the South Pacific required 2 W output; all the rest were worked with only 1 W! A real acid test of performance was operation at 250 mW (¼ W) from NZ5A. Stations in Colorado and California answered CQs, and gave RSTs of 559 and 449.

Directivity matched expectations as well. For example, when W5UDV switched from broadside (east-west) to endfire (north-south) after initial contact with John, KH3AE (Johnston Island, in the Northern Pacific), John reported being unable to hear W5UDV until the antenna was returned to broadside operation.

Conclusion

Both phased arrays described are delightful additions to our stations. They seem to perform consistently better than our inverted Vs, dipoles, and single verticals, even at close-in distances of 600 miles or so. Occasionally, the horizontal antenna will be better to one location but we almost never touch the antenna switch during a stint at the rig. If you're itching to upgrade your station with an inexpensive, easily constructed, low-profile gain antenna, the Beginner's Boomers may be for you.

Notes

[1] W. Orr and S. Cowan, *All About Vertical Antennas* (Wilton, CT: Radio Publications, 1986).

[2] F. Brown, "Antenna Gain Measurements, Part 1: Techniques,"*QST*, Nov 1982, pp 35-37.

New Products

GLOBE ON YOUR PC

◊ You may have seen amateur satellite-tracking software, but how about a program that puts *you* in space looking down? It's billed as "The Geography Program That Puts You in the Pilot's Seat," and *On Top Of The World* fulfills its promise. It isn't an atlas and has no maps. Instead, it lets you fly around a three-dimensional model of the Earth, whose position and sunlit side is correctly depicted for the time and date. You can move and look in any direction, and go to almost any altitude. It combines the fun of a flight simulator with learning about the planet. It's the closest you'll come to flying in space. You can select political- or physical-style views and whether individual states are shown; choose colors used for grids, cities and boundaries; measure the accurate distance between any points on Earth (not just major cities); change the time to see when the sun will rise or set anywhere; watch the cities light up when the sun sets; test your knowledge with the Quiz function; and get help at any time from the menus.

A prototype version was shown at the Educational Software Fair at the Museum of Science in Boston in March of 1993 and at the Boston Computer Society Mega Meeting last April. The finished version has refinements, improvements and new features, many of which were suggested by the museum and BCS visitors, including the ability to find any of 5000 geographic features from anywhere by typing a partial name and to identify those features, including oceans and seas, by clicking on the screen; display a screenful of information about any country or US state; display the correct local time for any place on Earth by clicking on the location; display the shortest Great Circle route between any two points on Earth; automatically simulate the view from a spacecraft orbiting Earth; export views of the Earth as standard .PCX graphics files for use in school or business reports. The current version doesn't import standard Keplerian elements, though; perhaps a future update will—and you could ride along with *Mir* or the space shuttle as a "virtual stowaway!"

You can launch *On Top Of The World* from DOS or Microsoft Windows. It requires an IBM-compatible PC with an 80386 processor (80486 recommended), VGA graphics, a mouse, 1 megabyte of RAM and 7 megabytes of hard disk space. *On Top Of The World* is appropriate for adults and any child who can use a globe or atlas. The human interface is easy enough that even a young child can learn the basics in minutes. Exploration Software, PO Box 961, Groton, MA 01450-0961; tel 508-649-3823

Strays

FAST CODE IN FLORIDA

◊ The West Palm Beach (Florida) ARC held a high-speed CW contest on October 23, 1993, during the Palm Beach County hamfest. The champs were (l-r) Don Murray, W4WJ (Winner, 41-60 Age Group), Edwin Roller, K4IA (Overall Winner), and Bill Jochimsen, K4ZK (Winner, Over-60 Age Group). Edwin copied with pen and paper at 41.18913 wpm for 2.02017 minutes. Other entrants earned Morse code proficiency awards. Officials received many compliments that the event had been well prepared and carried off successfully.

The contest officials have issued a challenge: The world speed record of 75.2 wpm, set by Ted McElroy in 1939, remains unbroken after 55 years. Can you produce at least one minute of error-free copy to top his record? Can anyone? Now planned as an annual event, all amateurs and code enthusiasts are invited to the High-Speed CW Contest at this year's hamfest on Saturday, October 22, 1994, at the South Florida Fairgrounds in West Palm Beach.

How did they clock the winning copy rate so precisely? Ted Herrman, AE8G, of Holiday, Florida, wrote a computer program for IBM-compatible PCs, *CWSPEED*, to precisely measure the speed and amount of the contestants' copy. It's available in a compressed .ZIP file with a keyboard-emulation program called, *CW*, and its documentation, by Dave Freese, W1HKJ.

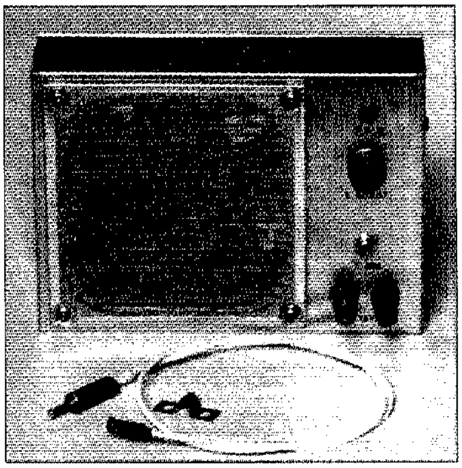
In addition to setting up the computerized contest set-up, Ted sends high-speed code practice on 10.122 MHz Monday and Thursday evenings at 8 PM Eastern time. Dave has custom modified this version of *CW* to send large ASCII files. The original version of *CW* had smaller buffers, but because Ted makes a huge ASCII text file for practice at speeds beginning at 30 wpm and going through 50 wpm, the large buffers in this version avoid a pause in execution when a buffer needs to be refilled with text. Now Ted pushes one key and an entire hour's worth of text is transmitted.

CWSPEED, *CW* and their documentation files, are available in compressed format as CWSPEED.ZIP. It's free for downloading via modem from the ARRL HQ BBS at 203-666-0578 and via the Internet. For details, send an e-mail message to info@arrl.org with the word HELP as the text of your message. If you try them out, drop Ted and Dave notes to say thanks and let them know how you like the programs.

An Automatic Temperature-Controlled Fan

Keep your equipment cool and lengthen its life span with this easy-to-build cooler.

By Bertram S. Kolts, WA0WZI
PO Box 122
Masonville, CO 80541



One of the great enemies of electronic components is heat. High temperatures can severely shorten the life span of many parts, leading to early failure. The prime heat-generating culprits are usually the final amplifiers in our rigs, or their power supplies. High operating temperatures are likely to be encountered if the rig is being operated in continuous key-down modes, where the transmitter's duty cycle is much higher than that of a CW or SSB transmission.

Some ham equipment manufacturers use built-in temperature-controlled fans for cooling, while others rely upon continuously running fans—or no fans at all. The best solution is to add a continuously running fan, but with just a little more effort, you can have the fan turn on only when it is needed. A temperature-controlled fan has the advantages of lower power consumption than a continuously running fan and, since it only runs when it is needed, the noise level in the shack is minimized.

The automatic fan described here uses a thermistor as a remote temperature-sensing element and requires a 12 to 14-V dc power source. The unit shown in the title-page photo is built in a 5x7x2-inch cabinet since there was no room for the fan inside the power supply. If your power supply (or rig) has enough room in the right area, you may be able to mount the fan in or on it.

Circuit Description

Figure 1 is the schematic of the temperature-controlled fan. The temperature-sensing element is a thermistor, RT1. Thermistors are devices whose resistance varies with their temperature. Most thermistors are designed to have a negative coefficient of resistance. That is, as the temperature increases, their resistance decreases. The temperature at which the fan turns on is determined by the resistor divider consisting of RT1, R1 and R2 and the base-emitter voltage of Q1. You set the desired turn-on temperature using R2.

In this circuit, as the temperature increases, the resistance of RT1 decreases

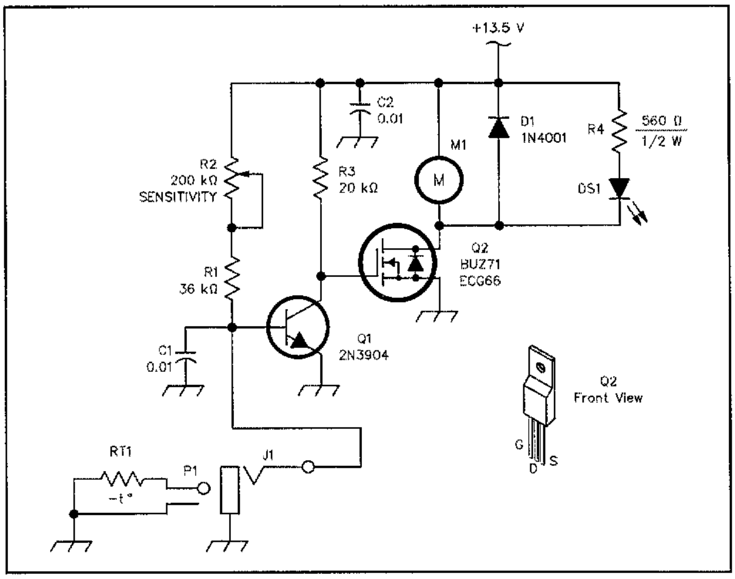


Figure 1—Schematic of the temperature-controlled fan circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are 5%-tolerance carbon-composition or film units.

C1, C2—0.01-µF, 50-V disc ceramic.
D1—1N4001.
DS1—LED.
J1—1/8-inch-diam phone jack.
M1—12 to 14-V dc muffin fan.
P1—1/8-inch-diam phone plug.
Q1—2N3904.
Q2—BUZ71, ECG66, or Digi-Key BUK-453-100APH-ND N-channel, enhancement-mode MOSFET. (Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel 800-344-4539, 218-681-6674, fax 218-681-3880.)
R1—36 kΩ, 1/8 W.
R2—200-kΩ, 1/2-W panel-mount potentiometer.
R3—20 kΩ, 1/8 W.
R4—560 Ω, 1/2 W.
RT1—10-kΩ thermistor. (Radio Shack 271-110, Digi-Key KC003T-ND or KC009G-ND.)
Misc: enclosure, hardware, binding posts.

and the voltage at the base of Q1 decreases. When the voltage at the base of Q1 drops below 0.6 V, Q1 is turned off and this allows the gate of Q2 to be pulled high by R3. This turns on Q2, which pulls the negative side of the fan to ground, turning it on. D1 prevents negative voltage spikes from damaging Q2 when the fan is turned off. DS1 is used to indicate that the fan is on. It is included to make turn-on adjustments easier. C1 and C2 help keep any stray RF fields from interfering with the circuit operation.

A word of caution is in order. Small dc fans typically draw from tens to hundreds of milliamperes during normal operation, but they can require as much as several amps at startup. The 4-inch muffin fan I got from my junk box ran at 300 mA, but it draws 1.8 A at startup! If you make a substitution for Q2, be sure that it can handle the fan's startup current.

Construction and Adjustment

Since only a few components are required for this circuit and because the layout is not critical, the circuit can be built using "dead bug" or "ugly" construction methods. A PC board is also available.[1] My circuit components are mounted on a small piece of universal PC board. No heat sink is needed for Q2, because it dissipates a very small amount of power.

The thermistor is soldered to the end of a length of small-diameter shielded cable and mounted in a short length of metal tubing. The metal tubing protects the fragile thermistor and acts as a thermal conductor. You can glue the tubing in place with epoxy or a silicone sealant. (If the end of the probe in the title-page photo looks familiar to you, that's because it's a .22-caliber shell casing—the amateur is always resourceful!)

Be sure that the leads of the thermistor *do not* make contact with the tubing, since the tubing will later be attached to a heat sink that might not be at the same potential. Use a bit of sleeving, or silicone sealant, to cover the thermistor and the leads to isolate them before the metal tube is glued over the thermistor.

The probe assembly is attached to the heat sink with a metal cable clamp in a location that will not be directly in the path of the fan's air flow. If the air from the fan blows directly on the thermistor, the thermistor will measure the air temperature, not the heat-sink temperature. The best solution is to mount the thermistor on the side of the heat sink opposite to the fan.

Adjustment is easy. Assume that the fan is set up to cool the rig's power supply. Attach the thermistor to the heat sink and connect the power and ground leads for the fan to the 13-V output of the power supply. Turn on the power supply and adjust R2 until DS1 and the fan are off. Continue to turn R2 in the same direction as far as it will go. This ensures that the fan won't come on while the power supply is warming up.

Let your power supply (or rig) thermally stabilize for about an hour without much of a load on it, other than perhaps your receiver. Then, turn R3 until DS1 and the fan turn on. Next, turn R3 back a bit until DS1 and the fan turn off and leave it in that position. As the heat sink warms up while transmitting, the thermistor senses the temperature change and turns on the fan. If you want the fan to turn on at a higher temperature, then simply rotate R3 a bit more in the *off* direction.

Summary

My automatic fan has been in operation for several months cooling my rig's power supply. Before the fan was installed, the heat sinks got so hot that they couldn't be touched.[2] (Maybe I ragchew too much!) Now, it takes about a half hour of operation for the fan to turn on and the heat sinks never get more than warm to the touch.

I would like to thank Brian, N1IEG, for suggesting a 12-V dc fan as a solution to my power-supply's hot heat sinks.

Bertram has been a ham for 34 years and currently holds an Advanced license. In 1975, he received a BSEE from Colorado State University. For the last 25 years, Bertram has been employed by Hewlett-Packard. His work primarily involves production and test engineering for data acquisition and precision measurement products.

Notes

[1] A PC board is available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price $3.75 + $1 shipping.

[2] This "rule of touch" doesn't necessarily mean that a heat sink is working beyond its capacity, but it doesn't hurt to keep things cool.

New Books

LOW POWER COMMUNICATIONS VOLUME 2, ADVANCED QRP OPERATING

Edited By Richard Arland, K7YHA

Tiare Publications, PO Box 493, Lake Geneva, WI 53147; 130 pp, B&W photos and tables; 8½x11 inches, $19.95. Available from dealers or direct from Tiare Publications ($2 s/h in the US and Canada, $3 s/h elsewhere).

Reviewed By Jim Kearman, KR1S
Assistant Technical Editor

I had the pleasure of reviewing the first volume of *Low Profile Communications* in November 1992 *QST*. I've been looking forward to the second volume since. Rich Arland wrote the first volume entirely by himself, but he chose to act as a contributing editor for this effort. Rich edited the QRP column in *Worldradio* for years (he now edits its Satellites column) and is well-versed in QRP operating. Still, his decision to ask other QRPers to contribute gives us a chance to learn from many, rather than one. And the list of contributors is impressive: Randy Rand, AA2U, on DXing; Bob Patten, N4BP, on Contesting; Paula Franke, WB9TBU, on DXpeditioning; Bill Smith, WA6YPE, and Bob Moody, K7IRK, on Milli/Microwatting (running less than 1-W output); "Red" Reynolds, K5VOL, and Fred Turpin, K6MDJ, on Field Day; Jim Thompson, W4THU, on Antennas; Mike Bryce, WB8VGE, on Solar Power; Rich Rinehimer, KA3QKI, and Rich Arland on Amateur Satellites.

The best way to describe this book would be to call it the equivalent of attending the QRP forums at Dayton for 10 years. Then you might gain access to all the information contained within its covers. Randy Rand, AA2U, who wrote the chapter on DXing, has worked more than 100 countries on each of seven HF bands with 5 W or less output. Yes, 5BDXCC, plus endorsements for 17 and 12 meters (the ARRL doesn't endorse DXCC for QRP operation, but I've been on the other end of pileups when Randy was calling me, and his operating skill can't be denied). Randy may not tell all he knows, but his chapter alone defrays much of the hefty cost of this book.

Not surprisingly, QRPers tend to be excellent operators. Having to substitute technique for horsepower does that for you. It's unfortunate that this book may never be read by those who choose not to run QRP. If a technique works for a 5-W station, it will surely work for a station running more power. On the other hand, I guess we QRPers are glad that more hams don't know our secrets!

The chapter on satellite operating is especially welcome. As the sunspots go the way of public morals, satellites offer us a reliable means of long-distance operation. The Russian RS series is accessible to anyone with an HF rig, so most QRPers can try out the "birds" before investing in UHF equipment for the other modes.

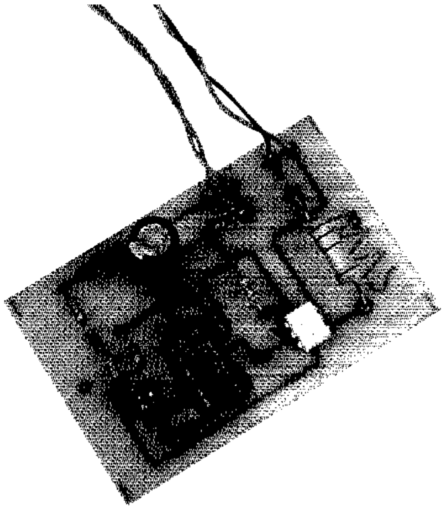
Another chapter I particularly enjoyed was the one on operating with less than 1-W output. This may sound radical, but hams were experimenting with fractions of a watt more than 70 years ago. Milliwatting isn't for everyone, but there's much to learn about minimizing losses in your antenna system—not to mention the propagation knowledge required to succeed. One of the authors of this chapter has worked all 50 states with less than 2 mW output! Who needs $800 final-amplifier tubes?

Considering that so much VHF, UHF and microwave operation is already done at QRP power levels, it's a shame that many hams think HF operation must start with a minimum of 100 W output. In less-complicated times, most hams got started with rigs that ran 2 to 50 W output. Although many of them went on to assemble stations that keep amplifier manufacturers smiling, some recognized the purity of simple, low-power equipment. The best and brightest of these amateurs are represented in this book.

Overvoltage Protection for AC Generators

This simple project can prevent expensive Field Day equipment damage!

By Jerry Paquette, WB8IOW
1966 Logans Lane
West Union, OH 45693



KIRK KLEINSCHMIDT, NT0Z

Most hams look forward to the challenges of Field Day: setting up and operating around the clock under less-than-ideal conditions. Gathering such items as antennas, coax, tables, chairs and tents is easy. Hams are reluctant to loan transceivers and computers, though, because of the possibility of damage by human error, such as coffee or soda spills.

And then, there are the generators. Using portable generators, there is always a possibility of damage to expensive equipment as a result of generator failure, especially from overvoltage.

A Field Day team operating in Class 3A (three transmitters plus a Novice/Technician station) could very easily have over $5000 worth of equipment in use. If the generator supplying power to this equipment puts out too much voltage, you run the risk of burning up power supplies or other electronic components. This article addresses the problem of increased voltage, not lower voltage or surges and spikes lasting for a few microseconds.

Using a portable generator overvoltage protection device (PGOVPD) 120-V circuit and ground-fault circuit interrupter (GFCI) is good insurance. *This overvoltage protection device must be used in conjunction with a GFCI at each station*!

Circuit Description

Refer to Figure 1 for this description,

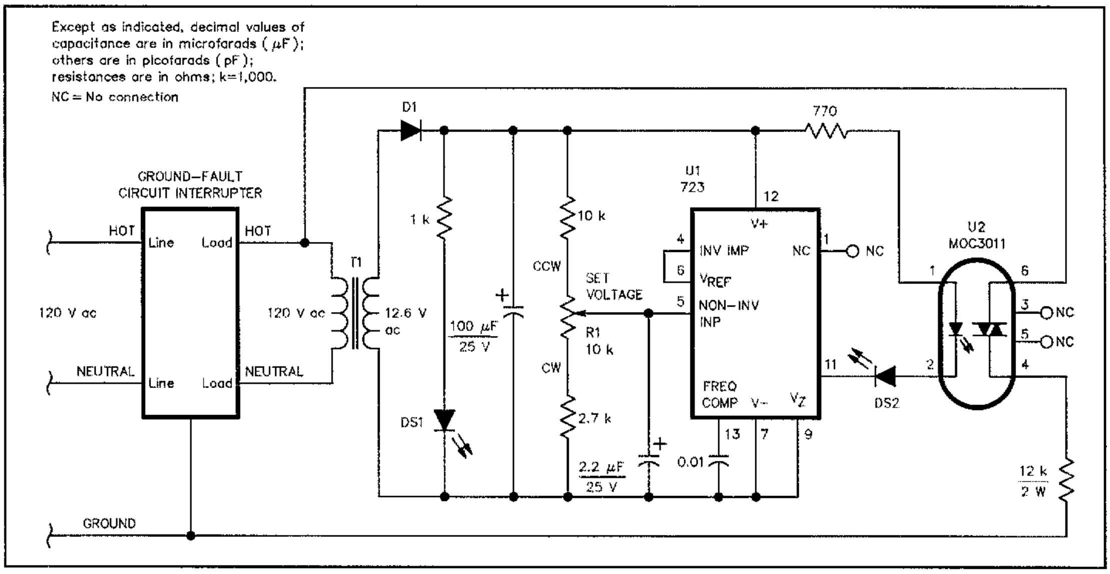


Figure 1—Schematic of the Field Day equipment overvoltage-protection circuit. This circuit must be used in conjunction with a ground-fault circuit interrupter (GFCI). A separate GFCI must be installed at each station. Resistors are ¼-W, 5% tolerance, unless otherwise specified.

D1—200 PIV, 1 A diode; 1N4003 or equivalent.
DS1, DS2—Small LEDs.
R1—10-kΩ board-mounted, multiturn potentiometer.
T1—12.6-V ac transformer (see text).
U1—723 adjustable voltage regulator IC.
U2—Optoisolator with triac output; Isocom MOC3011, MOC3021 or MOC3041, or equivalent.
(All parts available from Digi-Key, 1-800-344-4539; FAX: 218-681-3380).

R1 places an intentional fault on the load side of the GFCI. With the value resistor used, the fault is limited to 100 mA. (The normal tripping threshold of a GFCI is 5 mA.) This current forces the GFCI to trip in just a few milliseconds. This circuit will not function at all without the use of a GFCI. *A GFCI must be used at each station.* If a single GFCI were used at the generator, rather than one at each location, premature tripping could occur. Several hundred feet of extension cords could have enough leakage to trip the GFCI.

You can see that the GFCI has separate lines (inputs) and loads (outputs). GFCI input terminals must be connected to the generator output. The GFCI ground must be tied to the ground of the generator. The load (computers, radios, etc.) will plug into the GFCI or are wired to the load side of the GFCI. Likewise, this over-voltage protection device must be connected to the load side of the GFCI via a standard 3-conductor plug and can be mounted in a separate box.

T1 can be any 120 to 12.6 V transformer capable of delivering 100 mA or more. Mounting of this transformer varies depending on the type used. All remaining components mount on a circuit board. D1 rectifies the ac from T1; the 100-µF capacitor filters the dc. This voltage provides the power to the 723 voltage regulator.

What's a GFCI?

Article 100 of the National Electrical Code, published by the National Fire Protection Association (NFPA) says: The ground-fault circuit interrupter is "A device intended for the protection of personnel that functions to deenergize a circuit or portion thereof within an established period of time when a current to ground exceeds some predetermined value that is less than that required to operate the overcurrent protective device of the supply circuit." The GFCI can be purchased from any electrical supply house or most hardware stores that carry electrical equipment and usually costs about $10.

Two fixed resistors and a potentiometer form the voltage-divider network supplying voltage to the LM723 input, pin 5. R1, the board-mounted[1] potentiometer, has only three leads, but there are four pads on the circuit board, to accommodate different styles of pots. The 2.2-µF capacitor provides a slight delay, to prevent false tripping when the circuit is powered up. The 0.01-µF capacitor from pin 13 of the 723 to the negative supply bus should always be used. When the pin 5 voltage goes higher than the reference voltage at pins 4 and 6, pin 11 goes low, turning on the external LED and the LED inside the optical coupler. LED current is limited by the 1-k resistor. The optical coupler turns on the triac, which places a 10-mA fault between the hot wire and ground of the GFCI.

Adjustment

Adjustment is simple. You'll need a variable ac transformer (*Powerstat* or *Variac*). Turn R1 fully clockwise and use the variable transformer to adjust input to 130 V ac. Turn the pot counterclockwise until the GFCI trips.

[1] PC-board templates are available from ARRL. Send a self-addressed, stamped envelope to: Technical Department Secretary, 225 Main Street, Newington, CT 06111-1494. Ask for the Overvoltage Protection template from June 1994 *QST*.

New Books

JUST WHEN TOM HAD JEAN CONVINCED THAT HIS FRIENDS WERE NORMAL, HE TOOK HER TO A HAMFEST!

By Tom Irwin, AA0ME

The Grandpa Press, 1023 Kirkwood Ave, Iowa City, IA 52240; tel 319-337-9986. First edition, 1994. 103 pp, B&W cartoons. 8×5½ inches. Retail price $8.95, plus $2 s/h. ISBN 0-9640636-0-3.

Reviewed By Brian Battles WS1O
QST Features Editor

Horrifying as it may sound, there are people who consider ham radio a hobby. They think the purpose of becoming a licensed radio amateur, and building and operating radios is to have fun in their spare time. They see it as a recreational pursuit! For such fools, jokes and gag cartoons relating to the noble Amateur Radio service are appropriate, although "serious" hams know better. Keep them away from *Just when Tom had convinced Jean that his friends were normal....By Tom Irwin, AA0ME.*

Tom, who lives in Iowa City, Iowa, was born in 1936 and has been a ham since 1957 (until his nonrenewable one-year Novice ticket expired; he was relicensed as WD0BNO in the early 1970s), and he's been drawing since childhood. After earning a journalism degree, and several years of cartooning for newspapers and magazines, he's turned his sights on the solemn subject of Amateur Radio.

Over the years, a relatively few "regulars" have represented the field of ham cartooning. *QST* often presented the work of Bandel Linn, K4PP, in the 1960s and '70s; Harry Hick (ex-1ESS) drew gags and covers from before WW I to the 1960s. More recently, Jim Massara, N2EST, did cartoons and amusing illustrations for *QST* and ARRL books, and today Bob Beasley, K6BJH, is probably the most widely published ham cartoonist, producing gag panels for the ARRL, *WorldRadio* and other publications. There's a frustrating scarcity of hams who are professional cartoonists. Although someone occasionally mails in a random, awkward sketch, we *QST* editors wish more pros would send us high-quality original work for possible publication. The few submissions we see are generally doodles by radio amateurs who aren't professional artists. It's refreshing to find a ham who has a sense of humor and can draw. Also, unlike others who shall remain nameless, Tom doesn't make an embarrassing attempt to copy the unique style of the late, legendary Phil "Gil" (W1CJD) Gildersleeve, W1CJD, whose classics were a *QST* trademark and a *de facto* standard for ham cartoons for five decades.*

Tom brings a rare breath of fresh air in the ham book marketplace. This collection of 100 Amateur Radio-related cartoons offers hearty chuckles, and even a guffaw or two. Veteran hams, newcomers, and long-suffering nonham spouses and family members will get a charge out of many of the single-panel gags. Is Tom the next Bob Beasley or Gil? Probably not; although the cartoons are clever and neat, his artwork doesn't have Bob's crisp, clean lines or Gil's quaint, artistic style. Tom's material is less formal and sometimes weak in composition, but it has an appeal that places it several notches above much of the work you might find in a typical local club newsletter. Tom's technique reminds me of the simple, hasty-looking drawings done by humorist James Thurber during the 1930s through the '60s for the *New Yorker* and his many books.

There's no doubt that few hams will be able to resist photocopying Tom's cartoons to hang on their ham shack walls next to their FCC licenses, DXCC certificates and QSL cards. This paperback is sure to disrupt the proceedings at many ham club meetings as it's passed around during the treasurer's report or reading of the minutes. But that's okay because sometimes we need a good reason to laugh at our chosen avocation—and ourselves. At 11 cents a cartoon, *Just when Tom had Jean convinced that his friends were normal*...looks like a good value.

*Although it went out of print a year or two ago, the ARRL published an anthology of Phil Gildersleeve's best work in a paperback called *Gil: A Collection of Classic Cartoons from* QST.

Simple, Effective, Elevated Ground-Plane Antennas

Here's an easier and better way to use your grounded tower as a vertical antenna on 160 or 80 meters.

By Thomas Russell, N4KG
29836 Country Lane
Harvest, AL 35749

This article describes a simple and effective means of using a grounded tower, with or without top-mounted antennas, as an elevated ground-plane antenna for 80 and 160 meters.

Grounded towers have been used as shunt-fed verticals on the low-frequency amateur bands for many years. Generally, they required a gamma- or omega-type matching network and an extensive radial system for efficient operation. Recent computer studies reveal that simple elevated radial systems consisting of only four wires can produce results equivalent to 120 buried radials. Typically, these antennas are modeled as isolated monopoles. Presumably, grounded towers could be used with an appropriate shunt-fed matching network.

I've found an even easier method!

From Sloper to Vertical

Recall the quarter-wavelength sloper, also known as the half-sloper. It consists of an isolated quarter wavelength of wire, sloping from an elevated feedpoint on a grounded tower. Best results were usually obtained when the feedpoint was somewhere below a top-mounted Yagi antenna. You feed a sloper by attaching the center conductor of a coaxial cable to the wire and the braid of the cable to the tower leg. Now, imagine four (or more) slopers, but instead of feeding each individually, connect them together to the center conductor of a single feed line. *Voilà*! Instant elevated ground plane.

Now, all you need to do is determine how to tune the antenna to resonance. With no antennas on the top of the tower, the tower can be thought of as a fat conductor and should be approximately 4% shorter than a quarter wavelength in free space. Calculate this length and attach four insulated quarter-wavelength radials at this distance from the top of the tower. For 80 meters, a feedpoint 65 feet below the top of an unloaded tower is called for. The tower guys must be broken up with insulators for all such installations. For 160 meters, 130 feet of tower above the feedpoint is needed.

That's a lot of tower to dedicate to a single-band antenna, especially for someone with limited real estate. What can be done with a typical grounded-tower-and-Yagi installation?

A top-mounted Yagi acts as a large capacitance hat, top loading the tower. Fortunately, top loading is the most efficient means of loading a vertical antenna. The amount of loading can be approximated by using an empirical formula developed by John Devoldere, ON4UN.[1]

Devoldere found that the electrical height of a top-loaded tower can be approximated by:

$$L=0.38F\left(H+\sqrt{2S-H/500}\right) \qquad \text{(Eq 1)}$$

where

L is the approximate electrical length in degrees
F is the frequency in MHz
H is the height of the tower under the Yagi in feet
S is the area of the Yagi in square feet

To check Eq 1, consider the case of no antenna on top, where S = 0. Then $L = 0.38 \times 3.6 \times 65 = 88.9°$, which is very close to the desired 90° quarter wavelength.

The effective loading of a Yagi is the portion of the equation under the radical. The examples in Table 1 should give us an idea of how much top loading might be expected from typical amateur antennas. The term H/500 is ignored as insignificant compared with 2S.

Table 1
Effective Loading of Common Yagi Antennas

Antenna	*Boom Length (feet)*	*S (area, ft²)*	*Equivalent Loading (feet)*
3L 20	24	768	39
5L 15	26	624	35
4L 15	20	480	31
3L 15	16	384	28
5L 10	24	384	28
4L 10	18	288	24
3L 10	12	192	20
TH7	24	—	40 (estimated)
TH3	14	—	27 (estimated)

The values listed in the Equivalent Loading column of Table 1 tell us the approximate vertical height replaced by the antennas listed in a top-loaded vertical antenna. To arrive at the remaining amount of tower needed for resonance, subtract these numbers from the nonloaded tower height needed for resonance. Note that for all but the 10-meter antennas, the equivalent loading equals or exceeds a quarter wavelength on 40 meters. For typical HF Yagis, this method is best used only on 80 and 160 meters.

Construction Examples

Consider this example: A TH7 Yagi mounted on a 40-foot tower. The TH7 has approximately the same overall dimensions as a full-sized 3-element 20-meter beam, but has more interlaced elements. I estimate its equivalent loading to be 40 feet. At 3.6 MHz, 65 feet of tower is needed without loading. Subtracting 40 feet of equivalent loading, the feedpoint should be 25 feet below the TH7 antenna.

I ran 10 quarter-wavelength (65-foot) radials from a nylon rope tied between tower legs at the 15-foot level, to various supports 10 feet high. I tied nylon cord to the insulated, stranded, 18-gauge wire, without using insulators. The radials are all connected together and to the center of an exact half wavelength (at 3.6 MHz) of RG-213 coax, which will repeat the antenna feed impedance at the other end. Figure 1 is a drawing of the installation. I used a Hewlett-Packard low-frequency impedance analyzer to measure the input impedance across the 80-meter band.

An exact resonance (zero reactance) was seen at 3.6 MHz, just as predicted. The radiation resistance was found to be 17 Ω. The next question is, how to feed and match the antenna.

My approach to 80-meter antennas is to tune them to the low end of the band, use a low-loss transmission line, and switch an antenna tuner in line for operation in the higher portions of the band. With a 50-Ω line, the 17-Ω radiation resistance represents a 3:1 SWR, meaning that an antenna tuner should be in-line for all frequencies. For short runs, it would be permissible to

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[1] J. Devoldere, *Antennas and Techniques for Low Band DXing* (Newington: ARRL, 1994).

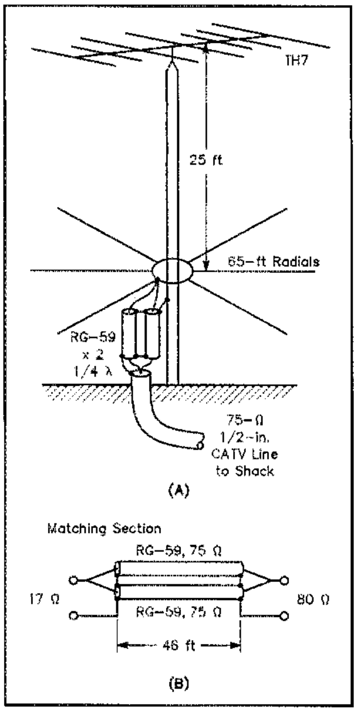


Figure 1—At A, an 80-meter top-loaded, reverse-fed elevated ground plane, using a 40-foot tower carrying a TH7 triband Yagi antenna. At B, dimensions of the 3.6 MHz matching network, made from RG-59.

use RG-8 or RG-213 directly to the tuner. Since I have a plentiful supply of low-loss 75-Ω CATV rigid coax, I took another approach.

I made a quarter-wave (70 feet×0.66 velocity factor = 46 foot) 37-Ω matching line by paralleling two pieces of RG-59 and connecting them between the feedpoint and a run of the rigid coax to the transmitter. The magic of quarter-wave matching transformers is that the input impedance (R_i) and output impedance (R_o) are related by:

$$Z_o^2 = R_i \times R_o \qquad \text{(Eq 2)}$$

For $R_i = 17\ \Omega$ and $Z_o = 37\ \Omega$, $R_o = 80\ \Omega$, an almost perfect match for the 75-Ω CATV coax. The resulting 1.6:1 SWR at the transmitter is good enough for CW operation without a tuner.

The Proof is in the Log

How effective is this antenna? Well, I used to install a 60-foot aluminum tower and 100 radials, 100 feet long in a clear one-acre field every winter, and remove it every spring for mowing. The top-loaded reverse-fed elevated ground-plane antenna has replaced that antenna with no regrets. My only other 80-meter antenna is a dipole at 110 feet, broadside to Europe and the South Pacific.

I use the elevated ground plane for South Africa, South America, the Caribbean, parts of the Pacific and Asia, both long and short path. With it I have worked everything I can hear, with 1200 W output, including HL (rare in Alabama); HS; UA9; UA0; UI; UJ; UL; most of the VK9s; XV; ZS1; ZS8MI; ZS9; 3Y5X; 8Q; 9M2; and 9V1. While running 5 W output, I have even worked two JAs with this antenna, which may say more for its effectiveness than anything else.

Will it Work on 160 Meters?

You bet it will, but it takes another tower. For the 160-meter band, a resonant quarter-wavelength requires 130 feet of tower above the radials. That's a pretty tall order. Subtracting 40 feet of top loading for a 3-element 20-meter or TH7 antenna brings us to a more reasonable 90 feet above the radials. Additional top loading in the form of more antennas will reduce that even more.

Recently, a friend moved to the country, and he needed a 160-meter antenna in a hurry for an upcoming contest. He had stacked TH6s on a 75-foot tower. I suggested he try four elevated radials at 10 feet above ground, with a tuner if necessary. He connected four radials about 120 feet long and a piece of RG-58. The SWR measured under 2:1 and he worked everything he heard in the contest. Figure 2 is a drawing of this installation.

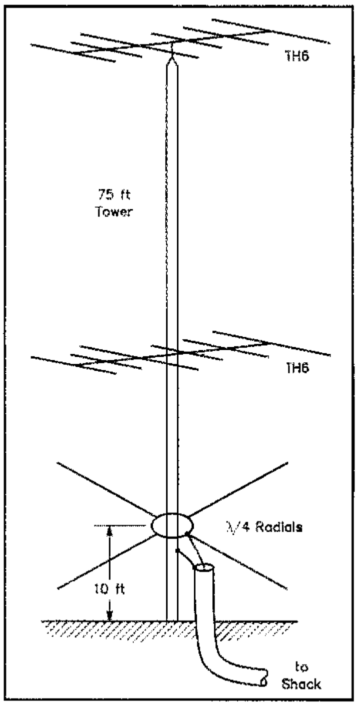


Figure 2—A 160-meter antenna using a 75-foot tower carrying stacked triband Yagis.

Another friend had a 120-foot tower with no antennas on it. He ran four elevated radials at 10 feet and obtained an SWR below 1.5:1 with a 50-Ω feed line. During the contest, he even beat out some big guns.

Elevated ground-plane antennas work! This simple, reverse-feed system makes it possible to feed grounded towers easily and efficiently.

References

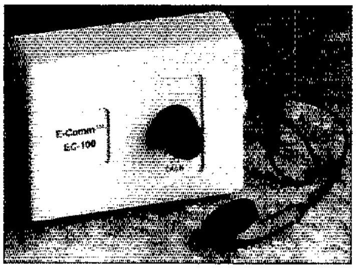
A. Christman, "Elevated Vertical Antenna Systems," *QST*, Aug 1988, pp 35-42 (Feedback, *QST*, Oct 88, p 44).

A. Christman, "More on Elevated Radials," *QST*, Mar 1993, p 72 (Technical Correspondence).

New Products

"SECRET" EARPHONE-MIKE

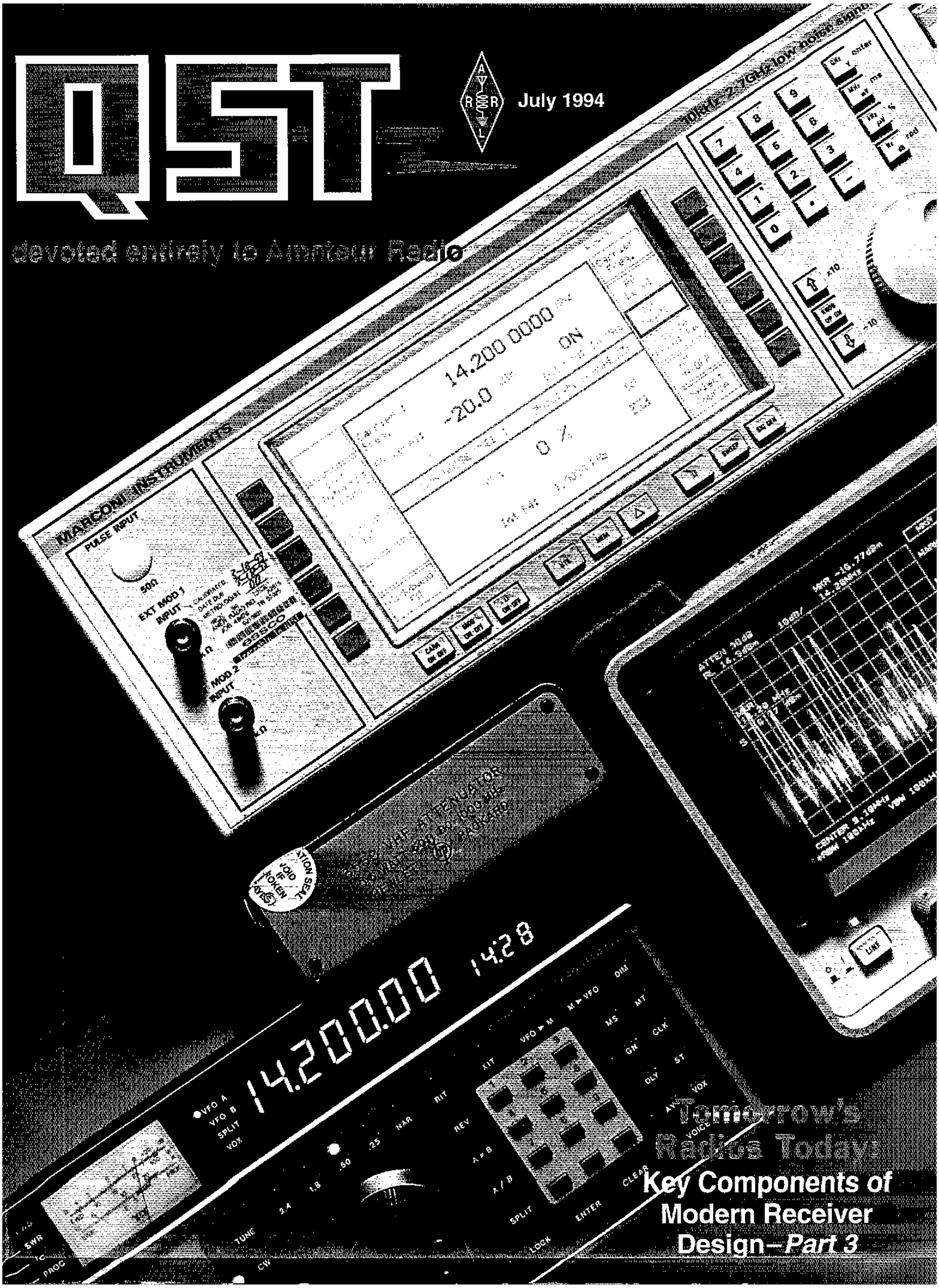
◊ In certain situations, operating a radio may disturb others or you may prefer not to draw attention to your communications. At a hamfest or other event, you don't want to strain to hear and shout to be heard. Here's a solution for hams who want to operate in "stealth" mode: The Ear-Mike is a combination of a tiny microphone and speaker developed for security applications that demand discreet communications. It permits convenient and unobtrusive two-way voice communications. A comfortable in-ear foam earbud allows only the wearer to hear received audio. Voice is transmitted through the same transducer, and is activated by pressing a push-to-talk switch on the interface module. The manufacturer states that even whispers can be transmitted. The miniature interface module is normally clipped to the belt, and features all-metal construction, a gain control and is powered by an internal AA-size battery. It's compatible with most commercial amateur transceivers. Retail price is about $135. Telex Communications Inc, 9600 Aldrich Ave S, Minneapolis, MN 55420; tel 612-884-4051, fax 612-884-0043.



QST

July 1994

devoted entirely to Amateur Radio



Tomorrow's Radios Today:

Key Components of Modern Receiver Design—*Part 3*

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OUR COVER

Is professional-grade radio performance achievable in amateur receivers? "Yes," says Dr Ulrich L. Rohde, KA2WEU, long involved in the design and production of high-performance commercial and military radio gear, "if we are willing to forgo standard techniques in favor of the new." The answer begins on page 42.
(photo by Kirk Kleinschmidt, NT0Z)

CONTENTS

July 1994
Volume LXXVIII Number 7

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

An SWR Detector Audio Adapter

With this adapter, you don't need a meter. You can *hear* your SWR detector responding to the forward and reflected voltages.

By Ben C. Spencer, G4YNM
33 New King St
Bath, Avon, BA1 2BL
England

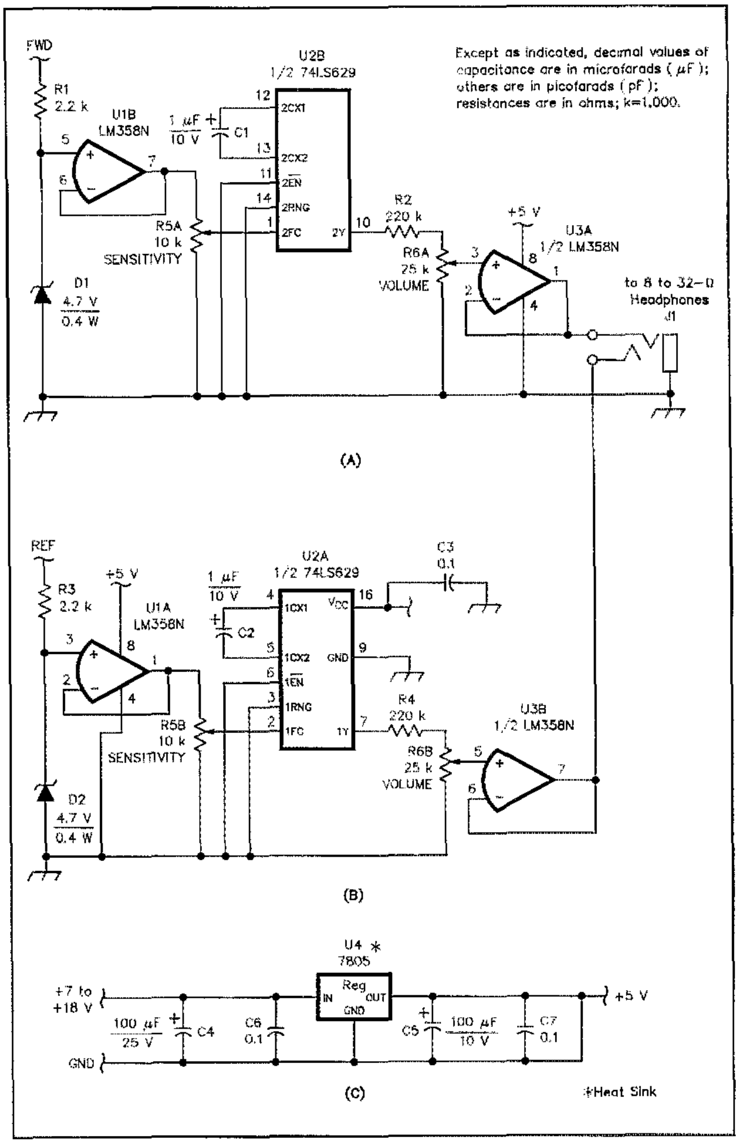


Figure 1—Schematic of the SWR detector audio-adapter circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. All capacitors are disc ceramic unless otherwise stated. The circuits of A and B are identical, each driving one earphone of an 8- to 32-Ω stereo headset. At A, the forward-voltage circuit; at B, the reflected-voltage circuit. A suitable voltage regulator for use with dc power supplies is shown at C.

R5A, R5B—10-kΩ horizontal-mount trimmer potentiometer; optionally, a 25-kΩ dual-gang, panel-mount potentiometer can be used.
R6—25-kΩ dual-gang, panel-mount potentiometer.
U1, U3—LM358N dual op amp (available from Jameco) or substitute an NTE928M.
U2—74LS629 dual VCO (available from Jameco, 1355 Shoreway Rd, Belmont, CA 94002-4100; orders, 800-831-4242, 415-592-8097; fax [domestic] 800-237-6948 or 415-592-2503; fax [international] 415-592-2503; $30 minimum order). Or, substitute an NTE74LS629. NTE devices are available from Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952, tel 800-524-6464, 614-264-6464, fax 614-264-5414; no minimum order, and from Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891, tel 800-866-6626, 401-596-3080, fax 401-596-3590; $10 minimum order.
Misc: PC board, stereo headphone jack, 8- to 32-Ω stereo headphones, mounting hardware.

This SWR detector audio adapter is designed specifically for blind or vision-impaired amateurs, but anyone can use it. Instead of using a meter (or meters) to indicate antenna system forward and reflected voltages, this adapter generates two tones with frequencies that are proportional to the respective voltages. The tones are fed to a pair of stereo headphones (the miniature types are ideal) so that one ear hears the forward-voltage tone and the other ear hears the reflected-voltage tone. (I've set mine up with forward on the left earphone and reverse on the right earphone.) Thus, tuning up a transmitter is simply a matter of tuning for the highest-pitched tone in the left ear and the lowest-pitched tone in the right ear.

Construction and testing are straightforward, requiring no special skills or equipment. All components are readily available.[1] The PC board can be installed in many existing SWR detectors, and the forward and reflected voltages obtained by tapping into the lines leading from the pick-up unit to the respective meters or the selector switch.

Circuit Description

The audio-adapter circuit is shown in Figure 1A and B. Because each half of the adapter circuit operates identically, describing the action of one (the forward-voltage sensing circuit) will suffice.

Most SWR detectors are reflectometers in which sampling lines are loosely coupled to the feed line. RF voltages induced into the sampling lines are diode-rectified. The resulting dc voltages are fed to panel meters that indicate relative forward and reflected power. With this circuit, the forward and reflected voltages are applied to the audio-adapter board and drive voltage-controlled oscillators (VCOs).

The forward dc voltage from the SWR detector is routed to R1, buffered and amplified by U1B, fed to **SENSITIVITY** control R5A and applied to VCO U2B. As the voltage on pin 1 of U2B increases, so does the frequency of the tone output at U2B pin 10. Adjusting the **SENSITIVITY** control sets the range of audio tones produced by the audio adapter. This signal is fed via **VOLUME** control R6A to a simple buffer/audio amplifier (U3A) that drives the left headphone. Zener diode D1 limits the maximum input voltage, partly to protect U1B, but also to limit the upper VCO frequency to about 3 kHz.

Without any dc input, each VCO runs at a low frequency (approximately 380 Hz) to tell you that the unit is operating. Increased voltage on the transmission line—even that from a low-power transmitter—is sufficient to cause the tone frequency to increase noticeably. As the voltage decreases, so does the frequency of the tone.

Construction

A single-sided PC board and template package are available (see Note 1). The PC board is small enough to fit inside most existing SWR detectors and the circuit can be battery operated if required. Simply mount a stereo headphone jack on the SWR detector's front panel to accept the headphone plug.

Note that R6A and R6B are parts of a dual-section, panel-mount potentiometer. R5A and R5B are PC-board mounted trimmer potentiometers. For those who want a panel-mounted **SENSITIVITY** control, a dual-gang potentiometer can be substituted for R5A and R5B.

Testing and Calibration

Once the unit is installed in an SWR detector, connect a 5-V power supply to the audio adapter board. When power is applied, you should hear two identical low-frequency tones in the headphones. Adjust the **VOLUME** controls to provide a comfortable listening level.

Next, connect your transmitter to a dummy antenna via the SWR detector. When you key your transmitter, the tone in the left earpiece should increase in frequency quite dramatically, representing increasing forward power. Theoretically, with a matched line and load, there should be no reflected voltage and therefore, the right-headphone tone shouldn't change. In all probability, however, the tone frequency will increase, but only slightly.

If you use an antenna tuner for matching your antenna system, you'll hear the two tones change frequency according to the degree of mismatch. The best match is indicated by the forward headphone tone reaching its maximum frequency while the reflected (right) headphone tone frequency decreases to its minimum.

Summary

This project is inexpensive, easy to build and sure to find use inside (and outside) many a ham shack. Although this adapter is designed for blind or vision-impaired amateurs, sighted amateurs are sure to find a use for it. Used in an SWR detector as outlined, it frees you from having to take your eyes from the controls you're manipulating to read a meter (or pair of meters).

---

[1] PC-boards are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269; price $2.50, plus $1.50 shipping. A PC-board template package is available free from the ARRL. Address your request for the SPENCER AUDIBLE SWR ADAPTER TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

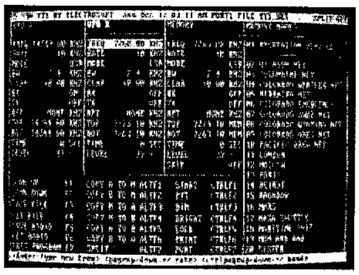
New Products

TEN-TEC KITS

◊ Making its debut in Ten-Tec's new line of Amateur Radio T-Kits is the Model 1208 6-meter transverter. It lets you operate SSB, CW and FM (if available from your HF transceiver). You use it to convert transmitted and received signals from 50 to 52 MHz to the 20-meter band. Three to five watts of 20-meter drive power provides 8 W output on 6 meters. It has RF-sensing transmit/receive switching so the connection to the HF rig is easy. A professionally painted and silk-screened cabinet is included. Retail price is $95. Coming soon are a mobile 2-meter FM transceiver, 2-meter power amplifier and more; request a Ten-Tec T-Kits catalog. Ten-Tec Inc, 1185 Dolly Parton Pkwy, Sevierville, TN 37862-3710; tel 615-453-7172, fax 615-428-4483.

YAESU FT-990 CONTROL SOFTWARE

◊ New computer-controlled software for the Yaesu FT-990 HF transceiver features user-friendly options that can control the rig's frequency, mode, bandwidth and so on, and can store an unlimited number of data files for programming the FT-990's memory channels. Several scanning modes are supported, as is split-frequency operation, simple function-key commands and many others. It requires the Yaesu FIF-232C level converter. Retail price is $100 and it has a money-back guarantee. Electrosoft, PO Box 1462, Loveland, CO 80539.

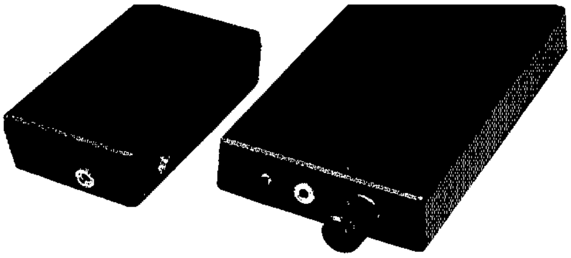


CT VERSION 9

◊ Just in time for Field Day 1994, K1EA Software has released version 9.0 of its popular *CT* contest-logging/operating program. *CT 9* works with 80386/80486 or better PCs, supports 15 contests, controls most transceivers, sends call signs and reports on CW, logs, dupes, scores and compiles statistics, supports *PacketCluster* input/output, and multitransmitting networking.

The latest version also adds 50-line mode, mouse support, sunrise/sunset tables, antenna-relay control, color-coded band maps and an automatic installer program. Registered users get unlimited free telephone BBS support. Retail price is $79.95 plus $4 s/h; upgrade from *CT* ver 8 $44.95 plus $4 s/h. K1EA Software, 5 Mt Royal Ave, Marlborough, MA 01752; tel 508-779-5054, fax 508-460-6211.

A Pocket-Size, *Talking* Morse Code Practice Computer



Here's a go-anywhere Morse-code companion to teach and entertain you.

By Ken Staton, KD6LHB
PO Box 2601
El Granada, CA 94018
Photos by Kirk Kleinschmidt, NT0Z

Have you ever wished you could practice copying Morse code during that boring commute to and from work, during a lunch break, or even while exercising? With the Morse Code Practice Computer (CPC), you can fulfill that wish. This pocket-sized Morse code generator can go almost anywhere you do. The CPC not only generates random Morse characters or groups, it *speaks* the code sent to provide feedback! The CPC costs about $50 to build—about the same price as a portable cassette player and set of Morse code tapes.[1] The CPC is self-contained, there are no tapes to lose or break and—because it generates random code—you'll never become familiar with the character sequence. Rather than the raucous tones produced by some code-practice devices, the CPC generates pleasant *sinusoidal* tones. You can set the CPC's code rate to Farnsworth speeds of 5 wpm (character rate of 16 wpm), 13 wpm (18 wpm character rate), or 20 wpm (23 wpm character rate). CPC software is available (see Note 1) and is easily modified.

Circuit Description

The CPC circuit (Figure 1) is a simple computer with an analog output section. U1 is a relatively inexpensive 8031 microcontroller. The 8031 addresses a full 64 kB of external program memory in a replaceable EPROM (U2), which makes software changes easy. U3 latches the least-significant 8 bits of the address bus from I/O port 0, which multiplexes the address and data.

To minimize the parts count and cost, I use digitized speech rather than a speech synthesizer IC. U4, the 8-bit digital-to-analog converter (DAC), is a single-supply-voltage IC that can be configured for voltage output. To operate U4 in the voltage-output mode, the R/2R ladder is driven backward by U6, a 2.5-V reference connected to U4's current-mode output pin. The DAC voltage output is obtained from the reference pin. U5A buffers the DAC output and drives a Sallen-Key low-pass active filter built around U5B. R4 through R6 set the output level to 0.5 V P-P across an 8 to 16-Ω load. Regulator U7 allows operation from a vehicle battery and 12 and 9-V sources.

A brief description of the software, and the process used to digitize speech, are presented in the sidebar.

Construction

The CPC is easy to build using any construction technique. Aim to keep the analog output circuitry away from the digital circuitry. Using the point-to-point wiring technique, you'll achieve the smallest possible assembly. It's possible to fit the entire circuit in a 0.995×2.25×3.6-inch box (see Figure 2), including a socket for the EPROM. (Use a socket for the EPROM so you can easily make software changes.) A slightly larger version of the CPC (Figure 3) uses a PC board, in a Serpac 051-I box. The PC board is single-sided to encourage individual fabrication.[2]

If you mount the CPC in an enclosure, you'll need to change the **VOLUME** control (R5) to a panel-mount potentiometer, or set the PC-board-mounted R5 for maximum output and use an in-line stereo, variable attenuator (such as RadioShack's 42-2459) with your earphone as a volume control. All components are available from JDR Microdevices and from Jameco.[3] Many components are available from other sources.

For best battery life, use CMOS parts. The CMOS 80C31 is somewhat more expensive than the HMOS 8031. The trade-off here is the long-term cost of batteries versus the initial cost of the microcontroller.

Operation

Headphone jack J1 is a stereo jack, but it's intended for use with a *monaural* headset. (Use an adapter if you have stereo headphones.) Plugging in the headphones turns on the unit: The sleeve of the monaural plug makes contact between the battery's ground and circuit ground.

For use during commuting, you can connect the CPC output to a CD player-to-cassette adapter (such as the Radio Shack 12-1951) and listen to Morse code using your vehicle's stereo system.

The CPC's random number generator produces a pseudorandom sequence. This means that the sequence will be the same for a given seed value. The seed is changed every time the push-button switch is held down during mode changes. To generate a different sequence while retaining the default setting, select **X** from the menu.

Operating Modes

The default mode is set to random characters at Farnsworth 5 wpm (this is what our local VEC uses). Push-button switch S1 (**MODE**) is used to change operating modes. Continually pressing the switch cycles through a verbal menu which says *G, R, S, T, 5, 1, 2,* and *X*. Releasing the switch during or immediately after the desired mode is voiced selects that mode. The modes are: G = 5-character groups, R = random characters, S = sequential characters, T = toggle speech on/off, 5 = Farnsworth 5 wpm (character rate of 16 wpm), 1 = Farnsworth 13 wpm (18 wpm character rate), 2 = Farnsworth 20 wpm (23-wpm character rate), X = eXit with no mode change.

Summary

The code-practice computer generates random code and provides feedback help-

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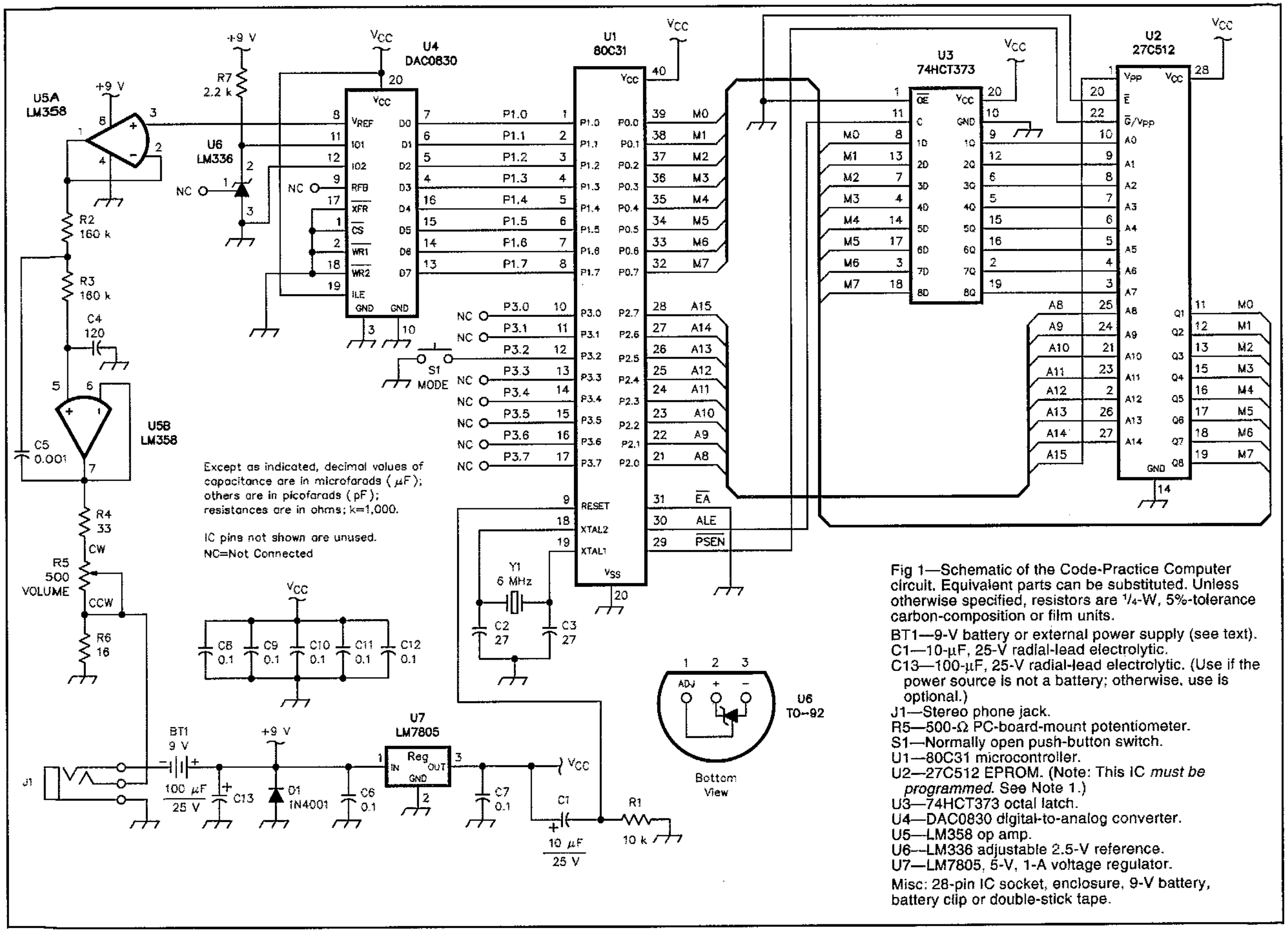
[1] Notes appear on page 28.



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

IC pins not shown are unused.
NC=Not Connected

Fig 1—Schematic of the Code-Practice Computer circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

BT1—9-V battery or external power supply (see text).
C1—10-µF, 25-V radial-lead electrolytic.
C13—100-µF, 25-V radial-lead electrolytic. (Use if the power source is not a battery; otherwise, use is optional.)
J1—Stereo phone jack.
R5—500-Ω PC-board-mount potentiometer.
S1—Normally open push-button switch.
U1—80C31 microcontroller.
U2—27C512 EPROM. (Note: This IC *must be programmed.* See Note 1.)
U3—74HCT373 octal latch.
U4—DAC0830 digital-to-analog converter.
U5—LM358 op amp.
U6—LM336 adjustable 2.5-V reference.
U7—LM7805, 5-V, 1-A voltage regulator.
Misc: 28-pin IC socket, enclosure, 9-V battery, battery clip or double-stick tape.

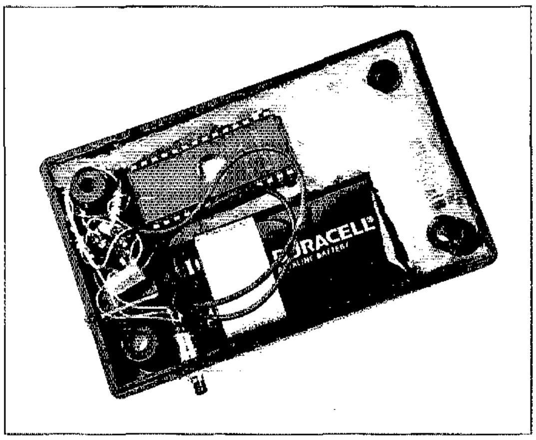


Fig 2—In this small version of the CPC, the micro-controller and attached components have been encapsulated in flexible sealant. The EPROM is left exposed for easy removal and reprogramming.

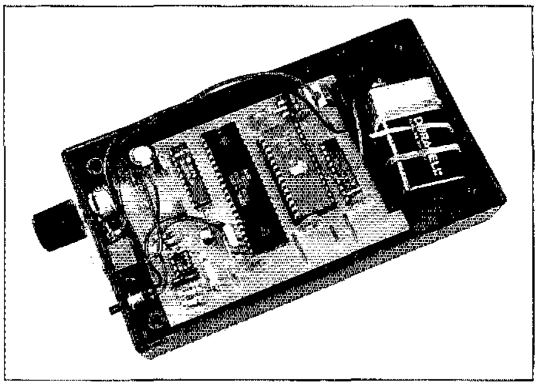


Fig 3—In this PC-board version of the CPC, everything is readily accessible. Double-stick adhesive tape can be used to hold the battery in place. The Serpac 051-I enclosure measures approximately 1x3x5½ inches (HWD) .

ful in learning code quickly. I still use the CPC when I want to practice copying Morse code but can't sit at the radio. My current code speed is 13 wpm; my goal is to reach 20 wpm.

When you no longer use the CPC for Morse code practice, you can turn it into a great single-board computer for another project. I'm sure you'll enjoy building and using this Morse code teaching companion.

The CPC's Software

Most of the software details are included in the source-code files. Since reading program source files is not a fun thing to do, I'll briefly describe the software modules.

I digitize the speech using a personal computer equipped with an 8-bit Sound Blaster sound card, then compress the speech using a first-order adaptive differential pulse-code-modulation algorithm (ADPCM) to achieve 2:1 compression with good sound quality. During operation, the compressed, digitized speech is decompressed and output to the DAC by the CPC software. The CPC is programmed in 8051 assembly. The main program loop is CPO_SPK. CPO_SPK calls subroutines MORSE and SPEAK. MORSE generates code based on a bit pattern encoded in a byte (8 bits). SPEAK speaks digitized speech for the character set (26 + 10 + 3 + 4 = 43 characters).

If you want to make modifications to the CPC program, there are several inexpensive 8051 cross assemblers available. One is *TASM,*[*] a shareware package from Speech Technology Inc. that supports a variety of processors, including the TI 32010 DSP. A comprehensive 8051 reference is *The 8051 Microcontroller: Hardware, Software and Interfacing,* by James W. Stewart.[†]

The binary speech data is generated by a series of programs written in Microsoft Visual C that run on a personal computer operating in the Microsoft Windows environment. These programs can be modified easily to compile with other C compilers for execution on different systems. *SPCH_LIB* input is a text file defining the sound-board data files; it outputs a text file for the assembler that defines the speech table and a binary file to be programmed into the EPROM for speech data. *SPCH_LIB* calls *VOC2BIN* and *ADPCM. VOC2BIN* converts sound-board data files (for the 8-bit Sound Blaster board) into binary speech files. *ADPCM* converts (compresses) binary files into adaptive, differential, pulse-code-modulated files with an approximate reduction of 2:1 in file size.

[*] Speech Technology Inc. 837 Front St S, Issaquah, WA 98027, tel 206-392-8150; CompuServe ID 73770,3612. *TASM* is available on CompuServe in the IBM Programming forum as *TASM.ZIP*.

[†] J. Stewart, *The 8051 Microcontroller: Hardware, Software and Interfacing*, (Simon & Schuster: Englewood Cliffs, NJ 07632), 1993.

Notes

[1] Source code, PC boards, programmed EPROMs and kits are available from The Staton Company, PO Box 2601, El Granada, CA 94018. An IBM-formatted floppy disk with source code is $10; PC board, $15; programmed EPROM, $15; kit (excluding battery and enclosure), $60. Shipping and handling in the US and Canada is $7; all other countries, $15. Payment must be made in US funds drawn on a US bank. California residents must add 8.25% sales tax.

[2] A PC-board template package is available free from the ARRL. Address your request for the STATON CODE-PRACTICE COMPUTER TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE. See Note 1.

[3] JDR Microdevices, 2233 Samaritan Dr, San Jose, CA 95124; tel 800-538-5000, 408-494-1400; fax 800-538-5005, BBS 408-494-1430. Jameco, 1355 Shoreway Rd, Belmont, CA 94002-4100, tel 800-831-4242, 415-592-8097, fax (domestic): 800-237-6948 or 415-592-2503, fax (international): 415-592-2503.

The software is available for downloading from CompuServe, the ARRL BBS (203-666-0578) as *STATON.ZIP*, by FTP on the Internet site oak.oakland.edu/pub/hamradio/arrl/infoserve/qst files area, or from the author on floppy disk.

Ken Staton is an electrical engineer at a large US corporation. He has a BSEE from Stanford and an MSECE from UCSB. Although Ken was interested in Amateur Radio since 1976, he didn't get his license until 1992, when he realized there was a no-code option. As this article was being prepared for publication, Ken upgraded to Advanced. QST

New Products

ADD-ONS FOR YAESU FT-1000

◊ Two CW receiving filters for the subreceiver on the Yaesu FT-1000 HF transceiver let the user select bandwidths of 250 and 400 Hz. Retail price is $130 plus $6 s/h. Tuning Upgraders for the main and sub-tuning controls cut the tuning rate to 2.5 kHz per revolution for finer tuning and bandspread. Retail price $46.95 plus $6 s/h. Robert Pohorence, KB4JMQ, International Radio and Computer Inc, 3804 S US 1, Ft Pierce, FL 34982; tel 407-489-0956, fax 407-464-6386. QST

The Null Steerer Revisited

Here's an improved version of a noise and interference fighter.

By Charles J. Michaels, W7XC
13431 N 24th Ave
Phoenix, AZ 85029
Photos by Kirk Kleinschmidt, NT0Z

In addition to the usual interference problems faced by other amateurs, my location suffers a wide variety of urban noise problems such as noise from lamp dimmers, "green plugs,"[1] line leaks, touch-controlled lamps, industrial processes, TV horizontal-sweep harmonics and a variety of wide and narrowband noise or signals of unknown origin, particularly on 40, 80 and 160 meters. In the summer, thunderstorms in the northern Arizona mountains render 160 meters virtually useless.

Why a Receive-Only Array?

Ordinary 2-element, continuously variable phased arrays covering 40, 80 and 160 meters are hardly practicable in my urban location. The theoretical gain of the arrays is seldom achieved, and the phasing and matching equipment to handle transmitting power in a continuously variable array is not simple, even for a single band. The primary value of such arrays usually turns out to be the ability to steer the null pattern to eliminate noise and interference arriving at any wave and azimuth angle in receiving. However, with every change in phasing to avoid noise and interference, the resulting feed-line SWR change requires rematching before transmitting.

My version of the Webb Null Steerer[2] has proven very effective, particularly with the noise problems on the 40, 80, and 160-meter bands. Those patches of raspy noise or wideband dimmer noise can be completely nulled. I can put a 50-dB null on a nearby 1-kW CW signal that overloads my receiver on 40-meter SSB. I can operate more often on 160 meters in the summertime. I can even copy signals—otherwise completely obliterated—by nulling on the popping to screaming corona discharge preceding nearby lightning strikes. (Hey, Ben Franklin took a chance!) The Null Steerer is more valuable to me than all the QRM and QRN combating features of my modern transceiver.

The system uses a short, loaded second antenna, a phasing device (the Null Steerer) at the operating table and the primary station antenna to form a 2-element receiving array.

Why a New Circuit?

Because the Webb circuit has problems:

- Feedback between the antennas can cause oscillation.
- The quarter-wave 50-Ω coax lines in the Null Steerer box (one is needed for each band) that produce the 90° delay are cumbersome for the lower frequency bands.
- The 250-Ω potentiometers are hard to find and expensive.
- Two amplifiers are used—one each for the **X** and **Y** lines. The amplifier gain is insufficient for small second antennas on the lower frequency bands.
- Lack of tuning permits HF and MF broadcast station overload to produce intermodulation products in the amplifiers.
- The paralleled amplifier outputs directly connected to the primary antenna signal line can cause system oscillation by feedback between the primary and second antennas.
- With the Null Steerer in the main antenna line, the full transmitter output power is routed through the control box (via the cumbersome main antenna coax) for relay TR switching.

I attacked each of these problems through a series of four experimental designs.

The W7XC Mark IV Null Steerer

For experimentation and general use, my fourth version is a band-switched, 10 through 160-meter unit. I'll discuss the circuit in single-band form with band-switching provisions shown optionally.

The circuit is shown in Figure 1; the amplifier portion is shown in Figure 2. The 90° pi network replaces Webb's quarter-wave transmission line. A hybrid summer[3] (R3 and T2) combines the **X** and **Y** potentiometer signals to produce a 360° variable phase and amplitude second-antenna signal, requiring only one amplifier, which is preceded by a tuner to prevent overload and intermodulation by strong out-of-band signals.

See the optional band-switching information in Figure 1 and Tables 1 and 2 to select appropriate pi-network and tuner-coil details for a single-band version, or to implement a chosen set for a multiband version. Note that the 40 and 80-meter bands can be covered by the same tuning circuit, but require different pi-nets. On the higher frequency bands—30 and 20 meters, and 17 through 10 meters—common tuning and pi-network circuits are used.

The amplified second-antenna signal is combined with the primary antenna signal in a second hybrid summer (R4, T3 and T4). When noise or signals from the primary antenna are matched by the same noise or signals of equal amplitude from the second antenna (but shifted in phase by 180°), a null is produced. Other signals are passed through with amplitudes appropriate to their position in the lobes of the resulting array pattern. The summer port-to-port isolation prevents inter-antenna coupling oscillation and isolates the primary antenna line from the amplifier. R5 sets the amplifier gain. R6 provides for attenuation of the main antenna signal while peaking the **TUNING** control.

S1 puts the Null Steerer in and out of operation by switching the transceiver lines and the power supplied to the amplifier and the **IN/OUT** indicator LED, DS1.

This Null Steerer is installed in the antenna line between the receiver and the TR relay using RG-174 coax and shielded phono plugs and jacks at the Steerer. Such a receiver/TR line is available in recently manufactured transceivers. In the Kenwood TS-830 and TS-940, these lines and the shield ground are available at the

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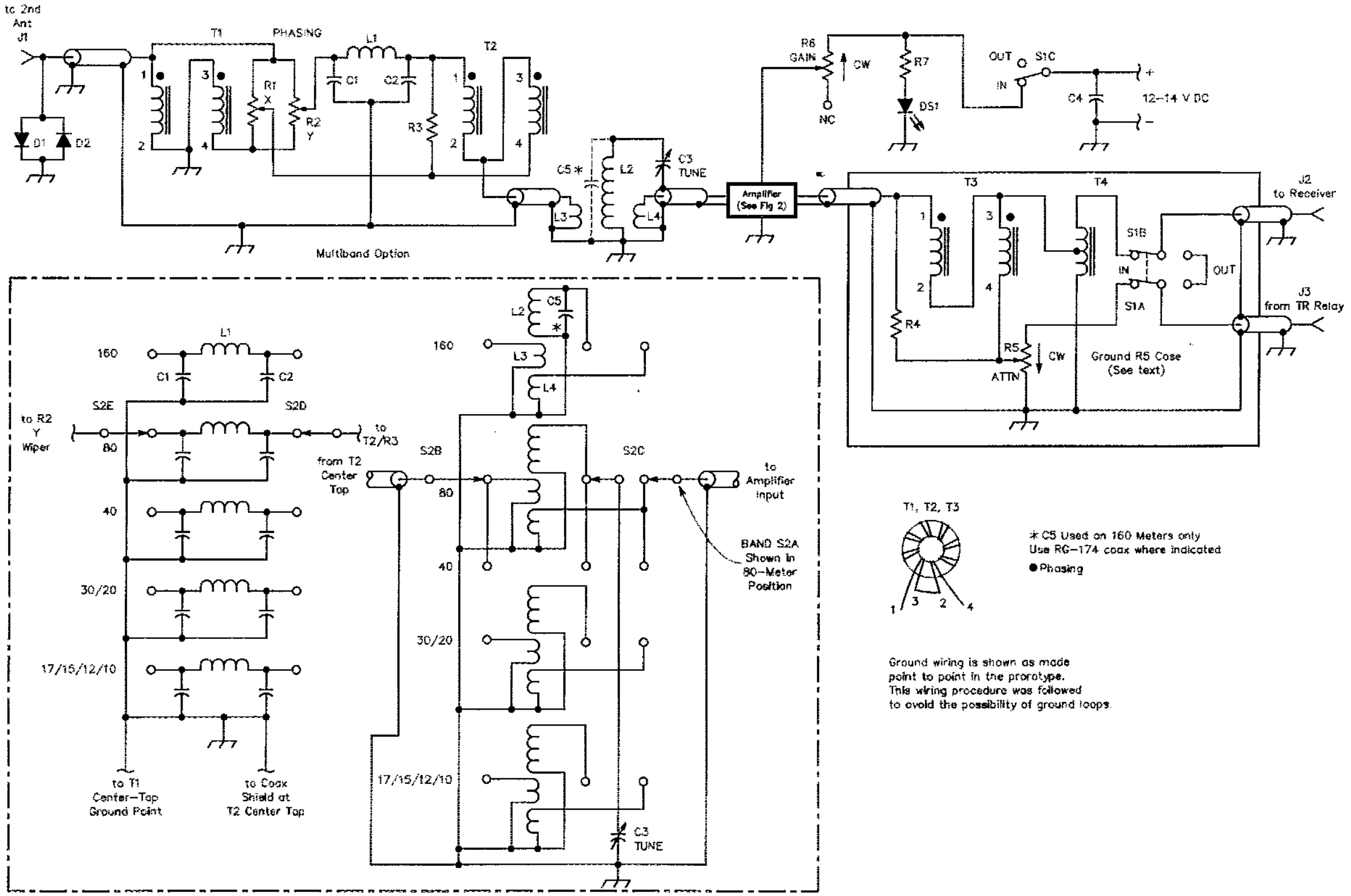
[1] Notes appear on page 33.



* C5 Used on 160 Meters only
Use RG–174 coax where indicated
● Phasing

Ground wiring is shown as made point to point in the prototype. This wiring procedure was followed to avoid the possibility of ground loops.

Figure 1—Null Steerer schematic with optional multiband function. Select the pi-network and tuner component values from Tables 1 and 2. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units.

C3—140-pF miniature air-variable capacitor.
C4—0.1 µF, 25-V ceramic capacitor.
C5—Approximately 300 pF across L2 (for 160 meters only). Combine mica capacitor values to allow C3 to cover the entire 160-meter band.
D1, D2—1N914 or equivalent diodes (276-1122).
DS1—LED.
J1-J3, incl.—Phono jacks (use shielded jacks on external lines).
R1, R2, R5, R6—500-Ω composition potentiometers (such as Mouser catalog no. 31VA205, Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76062, tel 800-346-6873, 817-483-4422; fax 817-483-0931).
S1—3PDT miniature paddle-handle switch. If such a switch is unavailable, use DPDT switches for S1A and S1B, and an SPST switch for S1C.
S2—5-pole, 5-position rotary switch to cover all bands (see text). A suitable (and inexpensive) 5-pole, 5-position switch may be difficult to find. The Electroswitch PA 1017, available from Newark Electronics (stock number 22F817) lists for $33.74 in lots of 1-9. (Newark Electronics has many branches throughout the US; check your telephone book for a branch near you. Main office: 4801 N Ravenswood Ave, Chicago, IL 06040-4496, tel 312-784-5100; fax: 312-784-5100,ext 3107.)
T1-T3, incl—10 bifilar turns #28 or #30 enameled wire twisted 8 turns per inch on Amidon FT-23-43 cores. Note: dots indicate phasing. (Amidon Associates Inc, 2216 E Gladwick St, Dominguez Hills, CA 90220, tel 310-763-5770, fax 310-763-2250.)
T4—14 turns #28 or #30 enameled wire tapped at 10th turn from the bottom.
Pi network—C1, L1 and C2, see Table 1; tuner L2, L3 and L4, see Table 2; amplifier, see Figure 2.

Table 1

Component Values for the 90° Pi-Network of Figure 1

Band (meters)	*C1 and C2 (pF)*	*L1 (µH)*	*Number of Turns*	*Length (inches)*
160	820	8.4	44	¾
80	420	4.24	30	⅝
40	220	2.2	18	⅜
30/20	110	1.32	14	⅜
17/15/12/10	68	0.66	10	⅜

Drill and epoxy #18 wire in dowel holes. For C1 and C2, use stable 5%-tolerance capacitors (such as mica, polystyrene, etc). Inductors are made of #30 enameled wire wound on ⅜-inch-diameter dowels.

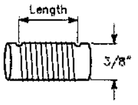


Table 2

Tuner Component Values for Figure 1

Band (meters)	*Core*	*L2 (turns)*	*L3 (turns)*	*L4 (turns)*	*Wire Gauge*
160	T-50-1	43	4	4	28
80/40	T-50-2	54	4	4	28
30/20	T-50-2	20	2	2	26 or 28
17/15/12/10	T-50-6	11	2	2	26 or 28

Wind L2 with enameled wire. Wind L3 and L4 near the ground end of L2. Use insulated wire of two colors for ease of wire identification. Core tolerances may require small changes in the number of turns on L2 to include the bands listed.

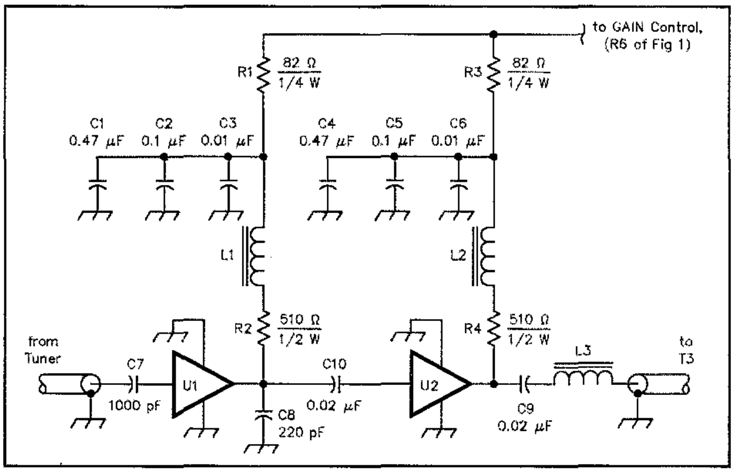


Figure 2—Schematic of the amplifier of Figure 1.

L1, L2—4 turns #28 enameled wire on Amidon FB-64-801 bead.
L3—One pass of #28 enameled wire through an Amidon FB-64-801 bead.
U1, U2—Avantec MSA 0685 MODAMPTM cascadable MMICs (available from Microwave Components, PO Box 1697, Taylor, Michigan 48180; tel 313-753-4581, evenings only.

TRANSVERTER DIN plug on the rear apron. (A jumper must be installed in the DIN plug to override disabling of the drive to the final amplifier when the DIN plug is inserted. Refer to your operator's manual.) In transceivers where this receiver/TR line is not already available, you can modify the transceiver to bring this line out to a pair of phono jacks. Or, the primary antenna line can be switched from the transceiver antenna connector to the Null Steerer primary antenna input by a relay. The relay can be operated by the transceiver's external amplifier-control relay and the Null Steerer's output can be connected to the auxiliary antenna input of the transceiver. In a separate receiver/transmitter arrangement, place the Null Steerer control relay in the receiver line from the TR relay or switch.

The Null Steerer's Amplifier

The Null Steerer's amplifier (see Figure 2) is designed to permit use of a simple single-sided PC board. You can make the resist pattern using a Radio Shack resist-ink pen (276-1530), or purchase a ready-made PC board.[4] This amplifier uses tiny monolithic microwave integrated circuit ICs (MMICs) for a low-component-count amplifier with a nominal gain of 40 dB.[5]

Construction

The Null Steerer must be enclosed in a metal cabinet to prevent hand-capacitance effects (see Figure 3). I used a 5×4×4-inch (HWD) utility box. A smaller box would suffice for a single-band version. I painted my box with Krylon flat black enamel, applied a protective Krylon matte finish and labeled the controls with DYMO tape. Suitable control knobs can be found at Radio Shack.

The most-used controls, **PHASING** pots **X** and **Y** (which actually adjust both phase and amplitude) and the **IN/OUT** switch, are placed along the bottom row (see the lead

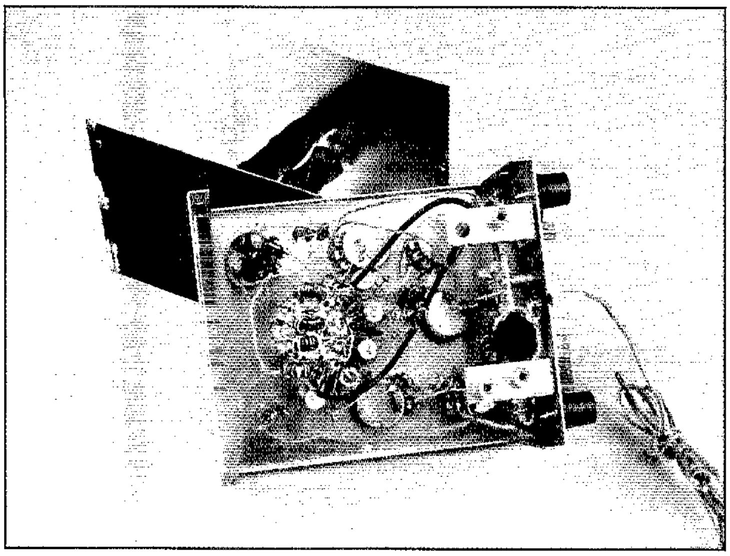


Figure 3—An inside view of the Null Steerer.

photo) with the **AMPLIFIER GAIN CONTROL** and **ATTN** pots above them and the **ATTN** pot near the **IN/OUT** switch. In my multiband version, the band switch is in the upper center of the front panel, with the **TUNE** capacitor and **IN/OUT** LED above and to each side.

Openings for the second-antenna and receiver-connection shielded phono plugs are on the rear slide-in portion of the utility box. The phono jacks are mounted in holes drilled in 1½-inch **L** brackets secured to the box bottom at the rear of the front section. The power cable enters through a grommet in a notch at the rear edge of the floor, so that after removing the phono plugs, the slide-in back portion can be removed. Place the **L** bracket for the high-signal-level receiver connection jacks at the same end as the **IN/OUT** switch and **L** bracket for the low-signal-level second-antenna connection jack at the opposite end to prevent feedback.

Parts other than those on the amplifier circuit board are mounted point-to-point on the pots and switches. T1, T2 and R3 are mounted between the **X** and **Y** pots (next to each other). High-signal-level portions of the circuit located within the boxed-in area of Figure 1 are mounted close to the **ATTN** pot and **IN/OUT** switch to avoid coupling to the low-level circuits preceding them.

In a single-band Null Steerer, mount the pi-net near the **X** and **Y** pots and mount the link-coupled tuning coil on the tuning capacitor. In a multiband Null Steerer, the pi-nets (C1, L1 and C2) and the link-in/link-out tuning coils are mounted on the **BAND** switch.

Ground the **TUNING** capacitor rotor and the **ATTN** pot's metal housing to the cabinet using soldering lugs secured beneath the capacitor and pot mounting hardware. The amplifier board is mounted on the box bottom atop a short metal spacer.

Add weight to the box and provide rubber feet for stability. Powered by 12 to 14 V dc, current drain is about 60 mA maximum and varies with the **GAIN** control setting.

Second-Antenna Considerations

Second-antenna spacing from the primary antenna of much less than about ⅛ wavelength makes null tuning critical. That's because the patterns containing nulls are confined to a smaller segment of the available 360° phasing.[6] Spacings greater than ½ wavelength result in multiple lobe and null patterns. This increases the chance of the null pattern falling on or near the desired received station when another portion of the null pattern is being used to reject an interfering noise or signal. This is a 2-element phased array, with patterns appropriate to its spacing. Consider all three dimensions, not just the zero-wave-angle patterns generally shown in handbooks (see Note 6), although they will apply to ground-wave-propagated local sources.

For nulling locally generated signals and noise, the polarization of the second antenna relative to that of the primary antenna is irrelevant, the ratio of vertical polarization to horizontal being rather stable in these fields. A short vertical base- or center-loaded[7] wire or whip antenna, similar to those used in mobile work, will suffice. Some top loading helps.

For best sky wave-propagated signal or noise nulling, the second antenna should mimic the polarization and shape (ie, vertical, horizontal, inverted **V**, etc) of the primary antenna. For horizontal or inverted **V** half-wave antennas, a low, short, loaded (see Note 7) or trap antenna, oriented parallel to the primary works best.

Do not place the second antenna where it will pick up noise *not heard on the primary antenna!* Such noise has nothing against which to null. A phono jack spliced to an appropriate RF connector permits using your SWR meter to tune second antennas.

The second antenna is in the local field of the primary antenna during transmission, but I have had no problems at a power level of 100 W. High power and close spacing may require a protective relay to disconnect the second antenna during transmit. If so, use the transceiver's amplifier control relay to operate it.

My primary vertical and horizontal antennas are located in my backyard. My short, base-loaded vertical second antennas are concealed in a bougainvillea trellis in the front courtyard. They are connected in parallel to RG-174 coax for the run to my nearby radio room. Ground is a 3-foot section of electrical conduit, thus ensuring a lossy broadband antenna. Of course, for sky wave nulling, they work best with my vertical primary antennas.

Initial Set Up and Test

With the **IN/OUT** switch set to **OUT**, the receiver should operate normally. Select a local noise or signal for testing and initial set up. Set the **GAIN** control at minimum. Center both **PHASING** pots and adjust the **ATTN** control for maximum primary antenna signal. Place the **IN/OUT** switch in the **IN** position. The selected noise or signal (and normal background noise) should be down by about 3 dB; note its level. Set the **ATTN** control for minimum primary signal. The signal should act as if the antenna feed line was shorted. Set one of the **PHASING** pots at one of its range extremes, the **GAIN** control at about 75%, and peak the signal with the **TUNE** capacitor. Set the **GAIN** control so that it somewhat exceeds the primary antenna signal level previously noted. Set the **ATTN** pot for maximum primary antenna signal. Rotate the **PHASING** pots to achieve a null. The pots interact, but a null should be found. If the null occurs near the center of both **PHASING** pot ranges, adjustment will be tedious; reduce the **GAIN** control setting. Settings are less critical when nulls occur near one (or both) **PHASING** pot extremes.

Operation

Noise or signals can be nulled by ear for best signal-to-noise-ratio (particularly in the presence of multiple noise or QRM sources), by S-meter observation, or by observing the noise or signal on a station monitor pan display. Nulling broadband noise (such as that emitted by lamp dimmers or distant thunderstorms) is often most easily done by tuning a little off the signal frequency and nulling the noise. Use minimum **GAIN** to avoid unnecessarily high amplifier noise, although that noise is usually masked by normal atmospherics.

Received signals may be stronger or weaker than the desired signal depending on their position in the lobes of their wave and azimuth angles relative to the wave and azimuth angles of the null pattern.

I have found less of a need for the Null Steerer system on 20 meters and the higher frequency bands since local noise is usually not a problem, and the constant variation of arrival angles on multi-hop sky wave signals—worsening with frequency—makes nulls less stable and effective as in all phased arrays.

Apartment dwellers, or others using indoor antennas, should find this system even more useful for the more severe noise problems of those environments throughout the HF range.

Notes

[1] Green plugs are devices into which you plug motor-driven devices (such as a refrigerator) and other equipment (such as computers and monitors) to reduce power consumption.

[2] J. Webb, "Electrical Antenna Null Steering," *QST*, Oct 1982, pp 28-32.

[3] W. Hennigan, "Broadband Hybrid Splitters and Summers," *QST*, Oct 1979, pp 44-46.

[4] PC boards for the *amplifier only* are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price: $4 plus $1.50 shipping.

A PC-board template package is available free from the ARRL. Address your request for the MICHAELS NULL STEERER TEMPLATE to: Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please enclose a business-size SASE.

[5] A. Ward, "Monolithic Microwave Integrated Circuits,"—*Part 1, QST*, Feb 1987, pp 23-29 and 32;—*Part 2, QST*, Mar 1987, pp 22-28 and 33.

[6] *ARRL Antenna Book*, 15th or 16th Ed. Array patterns, p 8-6.

[7] *ARRL Antenna Book*, 15th or 16th Ed. Chapter 16, Mobile and Maritime Antennas, p 16-1 to 16-8, and Loaded Antennas, p 6-6.

Charlie Michaels was born in Philadelphia in 1923, and licensed as W3IGR in 1939. After graduating high school, he joined the Navy V3 Reserve. After attending the Noroton Heights, Connecticut, Navy Radio School, he served aboard the USS Swan *and was at Pearl Harbor on December 7, 1941. Achieving Radioman First Class, he was appointed to the V12 Officer Training Program at the Universities of Wisconsin and Oklahoma. After WW II, Charlie earned a BSEE at the University of Pennsylvania, then did graduate studies in physics, electromagnetic radiation, ac networks and feedback theory.*

In 1948, Charlie joined the Eckert-Mauchly Computer Corporation (later called UNIVAC) as a Junior Engineer and advanced to Chief Engineer, Systems. Charlie was then Director of Engineering at Honeywell's Waltham, Massachusetts, Lab, held positions with GE Computer Operations in Phoenix, Arizona, and Xerox Corporation Corporate Staff before retiring.

An active amateur since high school, Charlie is an ARRL Life Member and Member since 1943. His other hobbies include amateur astronomy, old western, railroad and hillbilly ballads with guitar, noodling on the electronic keyboard, trout fishing and camping in his radio-equipped travel trailer. QST

W1AW schedule

Pacific	Mtn	Cent	East	Sun	Mon	Tue	Wed	Thu	Fri	Sat
6 am	7 am	8 am	9 am			Fast Code	Slow Code	Fast Code	Slow Code	
7 am	8 am	9 am	10 am			Code Bulletin				
8 am	9 am	10 am	11 am			Teleprinter Bulletin				
9 am	10 am	11 am	noon							
10 am	11 am	noon	1 pm		Visiting Operator Time					
11 am	noon	1 pm	2 pm							
noon	1 pm	2 pm	3 pm							
1 pm	2 pm	3 pm	4 pm	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code
2 pm	3 pm	4 pm	5 pm	Code Bulletin						
3 pm	4 pm	5 pm	6 pm	Teleprinter Bulletin						
4 pm	5 pm	6 pm	7 pm	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code
5 pm	6 pm	7 pm	8 pm	Code Bulletin						
6 pm	7 pm	8 pm	9 pm	Teleprinter Bulletin						
6^{45} pm	7^{45} pm	8^{45} pm	9^{45} pm	Voice Bulletin						
7 pm	8 pm	9 pm	10 pm	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code	Fast Code	Slow Code
8 pm	9 pm	10 pm	11 pm	Code Bulletin						
9 pm	10 pm	11 pm	Mdnte	Teleprinter Bulletin						
9^{45} pm	10^{45} pm	11^{45} pm	12^{45} am	Voice Bulletin						

Note: W1AW's schedule is at the same local time throughout the year. The schedule according to your local time will change if your local time does not have seasonal adjustments that are made at the same time as North American time changes between standard time and daylight time. From the first Sunday in April to the last Sunday in October, UTC = Eastern Time + 4 hours. For the rest of the year, UTC = Eastern Time + 5 hours.

☐ ***Morse code transmissions:***

Frequencies are 1.818, 3.5815, 7.0475, 14.0475, 18.0975, 21.0675, 28.0675 and 147.555 MHz.

Slow Code = practice sent at 5, 7½, 10, 13 and 15 wpm.

Fast Code = practice sent at 35, 30, 25, 20, 15, 13 and 10 wpm.

Code practice text is from the pages of *QST*. The source is given at the beginning of each practice session and alternate speeds within each session. For example, "Text is from July 1992 *QST*, pages 9 and 81," indicates that the plain text is from the article on page 9 and mixed number/letter groups are from page 81.

Code bulletins are sent at 18 wpm.

☐ ***Teleprinter transmissions:***

Frequencies are 3.625, 7.095, 14.095, 18.1025, 21.095, 28.095 and 147.555 MHz.

Bulletins are send at 45.45-baud Baudot and 100-baud AMTOR, FEC Mode B. 110-baud ASCII will be sent only as time allows.

On Tuesdays and Saturdays at 6:30 PM Eastern Time, Keplerian elements for many amateur satellites are sent on the regular teleprinter frequencies.

☐ ***Voice transmissions:***

Frequencies are 3.99, 7.29, 14.29, 18.16, 21.39, 28.59 and 147.555 MHz.

☐ ***Miscellanea:***

On Fridays, UTC, a DX bulletin replaces the regular bulletins.

W1AW is open to visitors during normal operating hours: from 1 PM until 1 AM on Mondays, 9 AM until 1 AM Tuesday through Friday, from 1 PM to 1 AM on Saturdays, and from 3:30 PM to 1 AM on Sundays. FCC licensed amateurs may operate the station from 1 to 4 PM Monday through Saturday. Be sure to bring your current FCC amateur license or a photocopy.

In a communications emergency, monitor W1AW for special bulletins as follows: voice on the hour, teleprinter at 15 minutes past the hour, and CW on the half hour.

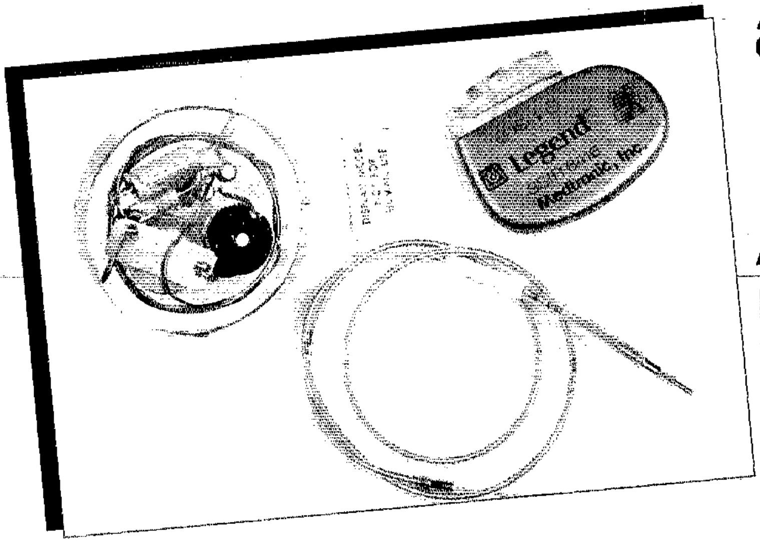
Headquarters and W1AW are closed on New Year's Day, President's Day, Good Friday, Memorial Day, Independence Day, Labor Day, Thanksgiving and the following Friday, and Christmas Day. On the first Thursday of September, Headquarters and W1AW will be closed during the afternoon.

Pacemakers, Interference and Amateur Radio

Can hams who use pacemakers safely pursue Amateur Radio? And if they can, what safety precautions should they take?

By Fred Weber, MD, AA2KI
145 E Atlantic Blvd
Ocean City, NJ 08226

Photo by Kirk Kleinschmidt, NT0Z



Pacemakers—old and new. The large round pacemaker—the one that looks as though it were made in someone's home electronics shop—is a very early model. The smaller unit is representative of today's compact pacemakers. Note the coiled lead.

It's safe to assume that hams who come home from the hospital with shiny new pacemakers implanted in their chests are at least a little bit worried about turning on their rigs. It doesn't matter how much they've been counseled by their surgeons, or how much literature they've read. All of that stuff was about *someone else*! They're turning on their radios now, and they want to know that they're not jeopardizing their lives by pushing the PTT or twitching the key.

Although pacemaker experts and various informal studies and user histories agree that normal Amateur Radio environments pose minimal danger to pacemaker users, it's a good idea for *everyone* to reduce their exposure to high-level RF and electric fields.

Before we explore specific safety recommendations for hams who have pacemakers, let's take a look at pacemakers themselves, and how they work.

Cardiac pacemakers—lifesaving devices used to increase or regulate heartbeat—were introduced in 1957, and many active hams have these devices in place. By understanding their principles, safe station operation can be assured.

In this article, we'll look at pacemaker hardware, function, the ways pacemakers can fail, history, interference problems and how pacemakers interact with devices and situations in our modern world.

How Pacemakers Work

The response of muscle to electrical stimulation is well known. In 1768, Galvani, using a voltaic pile, described frog muscle contraction in response to an applied voltage. Pacemakers use this method to induce heartbeats in the human body.

The pacemaker's transmit mode is called pacing, and the receive mode is called sensing. In the pacing mode, the pacemaker sends a signal to the heart. In the sensing mode, the pacemaker receives a signal from the heart.

The pacing mode results in the contraction of the heart muscle. During the sensing (listening) mode, the pacemaker receives signals indicating whether a normal heartbeat has occurred (that is, whether the patient's heart provided its own properly timed beat).

After the unit hears a normal heartbeat (and after a short blanking period), the pacemaker is switched again to receive mode. If another heartbeat is heard, the system is cycled and the pacemaker begins to listen again. The pacemaker listens to every heartbeat.

If no heartbeat is detected at the end of a set interval, the unit is switched to transmit mode, which sends a 0.5-ms, 5-V pulse to the heart. The heart muscle contracts and a heartbeat is formed. The pacemaker again switches to the receive mode and listens. It is during sensing (receive mode) that the system is most vulnerable to RF interference.

How Pacemakers are "Installed"

During the implantation surgery, an electrical connection is made between the pacemaker and the heart. The pacemaker lead is inserted into the heart through a small vein in the upper chest.

After the lead and the pacemaker are firmly in place, the electrical connection to the heart is analyzed. First, the threshold of stimulation (in volts) is found (the level required to cause the heart muscle to contract) and current flow at that voltage is measured. Typical values are 0.5 V at 1 mA.

Finally, the pacemaker's sensing characteristics are examined. The voltage generated by a normal heartbeat is measured; it must be above 6 mV to assure proper pacemaker operation.

After a final hardware check, the incision is closed.

Possible Problems

There are many ways for pacemakers to fail, the most common being dead batteries! The lithium-iodine batteries used provide about 1 Ah of capacity (normal battery life is seven years).

The insulation on the lead running from the pacemaker to the heart may fail, creating a short circuit. If this happens, the pacemaker may operate intermittently, or the short circuit may drain the unit's battery.

Because the heart is in constant motion, the lead may become dislodged or even poke through the heart muscle. This can cause hiccoughs (at the paced rate!) or other involuntary muscle contractions.

In some patients, scar tissue builds up at the lead tip, increasing the circuit's resistance and the stimulation threshold.

Pacemakers are also susceptible to interference from electric and RF fields. As hams we know that subjecting a wire to an electromagnetic field causes a current to flow in the wire. This is how radio receiving antennas work. The pacemaker lead is such a wire, and it can act as an antenna, receiving unwanted electromagnetic signals that can potentially interfere with normal operation.

Smaller induced currents may not be large enough to cause stimulation or damage, but they may be strong enough to fool the pacemaker into thinking it's heard a normal heartbeat. Thinking that a normal heartbeat has occurred, the pacemaker produces no pacing pulse.

Ham Radio Test Case

Ten days after I implanted a pacemaker in Charles Gilbert, W3YJM, a continuous 24-hour recording of his heart's activity was made as he went about his normal daily activities. Three on-air operating sessions (in his shack) were recorded.

Charles fed 100 W to several antennas on several bands. The first session involved contacts on 20 meters using a half-wave dipole. Later, operation was switched to 40 meters, also using a dipole. Finally, Charles keyed up on 80 meters using an inverted-V antenna.

No discernible effect on pacemaker activity was noted.

Although this study was not comprehensive, nor did it have scientific "controls," it does illustrate that hams who have pacemakers can expect to safely use their stations.

—*AA2KI*

As long as the interfering signal exists and continues to blank the pacemaker's "receiver," no pacing pulses will be sent to the heart muscle. Blackout and death can occur.

In addition to magnetically coupled interference (discussed above), interference can enter the body and influence pacemakers via galvanic coupling—when a voltage or current source is applied to the body and current flows through all or part of it.

The History of Pacemakers

Throughout history, physicians have experimented with applying voltage and passing current through the human body. In 1788, Kite reported an attempt at reviving a patient with dc voltage. This is the same principle as modern defibrillation.

In 1925, the first successful treatment using a pacemaker took place in Australia. A stillborn infant was resuscitated and was alive and well four years later.

The first adult application of pacemaker technology was made by a New York cardiologist who built a spring-wound, "three-speed" pacing device (it was good for six minutes). Informally, the device was reported as successful, but resistance to invasive medical treatments and a cruel stroke of fortune doomed the device.

In 1931, a few months before the announcement of the new mechanical pacer, the movie *Frankenstein* was released. Because this science fiction horror also featured electrical resuscitation, unfortunate comparisons were made and the unit never came into general use.

In 1952, doctors achieved control of cardiac rhythm with electrodes placed on the front and back of the chest. Because the voltage necessary for stimulation was between 30 and 150 V, it was quite uncomfortable and not suitable for long-term use.

In 1958, the first self-contained pacemaker was implanted by surgeons in Stockholm, Sweden. The implant lasted three hours, was replaced, and thereafter lasted eight days. The patient survived and did not require further pacing.

This initial approach required major surgery; an alternative was developed in the US that placed the lead inside the heart without major surgery and major risk. This is how pacemakers are implanted today.

Hazards

These wonderful devices were, however, not free of risk. Reports of unwanted interactions between pacemakers and electrical devices began to appear.

Less than five years after the first self-contained pacemaker was implanted, an electric shock from poorly grounded hospital equipment disrupted pacemaker function in another patient and caused abnormal heart rhythms. The patient died.

Shortly thereafter, RF interference was encountered. Diathermy—an obsolete RF tissue-heating treatment that has largely been replaced by ultrasound—provided a graphic illustration.

Diathermy machines used RF transmitters that were connected to hand-held radiators that were placed over the area of the body requiring treatment. The transmitters were tuned to various high frequencies—popular bands included 13.5 and 27 MHz. The tissues under the antenna were warmed. Severe pacemaker interference often resulted, however, and the use of diathermy in pacemaker patients was quickly discontinued.

By 1968, interference problems had become significant. The first step toward alleviating interference problems came via techniques used to place pacemaker leads during implantation. By placing the electrodes closer together in the heart, interference was dramatically reduced.

The second change was in pacemaker software. After pacemakers reportedly malfunctioned near active microwave ovens, the logic circuits were redesigned to differentiate between discrete and continuous interference. With this new ability, pacemakers could revert to fixed-rate pacing if interference was detected. The problem of pacemaker function near poorly shielded microwave ovens was overcome.

There are many potential hazards for pacemaker patients in our electrical world. Many devices encountered in daily life produce electric and magnetic fields. Here are some examples:

Our houses are rich in electric and magnetic fields, which vary from day to night, possibly reflecting appliance use. The fields from household electrical appliances, however, don't disturb pacemaker function.

In related tests, the electric field strengths in aircraft (at HF, VHF and microwaves) have also been measured. Patients experienced no change in pacemaker function.

A new antitheft technology known as electronic article surveillance (EAS) is in general use in many stores today. A tag is placed on the protected article, and devices near the store exits "scan" for the tags with RF or magnetic sensors. RF EAS systems didn't influence pacemaker wearers, but magnetic systems caused interference and pacemaker malfunction in two patients.

Weapons-detector gates commonly seen at airport security checkpoints had no effect on pacemaker function.

Work environments may produce significant amounts of electromagnetic interference. Recently, several potentially hazardous areas were studied. These extremes aren't usually encountered by most pacemaker patients (see Table 1).

Pacemaker failure as a result of radiation therapy for breast cancer has also been reported. At the end of therapy, the pacemaker was behaving irregularly. The unit was replaced and examined, and the doctors found that one of the unit's IC chips had been damaged.

Probably the most hostile environment for pacemaker patients is found within magnetic-resonance imaging (MRI) scanners, which are now in common use. Today's MRI devices produce extremely strong magnetic and RF fields. Needless to say, this situation is very dangerous, and the benefits of scanning must be carefully weighed against risks to the patient. An MRI scanner produces fields that are many times greater than those found in even high-power ham stations.

Tips for Safe Operation

When estimating potential dangers to pacemaker users, take steps not to generalize about pacemaker models, configurations, lead systems, and so on. Each patient is unique, and all variables must be carefully evaluated. Safe operation of a particular unit in a particular environment does not guarantee safe operation of that device in another.

So, what about hams who have pacemakers? Can they safely operate their stations without worry? The answer is straightforward: As long as accepted safety practices are maintained—the same practices recommended for all hams when it comes to electrical and RF safety—there is no increased danger. As the sidebar illustrates, interference to hams who use pacemakers (in Amateur Radio environments) is not expected with modern pacemakers.

Table 1
Pacemaker EMI Source Characteristics

EMI Source	*Output (max value)*
Spot-weld machine	1500 A
Arc welding machine	225 A
Industrial welder	300 A
Submerged arc welder	1000 A
TIG welder	220 A
Neon sign test room	4000 V
Electrical substation	138 kV
CRT assembly area	Magnetic field
Degaussing coil	Magnetic field
Jet engine plant	Magnetic field

Table values are from D. Marco, *PACE*, 15:2020 (1992).

Excellent discussions of RF exposure and recommended safety practices are found in the *ARRL Antenna Book* and in the *ARRL Handbook*.

One final precaution involves antennas. Most hams use external antennas that limit their exposure to RF energy. Everyone is encouraged to do this—especially hams with pacemakers. The increasing use of indoor loops and attic wires, however, brings RF closer to the shack, sometimes even bathing it in RF. As a precaution, hams with pacemakers should avoid these types of antennas.

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Strays

Shuttle Successes

The Shuttle Amateur Radio EXperiment (SAREX), once again thrilled thousands of hams during the April STS-59 mission, operated by Mission Specialist Jay Apt, N5QWL, and Payload Commander Linda Godwin, N5RAX. The primary payload, a three-frequency imaging radar, mapped over 70 million square kilometers of the Earth's surface—representing 6,480 gigabytes of data about soil erosion, deforestation, volcanism and tectonic activity, among other things. Crew members used the SAREX payload to make 1,674 packet contacts, dozens of voice QSOs, and to speak with students from nine schools.

Jay Apt said, "It was a real thrill for me to be able to operate from low Earth orbit for the third time. It really makes me feel connected with the people I'm flying over to work them on two meters. My biggest thrill was talking to my old crew mate Ken Cameron, R3/KB5AWP, in Star City, Russia, where he and cosmonauts talked to us with a 5-9 signal."

"Talking to hams fit right in with our purpose for being in space," reported Linda Godwin. "Our project, *Mission to Planet Earth*, was to study people and science."—*Rosalie White, WA1STO*

144-MHz Sporadic E

Every year, alert 2-meter operators make dozens of sporadic-E contacts up to 2300 km distant—and sometimes farther—but the openings are elusive. Analysis of 18 years of 144-MHz sporadic E suggests how you can maximize your chances of catching a great 2-meter opening.

By Emil Pocock, W3EP
625 Exeter Rd
Lebanon, CT 06249

Sporadic E (E_S) is best known on the 6-meter band, where it can provide hours of single- and multi-hop contacts from a few hundred to several thousand kilometers distant. E-skip signals are usually strong, even for modest stations. Sporadic E appears most often during summer mornings and late afternoons, but it can occur unpredictably at almost any time or season. These characteristics make it one of the most popular and fun modes of VHF propagation.

The cause of sporadic E is not fully understood, but it appears when especially dense concentrations of E-layer ions form into thin sheets, or clouds, at about 100 km altitude. These sporadic-E clouds may be as small as 10 km in diameter. Most E-skip openings are localized and often short-lived, but they can remain stable for hours. Sporadic E exhibits a maximum usable frequency (MUF) similar to the more familiar F layer, and on rare occasions, the sporadic-E MUF has exceeded 200 MHz.

Two-meter operators can also use sporadic-E propagation to make DX contacts. Openings occur only about a tenth as often as on 6 meters and generally do not last as long. Signals are usually loud, and typical contacts are in the 1500- to 2300-km range, with longer jumps being more common. Avid VHF DXers usually use SSB and CW to make sporadic-E contacts, but FM works nearly as well.

You don't need special equipment to work 2-meter sporadic E, but you do need to know when the band is likely to be open. One technique involves carefully monitoring the 6-meter band, TV channels or the FM broadcast band for telltale signs that the MUF is high. Another clue is knowing the days and times when 2-meter sporadic E is most likely to occur. This article reveals for the first time when 2 meters is likely to be open for sporadic-E DX.

More information about sporadic E can be found in the propagation chapter of the *ARRL Handbook* and in my article "Sporadic-E Propagation at VHF: A Review of Progress and Prospects" in April 1988 *QST.—W3EP*

Hams first stumbled upon sporadic-E (E_S) propagation on the old 5-meter band (at 56 MHz) during the summer of 1935. For half a dozen years, the annual summer E-skip season was the rage among VHF experimenters. Several articles appeared in *QST* during the late 1930s that analyzed E-skip events and offered explanations for the sudden appearance of strong signals out to 2300 km—well beyond the 150-km range the 5-meter gang expected.[1] The wartime ban on Amateur Radio interrupted these investigations, but amateurs were eager to try out the new postwar VHF bands at 50 and 144 MHz when they were made available in late 1945.

Sure enough, sporadic E was rediscovered on the new 6-meter band during the summer of 1946, but there was little expectation that it would ever reach 144 MHz. That was just too high for any form of ionospheric propagation! Scattered reports of E-skip signals on the new 2-meter band during the late 1940s were greeted with understandable skepticism. Skeptics were quieted when W5VY (Texas) and W8WXV (Ohio) reported a June 23, 1950, contact that could only be attributed to sporadic E. Operators became more alert to the possibilities. Even so, 2-meter E-skip contacts continued to be rare through the 1950s and into the 1960s.

That has changed during the past 20 years. With more sophisticated equipment and alert operating practices, VHF operators now report at least a dozen 2-meter sporadic-E openings each year across the US and southern Canada, almost all from May through August.[2] Booming 2-meter contacts in the 1500- to 2500-km range are no longer isolated events, but they're still uncommon enough to make news when-ever they occur. In spite of advances, we probably miss many more openings because it's still uncertain exactly when sporadic E is most likely to reach as high as 144 MHz.

Use of Historic Data

Previous studies have shown that most 50- and 88-MHz sporadic-E openings occur during the daylight and early evening hours between May and August, but it's not certain whether these are also the optimal times at 144 MHz. More data is needed to answer this and other questions, but apparently no one has kept systematic records of 144-MHz sporadic-E activity over a long period of time (comparable to the nearly 20-year record Patrick Dyer, WA5IYX, has accumulated for sporadic E on 50 and 88 MHz from Texas).[3]

Obtaining the necessary data is possible even without systematic observations, because 2-meter openings do not go totally unrecorded. They are unusual enough that they have been reported for years in *QST* and in other publications. Most VHFers also retain their logs.[4] These sources can provide the needed long-term data.

The study reported herein is based on a compilation of all the 144-MHz sporadic-E reports that could be found from 1976 to 1993 in *QST* and six VHF newsletters; additional sources included some 24 station logs. Care was taken to weed out contacts made via tropospheric ducting, meteor scatter, aurora and other forms of propagation that could have been confused with sporadic E. Single contacts were generally eliminated as well, because of the uncertainty of the propagating mode.

The resulting data includes thousands of 144-MHz sporadic-E contacts, most between 1000 and 2500 km. The midpoints of all contacts were plotted to determine the probable locations of the sporadic-E regions responsible for creating the paths. They were then grouped into 209 distinct events (or openings) that took place on 161 different dates. An event was defined as a continuous series of contacts that could be attributed to the same sporadic-E reflecting region. The beginning and ending times of each opening were compiled in local time, corresponding to the location of the responsible sporadic-E region.[5]

Two-meter sporadic-E openings lasted from just a few minutes to more than eight hours. An unusual 8-hour opening occurred on July 11, 1982, involving a wide area from

[1] Notes appear on page 41.

Table 1

Cumulative Number of Days per Month with 144-MHz Sporadic E, 1976-93

Month	Jan	Feb	Mar	Apr	May	Jun	Jul	Aug	Sep	Oct	Nov	Dec
Days	2	3	0	1	30	53	51	12	3	0	1	4

the Canadian Maritimes, across much of the Midwest and upper South, to as far west as Wyoming.[6] The average of all recorded 144-MHz openings was just more than 90 minutes. This is probably on the high side, because briefer and less spectacular openings were more likely to be overlooked and remain unreported. Indeed, this data undoubtedly undercounts actual 144-MHz E-skip openings. Some observed events were not recorded in available sources, while others probably went undetected.

Relying on data derived from reported contacts has inherent problems relating to operating habits. Few VHFers are on the air between midnight and 7 AM (local time), for example, so any 144-MHz E-skip openings that might appear during that span would likely go unnoticed. Most operators work during weekdays, thus it would be expected that proportionately more Saturday, Sunday and evening events would appear in the record. VHFers are also more alert for E-skip during the summer months, a factor that might also subtly skew the data. Finally, geography also plays a role, because openings that affect low-density population areas are more likely to go unnoticed as well.

Rigorous statistical analysis of 144-MHz E_S data is therefore compromised by inconsistent reporting and inherent biases. Nevertheless, this compilation is probably the best source of 2-meter E-skip information that can be obtained over so long a period and over so wide an area. Comparisons with similar data collected from European sources (1977-92) and 88-MHz data from Texas (1980-90) provide additional opportunities to support some general conclusions about the appearance of sporadic E at 144 MHz.

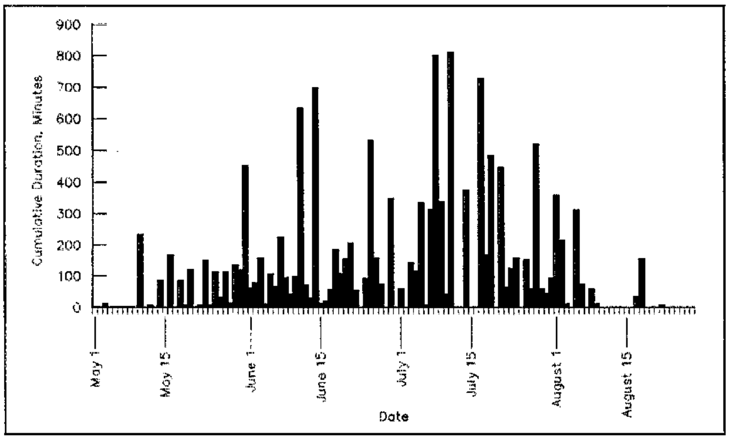

Figure 1—Cumulative minutes of daily 144-MHz sporadic E, May to August, 1976 to 1993. The data appears to be trending toward a bell curve, with a peak around July 1. Compare this graph with that in Figure 2.

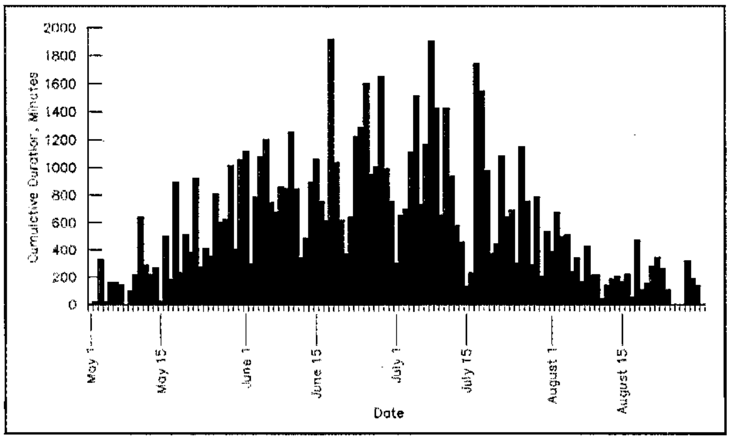

Figure 2—Cumulative minutes of daily 88-MHz sporadic E, May to August, 1980 to 1990. A bell curve distribution, with deep nulls, appears to peak around July 1. Source: Pocock and Dyer (1992).

Monthly Occurrence

Sporadic E at 144 MHz is most common from May through August, as shown in Table 1. This is consistent with what is known about 50- and 88-MHz E-skip distribution. Although 147 of the days (91%) occurred in the May to August peak season, 144-MHz E_S occurs almost every month, but it's least likely during the weeks surrounding the spring and fall equinoxes.

A daily analysis of the four-month peak summer season, summarized in Figure 1, shows a familiar sporadic-E distribution.[7] The cumulative minutes of 144-MHz sporadic-E dates reveal an emergent bell-curve distribution centered around July 1. This pattern closely resembles 88-MHz data for 1980-90, shown in Figure 2, and suggests that the ultimate distributions might be similar.

Analysis of 88-MHz sporadic-E data from Texas uncovered a five-day cycle of date-specific peaks and troughs in E-skip activity. Efforts to detect this same unusual cycle in the 144-MHz data set are not conclusive. Figure 3, which superimposes the 144-MHz cumulations on top of those for 88-MHz, indicates some correspondence of peaks and troughs, but the variation is great enough to cast doubt on the relationship. The much wider geographic scope of the 144-MHz data may invalidate the comparison, but some coincidence could be expected in any case. More than half of 144-MHz events that fell within the capture area of the Texas observations, for instance, occurred on designated 88-MHz peak dates.[8]

Daily and Hourly Occurrence

A breakdown of reported E_S occurrence by days of the week reveals a distribution that favors Saturday and Sunday. This undoubtedly results from data reported primarily by operators who aren't at their stations during workdays. If the data was not biased in this way, E-skip could be expected about equally as often on each day of the week. Table 2 shows that this is not the case. Slightly more than 38% of all 144-MHz E_S days (56 of the 147) were reported on weekends, somewhat greater than the 29% share that ordinarily could be expected for Saturday and Sunday alone.

The 144-MHz data for May-August also yields the familiar double peaks at about 1000 and 1900 local time daily. This is

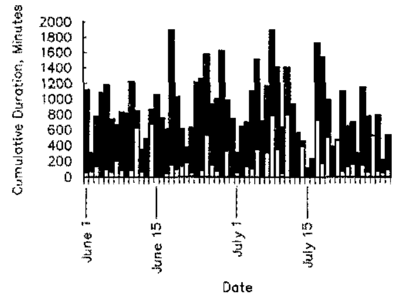

Figure 3—Cumulative minutes of daily 144-MHz sporadic E (light bars) superimposed on cumulative minutes of daily 88-MHz sporadic E (dark bars). The June to July period covers the same years as the Figure 1 data. Six of the nine 144-MHz peaks over 400 minutes correspond to 88-MHz peaks; fewer than half of those over 1000 minutes at 88-MHz correspond to apparent 144-MHz peaks.

Table 2

Distribution of 144-MHz Sporadic-E Days by Day of the Week, May to August, 1976 to 1993

Day of the Week	Mon	Tue	Wed	Thu	Fri	Sat	Sun
Total Days	317	318	319	316	315	313	316
E_S Days	18	19	19	13	19	23	33
Percent with E_S	5.8	6.0	6.0	4.1	6.0	7.3	10.4

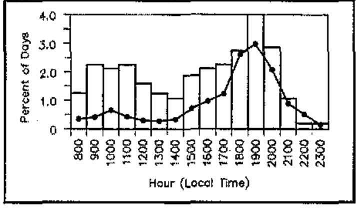

Figure 4—Percent of Saturday to Sunday hours (bars) and weekday hours (points and line) with reported 144-MHz sporadic E, May to August, 1976 to 92.

Table 3

American and European 144-MHz Sporadic-E Occurrence on Successive Days, May to August, 1977 to 1992

	US/Canada	*Europe*
Total Days	1968	1968
Days with E_S	139	182
Successive E_S Days	54	97
Portion of E_S Days in Succession	38.8%	53.3%

shown in Figure 4, which also reveals a significant difference between weekdays and weekends. That difference is not real (there is nothing to indicate that sporadic E is actually more common on Saturdays and Sundays)—it's another artifact of on-the-air habits. Operators are more likely to be at their stations during the mornings on weekends than during the week; indeed, they are likely to spend more time in general at their stations on Saturdays and Sundays than on other days of the week.

As a result, Saturday-Sunday data undoubtedly provides a better indicator of actual daily and hourly distributions. Indications are that 144-MHz E-skip can be observed on as many as 10% of the days during the May to August season somewhere in the US and Canada. This suggests that as many as half the 144-MHz sporadic-E events that actually occur during the rest of the week go unnoticed or at least unreported. Sporadic E as high as 144 MHz is also somewhat more likely to occur at 1900 local time (4% of the days) than at 1000 (less than 2.5% of the days). Most missed openings probably occur on weekday mornings.

European 144-MHz Data

The English *VHF-UHF DXer* newsletter published a similar set of 144-MHz sporadic-E data covering the May-August seasons from 1977 to 1992.[9] It lists the dates when sporadic E was reported anywhere in Europe in three categories of duration, ranging from "short with few reports" to "widespread openings lasting more than an hour." This compilation probably has the same inherent weaknesses as the American data, with an additional problem in that durations were not precise and time periods were omitted. Nevertheless, the European data provides an opportunity to compare key features of 144-MHz E_S between two continents of comparable size and geographic latitude.

Europeans reported 182 days of 144-MHz E-skip during May through August over 16 years (1977-92), for an average of 11.4 days per season. This is slightly higher than the 139 days for the same period in the US, for an average of 8.7 days per season. The higher European figure may reflect a greater VHF population density and more efficient reporting—or it might signify an actual difference in the occurrence of sporadic E. Neither data set is reliable enough to make either conclusion. The distribution of European E-skip over the summer season is similar to the American data, except that there are more pronounced peaks around June 5 and July 10.

A direct comparison of European and American E_S dates for May through August, 1977 through 1992, yields just 18 occasions when 144-MHz E-skip appeared on both continents on the same days. That works out to 12.9% of the American E-skip dates and 10.2% of those in Europe. No other intercontinental associations were evident, such as sporadic E appearing in the US a day before or after its appearance in Europe. These results rule out using 144-MHz sporadic-E openings in Europe as a forecast for what might happen in the US. They also suggest that the causes of VHF sporadic E are probably more dependent on local circumstances than on global conditions.

Consecutive Dates

American and European data reveal that 144-MHz sporadic E tends to occur on successive days. This is tabulated briefly in Table 3. More than a third of the American 144-MHz E-skip days and more than half of those in Europe appeared in successions of two or more consecutive days. The longest string of consecutive days in the US occurred during the five days from July 5 to 9, 1989, primarily over the eastern half of the country. Consecutive sporadic-E days do not necessarily mean that the same regions were active each day, although that was often the case.

These strings of consecutive E-skip days occur more often than would be expected by chance, but it's difficult to draw further conclusions about why this happens. Many of the reported consecutive E-skip days are the result of operating psychology. When VHF operators learn of a 144-MHz E_S event, they are much more likely to pay closer attention to the band on the following day. Their attentiveness is often rewarded with additional E-skip that otherwise might have gone unnoticed. It's possible that 144-MHz sporadic E is simply more apt to be observed on days immediately following earlier reports. Nevertheless, this clumping of E_S days can't be dismissed altogether as another result of operating habits. Despite the potential bias, the data may be reflecting an actual sporadic-E characteristic.

A Five- to Six-Year Cycle

An analysis of sporadic-E data from 1949 to 1986, including 88-MHz data from Texas, suggested that there may be a regular five- to six-year cycle of annual occurrence in sync with the solar cycle.[10] The European and American 144-MHz data provides some additional support for this observation. The lines in Figure 5 result from calculating annual accumulations (minutes of observed E-skip each year) as a percentage of total minutes of E-skip for the entire multiyear series. This normalizes the data so that the three different sources, Texas 88-MHz,[11] American 144-MHz, and European 144-MHz data can be compared directly.

The composite results from these three independent sources show that periods of low cumulative sporadic-E activity—the 1979 and 1984-85 troughs—are separated by five or six years. Similarly, the periods of high cumulative activity—the 1977-78, 1982-83 and 1987 peaks—are also separated by five- or six-year intervals. Peaks seem to correlate roughly with the rising and declining slopes of the solar cycle, as measured by the 2800-MHz solar flux, and thus demonstrate a degree of synchrony with solar activity. The possible significance of the observation (if valid) is unknown, but it might prove a reliable indicator for forecasting E-skip activity.

Geographic Distribution

The midpaths of all May to August sporadic-E events, and therefore the presumed active E-layer regions, were plotted on a single map. The result was a very crowded jumble of dots and arrows indicating the movement of E-skip activity centers, but certain general patterns emerged. More than 90% of all 144-MHz E_S centers fell within one of the four outlined regions shown in abstracted form in Figure 6.

Region C was by far the most crowded, followed by regions B, D, and A. Much of the distribution can be accounted for in terms of geography of population and path geometry. Active sporadic-E centers that fell in region C, for example, formed the midpaths of contacts between the populated East Coast and the Midwest. The high population density in these two areas could account for much of the relatively higher occurrence of sporadic-E centers from Ohio and Indiana south to the Gulf Coast.

Similarly, the lower concentrations in regions A and B can be accounted for by the relatively smaller population densities in the Midwest and West, which provided the endpoints for E-skip paths that crossed these two regions. Region D centers supported sporadic-E paths that primarily lay parallel to the coast, especially between the Northeast and the deep South.

No centers appeared along the edges of the country—northern New England, Florida, southern Texas, and the Pacific Northwest. These areas do not provide many useful midpoints between two areas of population at the distances observed most often for 144-MHz E-skip. The near absence of centers in California and Nevada, at latitudes similar to the active Midwest region, suggests low station activity and reporting levels.

Other features of the map may reflect actual distribution of sporadic E. Professional studies at lower frequencies show that sporadic E is twice as common at southern latitudes than more northerly ones across the US.[12] Thus the relative paucity of E_S centers across the northern states and southern Canada may not be due to sparse reports and low adjacent population densities, but may reflect an actual sporadic-E distribution. By far the highest concentration of 144-MHz E-skip activity is centered on the southern half of region C. The combination of southerly latitude and relatively high adjacent population densities may have created this apparent E-skip hot spot.

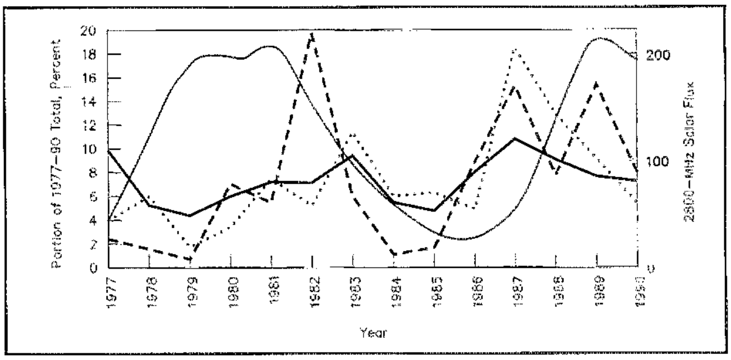

Figure 5—Annual portion of sporadic E reported for three series of data: 88-MHz Texas data (line), 144-MHz US data (dashed line), and 144-MHz European data (dotted line). The light curved line represents the 2800-MHz solar flux, superimposed to reference the solar cycle.

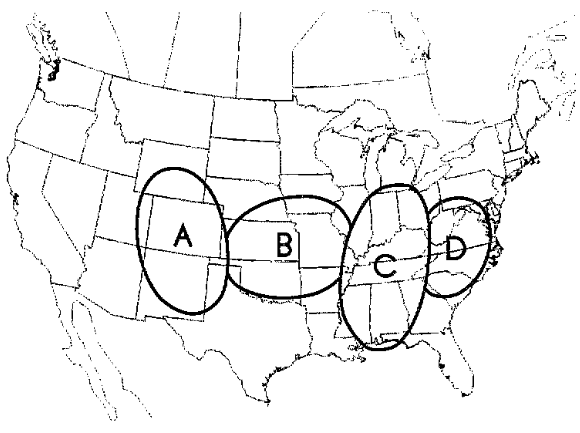

Figure 6—More than 90% of all 144-MHz sporadic-E path midpoints fell within these four regions. Region C had the highest density of midpoints.

Conclusions

Sporadic E in the states appears much more frequently at 144 MHz than is commonly observed and reported. Analysis suggests that 144-MHz E-skip occurs somewhere in the US on at least 10% of all days during May through August. The relative occurrence is even higher during the three-month core period between May 15 and August 15. Nevertheless, the chances that any particular location would observe 144-MHz sporadic-E propagation during the summer season is smaller than one day in ten. It may be somewhat more common across the southeastern quadrant of the US than elsewhere in the country.

Although 144-MHz E_S has been observed throughout the daylight and early evening hours, it is somewhat more likely to form around 1000 and 1900 local time. The evening peak appears to be somewhat stronger. Prime on-the-air times may vary by one hour earlier or later, because these times indicate when actual sporadic-E midpoints are active. Stations in adjacent time zones to the east or west of active regions may have to adjust the times to correspond to path midpoints.

Sporadic E at 144 MHz can take place in any month of the year, but more than 90% of all reported 144-MHz E-skip occurred in the four-month May to August season. There are comparatively few events from September to April. The December to January miniseason, also observed at 50 and 88 MHz, provides the best opportunity for E-skip during the eight-month lull. Sporadic E is rarest during March through April and September through October. In spite of the documented distribution, many sporadic-E events are probably overlooked during September to April because expectations are low.

There may be a tendency for VHF sporadic E to appear on consecutive days, often in the same area. There are also indications that there is a five- or six-year cycle of sporadic-E activity that coincides with the 11-year solar cycle. E-skip activity appears to peak during the rising and falling portions of the cycle.

Further long-term studies, especially based on systematic observations, may clarify many of the indications apparent in the 144-MHz data analyzed here. For the practical operator, the patterns that emerged from historic analysis may aid the discovery of elusive 144-MHz sporadic-E openings in the future. Yet too great a reliance on past experience could also mean that much E-skip may remain hidden from view.

Notes

[1] See Pocock, ed., *Beyond Line of Sight*, Ch 2; "Sporadic E," for a short historical account of sporadic-E propagation at VHF and reprints of three *QST* articles from the 1930s.

[2] E. Pocock, "Sporadic E at VHF: A Review of Progress and Prospects," *QST*, Apr 1988, pp 33-39.

[3] See, for example, P. Dyer, "A Seven-Year Study of 50-MHz Sporadic-E Propagation," *CQ*, Aug 1972; E. Pocock and P. Dyer, "Eleven Years of Sporadic E," *QST*, Mar 1972, pp 23-28.

[4] The sources of 144-MHz sporadic-E events are *Anomalous Propagation*, *Northeast VHF News*, *SWOT Bulletin*, *QST*, *Upper Midwest VHF/UHF Newsletter*, *VHF+*, *VHF-UHF DXer [England]; West Coast VHFer*, and K5SW, K7ICW, K8AXU, K9AKS, KA2WKA, KB0ZQ, KE0YG, N4MW, N4VC, N0BSA, N0HJZ, N0HRF, N0LL, VE6TA, W1AJR, W2BOC, W2MGF, W3EP, W3ORD, W4AZR, W4FSO, W5SXD, W5UWB, WA5IYX, WB4MJE, WB0KEK, WD4AHZ and WD4FAB station logs. My thanks to everyone who contributed.

[5] In cases where the sporadic-E region straddled time zones or moved during an opening, the more easterly time zone was used. Two or more active E-layer regions, separated in time or space, sometimes appeared on the same day.

[6] *QST*, Sep 1982, p 73 and Oct 1982, p 72; *SWOT Bulletin*, Aug and Sep 1982, and Feb 1983.

[7] E. Pocock, "Sporadic-E Propagation at VHF," loc cit.

[8] Patrick Dyer, WA5IYX, monitored the FM broadcast band (88 to 108 MHz) and the 2-meter band for evidence of sporadic E for an 11-year period, 1980 to 1990. The effective area of sporadic-E midpoints for signals arriving at his Austin, Texas, listening post encompassed a semicircular region arcing across New Mexico, Kansas, Oklahoma, northern Texas, Arkansas, Mississippi and Louisiana. See Pocock and Dyer, "Eleven Years of Sporadic E," loc cit.

[9] European 1977-91 data was compiled from several sources by Mick Toms and published in the *VHF/UHF DXer*, May 1992, pp 14-15. Additional 1992 data was added from subsequent *VHF/UHF DXer* issues.

[10] Pocock, "Sporadic-E Propagation at VHF," loc cit, pp 34-36.

[11] Thanks to Dyer for providing 1977-79 88-MHz data to augment that used in our 1992 study.

[12] E. K. Smith Jr, "The Occurrence of Sporadic E," in K. Smith and S. Matsushita, eds., *Ionospheric Sporadic E* (New York: Macmillan, 1962), pp 3-11.)

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Reviewed By Brian Battles, WS1O
QST *Features Editor*

These aren't books for ventriloquists' sidekicks; they're manuals for computer users, aimed at beginners, not "power users" who can make operating systems and programs jump through hoops by casually tickling the keyboard. In Amateur Radio, we call them newcomers (as the Novice license class once indicated); in computing, the authors call them "non-nerds."

You don't need to know all the details of how a transceiver works to operate a ham radio station with a level of proficiency, and the same may be said of personal computers (PCs). Certain radio amateurs are sometimes condescendingly tagged by their more technically proficient peers as "mere appliance operators" (some hams amiably accept the classification), and there are plenty of computer users worthy of the same sobriquet—perhaps many of whom are the same people. These books aren't published with the intent to keep people in the dark about computers, though; their goal is to get people up and running on the mysterious machines, and to help them fearlessly join the ubiquitous world of automated data processing brought on by the deluge of inexpensive microprocessors in everything from desktop PCs to telephone answering machines and video cassette recorders.

Looking briefly at each of these books' topics:

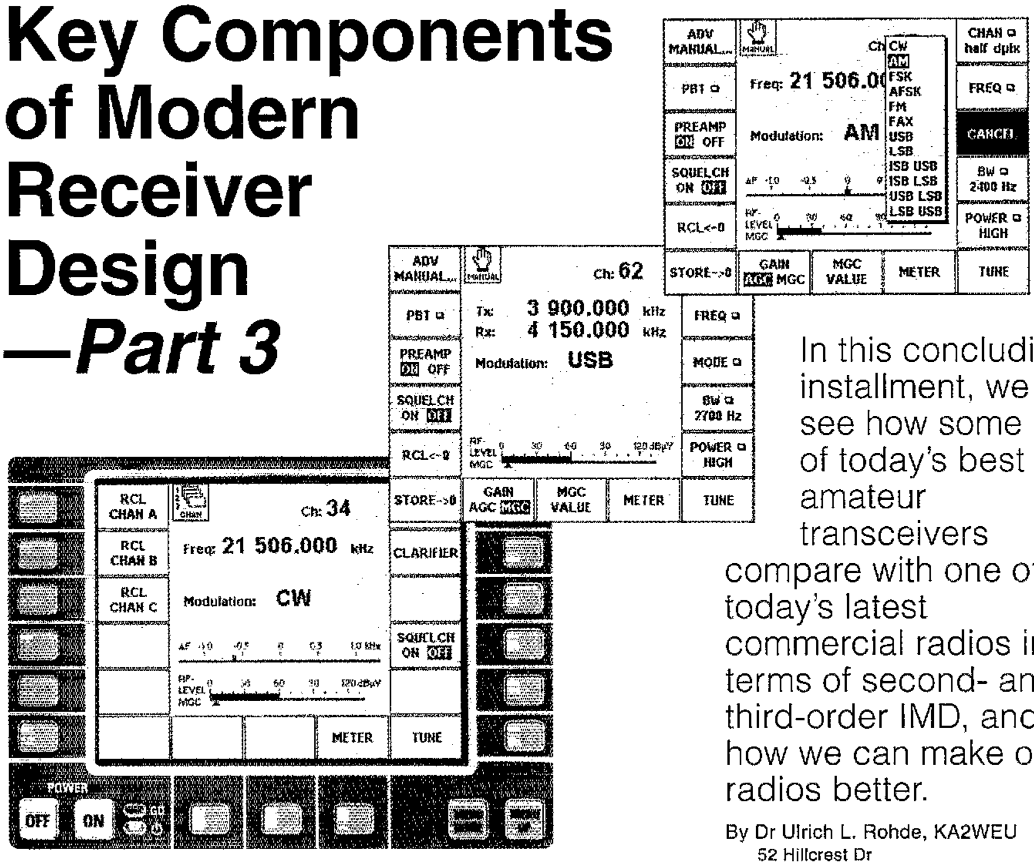
- ***DOS*** (or, more precisely, MS-DOS, as marketed by Microsoft Corp) is the most common PC operating system in use today. Although flexible and found on more machines than any other operating system, DOS presents the user with a stark monitor screen displaying only a text-based prompt, and it requires the user to memorize many commands, which must be typed on a keyboard.
- ***Windows*** is Microsoft's operating environment (or "graphical user interface") that runs on DOS-based computers. The idea is that it replaces the alphanumeric-text-only DOS (which requires commands to be entered via a keyboard) with a graphical front-end that lets a user select activities and perform many tasks by pointing at pictures and icons with a mouse. It also makes it possible to run more than one program at a time.
- ***OS/2*** is IBM's PC multitasking, multithreaded operating system that's an alternative to DOS and Windows. OS/2 is newer than DOS, UNIX and Windows, but has rapidly become popular in the past couple of years, largely because of its unique ability to run multiple programs written for OS/2, DOS and Windows simultaneously, and it shows great promise. OS/2 offers a graphical user interface similar in some ways to that provided by Windows.
- ***UNIX*** is a multitasking operating system that's an alternative to DOS and OS/2, developed for scientific and engineering work, and often found on machines used by universities, businesses and networks. There are several varieties of UNIX, some available free and others as commercial products. UNIX is widely known as a powerful system, but challenging for new users to learn. Most versions of UNIX provide a text-based interface that requires the user to memorize many commands that must be entered by typing them on a keyboard.
- ***The Internet*** is a worldwide network of computers and smaller networks that share access, data and programs with each other and via outside links, such as dial-up services and commercial online systems (like CompuServe, America On-Line, etc). The Internet grew from the now-defunct ARPANET, a resource created in 1969 primarily to share data and computer power among researchers and subcontractors for the US Department of Defense Advanced Research Projects Agency. It was superseded around 1980 by the NSFNET, funded by the National Science Foundation. Today the Internet has essentially subsumed the NSFNET and expanded exponentially in size, with the ongoing (daily) addition of networks and computers.

IDG Books claims to have more than five million ...*For Dummies* books in print, and it's easy to see why. Billed as "A Reference for the Rest of Us!" each title in this extremely popular series is written in plain English to take the reader through the essential steps necessary to use the computer product in question without clogging the brain with unnecessary technical gibberish. In fact, there are sidebars and information boxes with a "warning" icon to let the reader know that the material therein is mainly "nerdy technical discussions you can skip if you want."

If a person has just started using DOS, UNIX, OS/2, Windows or the Internet, he or she can choose to (A) read the manufacturers' often obfuscating manuals (if there's one available), (B) purchase "complete" third-party guides to using them, which are commonly full of the cumbersome details eschewed by the . . .*For Dummies* books, or (C) get books that explain what he or she can do and how to do it without being buried in the stuff the average person doesn't care about or need to know. The more complex levels can be explored later, after the user becomes familiar with the basics and at least seems to know what's going on.

Radio amateurs are a decidedly curious lot, however, so it may be that many hams would prefer more technical books on computer topics. For those whose main interest is *operating* a PC, not in wondering how the operating system or software does what it does, these may be just right to help you get started. The next question is how do you hide the covers so that no one can see the titles when they spot these references on your bookshelf? (And will they ever publish *Packet Radio for Dummies, Satellites for Dummies, Microwaves for Dummies, Moonbounce for Dummies*, etc?)

Key Components of Modern Receiver Design
—*Part 3*



In this concluding installment, we see how some of today's best amateur transceivers compare with one of today's latest commercial radios in terms of second- and third-order IMD, and how we can make our radios better.

By Dr Ulrich L. Rohde, KA2WEU
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Upper Saddle River, NJ 07548

In the preceding two parts of this article,* I have covered dynamic range, AGC and synthesizer concepts that are important in high-performance MF/HF reception, and offered some practical circuits for implementing these ideas. In this, the third and last installment, we investigate how the strong-signal performance of today's MF/HF transceivers can be improved, and compare the performance of a recent commercial counterpart.

Applying These Concepts to Commercial Transceivers

During a recent trip to Germany, I installed my favorite Hustler mobile antenna system (a combination of the MO-1 and RM-40) on the aluminum railing of my 19th-floor hotel window in the Munich Sheraton. This translates into an elevation of 180 feet above ground level, and among other things provides a spectacular view of Munich. With the Hustler antenna adjusted for resonance at 7.075 MHz (–3 dB bandwidth, 50 kHz), I scanned the 40-meter band with my TS-450S soon after sunset. Despite the antenna's short length—about 2 meters—incoming signals immediately overloaded the TS-450S's receiver. The radio's speaker put out nothing but noise, and its S meter indicated a minimum of S9 as I scanned 40 meters—yet I could recognize no amateur signals whatsoever! Only strong broadcast stations above 7.1 MHz—"strong" meaning 50 to 60 dB over S9—were clearly detectable. The receiver became marginally usable for weaker signals only after I turned on its 20-dB attenuator and its "AIP" feature. Using the same antenna (still resonant on 40 meters) for 20-meter reception with no input attenuation, I could hear several "ghost" broadcast stations (meaning that I found, on several frequencies, signals apparently carrying two different voice transmissions, one English and one Russian, with music from a third station superimposed on them.) These signals immediately disappeared—plunged from S9 to nothing—when I turned on the TS-450S's 20-dB attenuator. Switching from the 40-meter antenna to a properly tuned 20-meter mobile antenna also made these spurious signals disappear.

Despite recent advances in transceiver design, even the users of high-end transceivers, such as the IC-781, IC-765, TS-950SDX and FT-1000, sometimes find

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*Part 1 appeared on pp 29-32 of May 1994 *QST*, and Part 2 appeared on pp 27-31 of June 1994 *QST*. Corrections: Transistors Q1 and Q2 in Fig 7 are BFR96s and not BRF96s; the part-number prefix for the HP diodes mentioned is 5082 and not 8052, and these are PIN and not hot-carrier devices. The response spikes at non-line-harmonic offsets in the right half of Figures 15 and 18 are artifacts of the phase-noise-measured system used.

strange mixing products that in no way correlate with intermod measurements and calculations appropriate for third-order-IMD prediction. Readers of the first two parts of this article have probably already realized that these products come from *second*-order IMD. A typical frequency relationship involved in real-life second-order IMD is 6 MHz + 8 MHz = 14 MHz. In this case, 49-meter broadcast signals mix with signals from the fixed or maritime mobile services near 8 MHz, generating ghost broadcast signals every 5 or 10 kHz in the 20-meter amateur band. My experience with my TS-450S and balcony-mounted mobile antenna clearly show that a good-quality transceiver can get into trouble with second-order IMD even when using a fairly small antenna in a good location.

Improving Amateur Transceiver IMD Performance

In earlier parts of this article, I discussed switching-diode characteristics with emphasis on good second-order-IMD performance. To prove that installing PIN diodes in place of a radio's stock front-end-filter switching diodes can improve its second-order-IMD performance, I modified an ICOM IC-765, had Amateur Electronic Supply modify a Yaesu FT-890, and had Kenwood modify a TS-50 for me.[13] I measured the following results (all tests done at 14 MHz with preamp off, unless otherwise noted):

- The IC-765's second-order intercept increased from +60 to +95 dBm, and its third-order intercept point improved by 6 dB (to about +30 dBm).
- The FT-890's second-order intercept increased from 63 to 93 dBm, and its third-order intercept point increased from +20 dBm to +26 dBm.
- Unmodified, my TS-50 exhibited a second-order intercept point of approximately +35 dBm. Installing PIN diodes improved this to +48 dBm. Using the radio's optional external antenna tuner, which has a peaked high-pass characteristic, improved the second-order intercept point still further, from +85 to +90 dBm. (With two −20 dBm test tones, one at 3.5 MHz and the other at 3.6 MHz, and listening at 7.1 MHz, the TS-50's second-order intercept point is only +30 dBm because the diodes' IMD products become stronger as frequency decreases. With the tuner activated, they disappear.) Installing the PIN diodes also increased the TS-50's third-order input intercept point from +12 to +18 dBm.
- With its original front-end switching diodes in place, my TS-450S exhibited a second-order intercept point of +55 dBm. Reconfiguring the TS-450S's antenna tuner to operate in receive as well as transmit pushed its second-order intercept to

Table 2

Selected XK2100L Transceiver Specifications

General		
Switchover times		
TX/RX, RX/TX	< 10 ms	
Frequency Change	< 30 ms	
Receiver		
Input Impedance	50 Ω, VSWR < 3	
Noise figure		
without preamplifier	17 dB	
with preamplifier	9 dB	
Input sensitivity (for S/N of 10 dB from 0.2 to 30 MHz), without preamp		
A1A (CW)	0.4 µV EMF (−121 dBm), BW = 300 Hz	
J3E (SSB)	1.0 µV EMF (−113 dBm), BW = 2.7 kHz	
H3E (AME), 60% mod by 1 kHz	2.7 µV EMF (−104 dBm), BW = 6 kHz	
With preamp		
A1A (CW)	0.15 µV EMF (−130 dBm), BW = 300 Hz	
J3E (SSB), J7B	0.4 µV EMF (−121 dBm), BW = 2.7 kHz	
H3E (AME), 60% mod by 1 kHz	1.0 µV EMF (−113 dBm), BW = 6 kHz	
Receiving bandwidths	*−3 dB*	*−60 dB*
	150 Hz	300 Hz
	300 Hz	450 Hz
	600 Hz	860 Hz
	1 kHz	1.54 kHz
	1.5 kHz	1.98 kHz
	2.1 kHz	3.2 kHz
	2.4 kHz	3.52 kHz
	2.7 kHz	3.8 kHz
	3.1 kHz	4.2 kHz
	6 kHz	8.4 kHz
	8 kHz	10.4 kHz
AGC	< 3 dB output change over the range 1 µV to 1 V EMF	
Response to 60-dB step variation		
Attack	<10 ms	
Decay	25/150/500 ms, 1 s/3 s	
AF distortion		
Line output, 0 dBm	< 1%	
Headphones, loudspeaker	< 10% at rated power	
Blocking	3-dB signal attenuation (Δf = 30 kHz, desired signal 2 mV EMF, interfering signal 5 V EMF)	
Desensitization	> 20 dB SINAD (Δf = 30 kHz, BW = 2.7 kHz, desired signal 30 µV, interfering signal 100 mV)	
Intercept point (IP_3)	> 30 dBm (Δf > 30 kHz, interfering signals 2 × 0 dBm)	
Cross-modulation	< 10 % (Δf > 30 kHz, desired signal 1 mV EMF, 1 kHz, m = 30%)	
Inherent spurious signals	< −113 dBm, with few exceptions	
Immunity to interference (Δf > 30 kHz)		
Image-frequency rejection	> 80 dB, typically > 90 dB	
IF rejection	> 80 dB, typically > 90 dB	
Oscillator reradiation	< 10 µV (at antenna input)	
Protection of receiver input	up to 100 V EMF (f < 30 MHz)	
with digital preselector	up to 200 V EMF (f < 30 MHz)	
Transmitter		
Spurious suppression	> 70 dB, typically > 80 dB (into 50 Ω)	
Harmonic suppression	> 45 dB, typically > 60 dB (into 50 Ω)	
Intermodulation products (with two-tone modulation and 26.5-V supply voltage)	> 32 dB down, typically > 36 dB (referred to PEP)	
S/N ratio	> 150 dB, referred to 1-Hz test bandwidth, Δf > 1 MHz	
	> 165 dB with digital preselector option	
Carrier suppression	> 60 dB referred to PEP, typically > 70 dB	
with voice compression	> 30 dB	
Suppression of unwanted sideband	> 60 dB referred to PEP	

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[13] Notes appear on page 45.

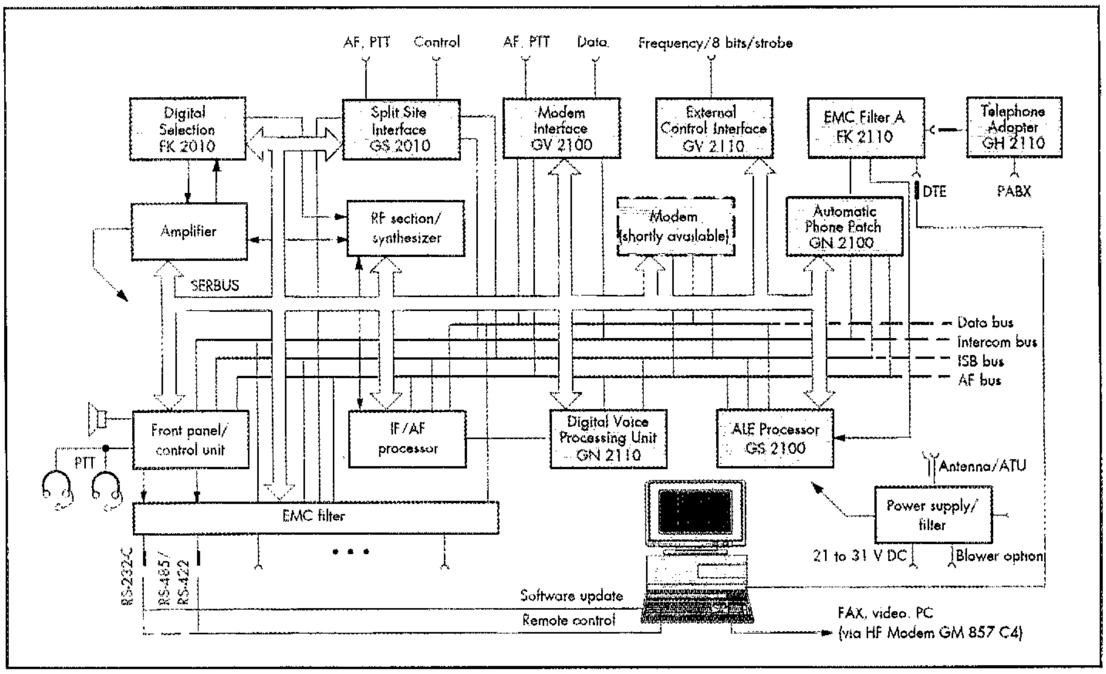

Figure 20—All of the XK2100L's modules use the Rohde & Schwarz SERBUS for control. Shaded modules are options.

about +81 dBm. Replacing its stock switching diodes with PIN devices increased the TS-450S's second-order intercept point to +60 dBm, while its third-order intercept point went from +22 dBm to 28 dBm. Operated from the same hotel with the same balcony-mounted antenna I mentioned earlier, the modified TS-450S exhibited none of the overloading problems it had exhibited before modification.

I believe that the second-order-IMD numbers for the IC-765 and FT-890 after modification are slightly higher because we also modified their antenna tuners to work in transmit and receive. (The inboard antenna tuners in today's Amateur Radio transceivers are usually in line only in transmit to avoid tuner hunting during split-frequency operation. Because tuner selectivity can directly improve a receiver's second-order IMD dynamic range, I urge equipment manufacturers to build in the option of using our radios' tuners in receive mode.)

I also evaluated a Collins KWM-380 (second-order intercept, +60 dBm), an unmodified TS-950SDX (second-order intercept, +62 dBm), an unmodified Ten-Tec OMNI VI (second-order intercept, +43 dBm) and finally an unmodified FT-1000 (second-order intercept, over +70 dBm). The TS-950SDX and the FT-1000 tested both included antenna tuners (transmit mode only) and stock switching diodes.[14]

How a Current Professional Transceiver Compares

With second- and third-order-IMD numbers for current high-end Amateur Radio transceivers in hand, it's time to look across to one of our radios' newest military/commercial relatives to see how good "good" radio performance can be. Then we can challenge Amateur Radio equipment manufacturers—who, quite understandably, operate under considerably tighter cost constraints than producers of professional/military radios—to find ways of working toward that performance in Amateur Radio products.

The new Rohde & Schwarz XK2100L, an MF/HF transceiver with receiver coverage from 10 kHz to 30 MHz, implements all of the design concepts discussed in this article. Software menus, not front-panel buttons and knobs, access most of its control functions. The title graphic shows part of the radio's control-menu structure.

Figure 20 diagrams the XK2100L's internal architecture. Like a personal computer, the XK2100L controls and communicates with its system modules via an internal bus (the Rohde & Schwarz SERBUS). Unlike current amateur MF/HF transceivers, which require the installation of new ROMs or EPROMs for firmware upgrades, the XK2100's operating firmware can be updated with a PC via its RS-232-C interface.

Digital signal processing (DSP) handles all of the XK2100L's radio's modulation and demodulation functions at IF and AF, allowing:

- A variety of emission classes, such as AME,[15] CW, SSB (USB and LSB), ISB (two independent sidebands), FM, FSK and AFSK (the deviation/shift and baud rate of which can be optimally matched in receive)
- 11 bandwidths from 150 Hz to 8 kHz, with group-delay-equalized filtering for data transmission
- Five AGC decay time constants from 25 ms to 3 s
- Passband tuning (with bar-graph passband indication)
- An IF notch filter (with bar-graph notch position indication)
- A noise blanker that automatically adapts to the interference's pulse repetition rate and pulse width
- Syllabic squelch (no threshold setting necessary)
- Voice compression that approximately doubles the transmitter's average output power during voice transmission

The XK2100L's receiver typically exhibits intercept points of +70 dBm (IP_2) and +35 dBm (IP_3), and 10% cross-modulation with an interference source of +20 dBm. Its 9-dB noise figure affords good reception even with short rod or whip antennas, and it can withstand antenna-terminal overvoltages up to 100 V EMF indefinitely. An

Second-Order IMD Testing Coming to *QST* Product Reviews

If Dr Rohde's findings on second-order IMD have convinced you that including the results of second-order-IMD testing in *QST* Product Reviews would be a good idea, you're not alone: We're convinced, too! As a result, we're working second-order-IMD testing into the ARRL Lab tests performed on Product Review units, and future *QST* Product Reviews of MF/HF transceivers and receivers will include the results of these tests.—*Mark J. Wilson, AA2Z,* QST *Editor*

optional digitally controlled preselector can be installed to provide a −20 dB front-end bandwidth equal to ±10% of the operating frequency. This preselector is in line during transmit and receive, and its tuning time is 20 ms.

On the transmit side, the XK2100L operates at 150 W (PEP SSB) and 100 W (CW mode), and supports automatic link establishment (ALE) in line with FED-STD 1045. Properly configured, the transceiver can act as a remote telephone link with all the amenities of a modern telephone set.

Table 2 shows selected XK2100L technical specifications for comparison with those of amateur transceivers. This raises the obvious question of cost. The simple answer is that a basic XK2100L without options sells in the $18,000 range. The useful answer is that I believe that strong-signal RF and AGC performance comparable to that of transceivers like the XK2100L can be achieved for amateur-market prices through more stringent system design. As an example of this, consider that, per *QST* Product Review testing, the ICOM IC-761 transceiver achieved a third-order intercept of +5 dBm at 20 meters (preamp on),[16] while a Kenwood TS-830S, designed over a decade earlier, achieved an input intercept of −5 dBm at 20 meters (preamp not defeatable)[17]—yet both radios use first mixers consisting of two 2SK125 JFETs. Being given the option of switching out an RF preamp is another aspect of progress. Switching off the preamp in the *QST* Product Review IC-761 increased its third-order intercept from +5 dBm to +21 dBm at 20 meters.

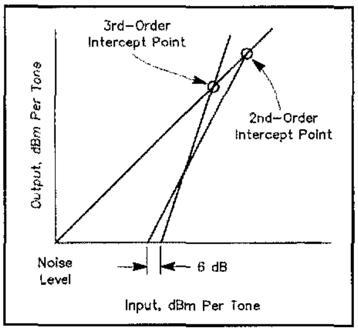

Figure 21—As the levels of input signals rise, second-order IMD products emerge from the noise first, and predominate a receiver's IMD products until signal levels rise far enough for third-order products to overtake them. (After Sabin 1981; see Note 8)

Summary

In this three-part article, I have described an integrated system approach for improved AGC, better selectivity and much-simplified frequency synthesis in MF/HF receivers and transceivers. Considerable better phase-noise performance should be achievable for little, if any, cost increase over that of circuitry already present in our transceivers.

The importance of good second-order IMD performance may be this article's most controversial finding. Figure 21 shows how second-order IMD can be more important than the third-order IMD products: As input-signal levels rise, second-order products emerge from the receiver noise first. Minor changes in input signal level, coupled with falling conditions, can therefore allow second-order products to predominate in some circumstances. Equipment manufacturers listened when radio amateurs asked for receivers with better third-order-IMD performance; now, we must work with them to minimize second-order IMD.[18]

The measurements I've presented show that replacing an amateur transceiver's front-end-filter switching diodes with PIN devices can significantly improve second-order-IMD performance, but innovation is still necessary in the area of input selectivity, possibly using electronic tracking filters. Even current high-end professional radios need highly selective sub-octave input filtering to achieve the maximum possible second-order intercept point. Being able to use our transceivers' antenna tuners in receive as well as transmit would be one step in the right direction.

Finally, a glimpse at a new commercial-grade MF/HF transceiver confirms that the latest professional/military radios are already implementing IF selectivity *digitally*, by means of DSP, while simplifying user interface through the use of multilevel menus rather than a forest of knobs, switches and buttons. Studying this radio's technical specifications suggests that state-of-the-art strong-signal and AGC performance may be within reach at amateur-market prices—if we are willing to forgo standard techniques in favor of the new.

Notes

[13] This procedure was complicated because the TS-50 and similar units from Kenwood use diodes in special SMD packages. I thank the managers at the Siemens Semiconductor Group for making custom diodes available for this test by taking standard 4-µs minority-carrier-lifetime diode chips and packaging them in the appropriate configuration.

[14] The considerable variation in IP_2 among the unmodified radios tested is surprising because all use similar switching diodes. One possible reason for this is that some radio designs, such as the FT-1000, may bias their diodes more optimally than others.

[15] AM equivalent (full-carrier SSB, or H3E).

[16] B. Schetgen, ed, *The ARRL Radio Buyer's Sourcebook*, 1st ed (Newington: ARRL, 1991), pp 2-37 to 2-42.

[17] B. Schetgen, ed, *The ARRL Radio Buyer's Sourcebook*, 1st ed, pp 2-89 to 2-92.

[18] For details on how to calculate IP_2 and IP_3, see U. Rohde, "Testing and Calculating Intermodulation Distortion in Receivers," *QEX*, Jul 1994.

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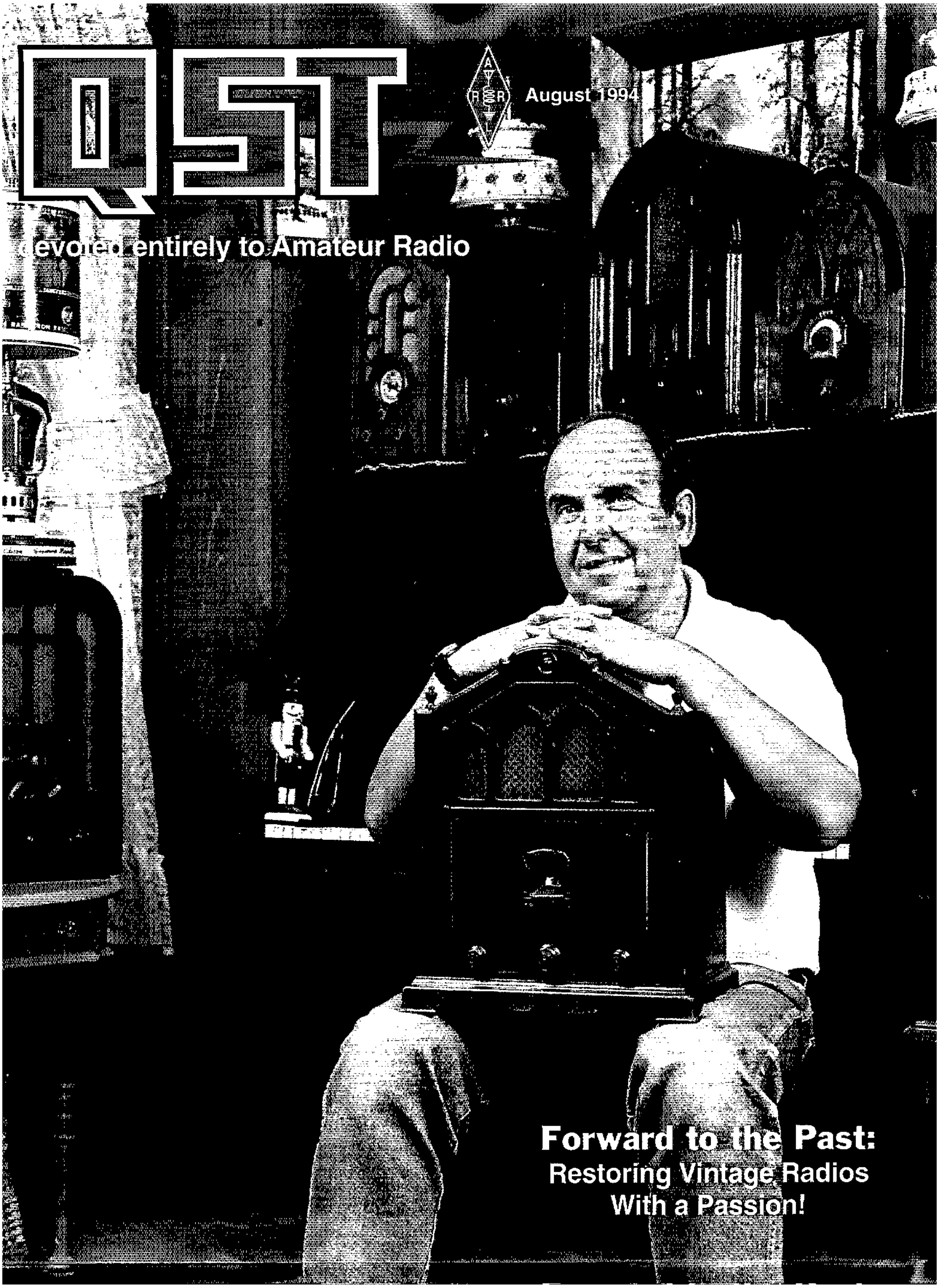

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OUR COVER

When Bob Eslinger, KR1U, bailed out of the corporate world a few years ago, he turned a life-long passion for radio into a business, restoring old receivers of all kinds (just so they use vacuum tubes!). Bob's rural Connecticut home is a virtual museum of old radios. An article about antique radio restoration and repair appears on page 42.
(photo by Kirk Kleinschmidt, NT0Z)

CONTENTS

August 1994
Volume LXXVIII Number 8

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

Two New Multiband Trap Dipoles

W8NX details a new coax trap design used in two multiband antennas; one covering 80, 40, 20, 15 and 10 meters, and the other covering 80, 40, 17 and 12 meters.

By Al Buxton, W8NX
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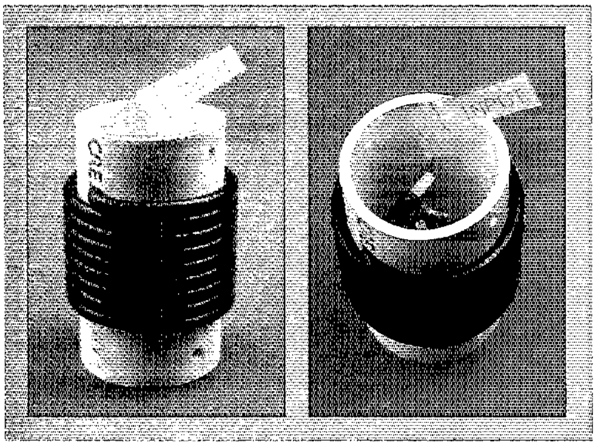

Over the last 60 or 70 years, amateurs have used many kinds of multiband antennas to cover the traditional HF bands. The availability of the 30, 17 and 12-meter bands has expanded our need for multiband antenna coverage. A fortunate few have the space and resources for multiband antennas like rhombics or long Vs, but many hams have employed inverted-L long wires or parallel dipoles. Old-timers will recall the off-center-fed *Windom* of the '30s—the first version using a single-wire transmission line, and the later design using two-wire feed line. Over the years, random-length dipoles with open-wire feeders and associated tuners have been used successfully as multiband antennas. The G5RV multiband antenna is a specialized example of this approach.[1]

The *log periodic array* represents a kind of brute-force approach to the goal of achieving coverage of multiple HF ham bands. It seems inefficient because of the large gaps between our relatively narrow amateur HF bands.

Over the last few decades, two factors have affected the development of multiband antennas—the popularity of low-impedance (usually 50-Ω) coaxial feed lines, and the appearance of untuned, 50-Ω solid-state amplifiers. The impedance of an antenna is relatively low only at its fundamental frequency and at odd-order harmonics. Although antenna tuners are often necessary to resonate an antenna system, the quest for expanded multiband coverage with simple antennas continues.

At the end of the 1930s, a different technological approach appeared in the form of resonant traps in antennas. The *Mims Signal Squirter* is the grandfather of modern day tribanders.[2] This article discusses in detail an innovative trap design employed in two multiband dipoles.

One W8NX Trap Design—Two Multiband Dipoles

Two different antennas are described here. The first covers 80, 40, 20, 15 and 10 meters, and the second covers 80, 40, 17 and 12 meters. Each uses the same type of W8NX trap—connected for different modes of operation—and a pair of short capacitive stubs to enhance coverage. Both antennas were designed using my "All About Trap Dipoles" software package.[3] The new W8NX coaxial-cable traps have two different modes: a high- and a low-impedance mode. The inner-conductor windings and shield windings of the traps are connected in series in the conventional manner for both modes. However, either the low- or high-impedance point can be used as the trap's output terminal. For low-impedance trap operation, only the center conductor turns of the trap windings are used. For high-impedance operation, all turns are used, in the conventional manner for a trap. The short stubs on each antenna are strategically sized and located to permit more flexibility in adjusting the resonant frequencies of the antenna.

Figure 1 shows the configuration of the 80, 40, 20, 15 and 10-meter antenna. The radiating elements are made of #14 stranded copper wire. The element lengths are the wire span lengths in feet. These lengths do not include the lengths of the pigtails at the balun, traps and insulators. The 32.3-foot-long inner 40-meter segments are measured from the eyelet of the input balun to the tension relief hole in the trap coil form. The 4.9-foot segment length is measured from the tension relief hole in the trap to the 6-foot stub. The 16.1-foot outer-segment span is measured from the stub to the eyelet of the end insulator. The coaxial-cable traps are wound on PVC pipe coil forms and use the low-impedance output connection. The stubs are 6-foot lengths of 1/8-inch stiffened aluminum or copper rod hanging perpendicular to the radiating elements. The first inch of their length is bent 90° to permit attachment to the radiat-

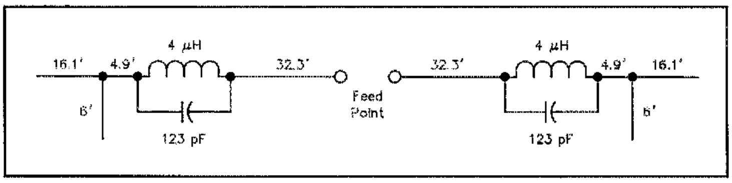

Figure 1—A W8NX multiband dipole for 80, 40, 20, 15 and 10 meters. The values shown (123 pF and 4 µH) for the coaxial-cable traps are for parallel resonance at 7.15 MHz. The low-impedance output of each trap is used for this antenna.

[1]Notes appear on page 29.

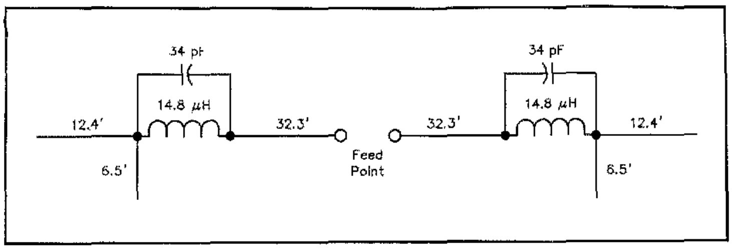

Figure 2—A W8NX multiband dipole for 80, 40, 17 and 12 meters. For this antenna, the high-impedance output is used on each trap. The resonant frequency of the traps is 7.15 MHz.

ing elements by large-diameter copper crimp connectors. Ordinary #14 wire may be used for the stubs, but it has a tendency to curl up and may tangle unless weighed down at the end. I recommend that you feed the antenna with 75-Ω coax cable using a good 1:1 balun.

This antenna may be thought of as a modified W3DZZ antenna[4] (shown for many years in various ARRL publications) with the addition of capacitive stubs. The length and location of the stub give the antenna designer two extra degrees of freedom to place the resonant frequencies within the amateur bands. This additional flexibility is particularly helpful to bring the 15 and 10-meter resonant frequencies to more desirable locations in these bands. The actual 10-meter resonant frequency of the W3DZZ antenna is somewhat above 30 MHz, pretty remote from the more desirable low frequency end of 10 meters.

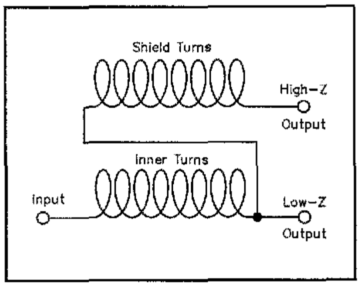

Figure 3—Schematic for the W8NX coaxial-cable trap. RG-59 is wound on a $2\frac{3}{8}$-inch OD PVC pipe.

Figure 2 shows the configuration of the 80, 40, 17 and 12-meter antenna. Notice that the capacitive stubs are attached immediately outboard after the traps and are 6.5 feet long, 0.5 foot longer than those used in the other antenna. The traps are the same as those of the other antenna, but are connected for the high-impedance output mode. Since only four bands are covered by this antenna, it is easier to fine tune it to precisely the desired frequency on all bands. The 12.4-foot tips can be pruned to a particular 17-meter frequency with little effect on the 12-meter frequency. The stub lengths can be pruned to a particular 12-meter frequency with little effect on the 17-meter frequency. Both such pruning adjustments slightly alter the 80-meter resonant frequency. However, the bandwidths of the antennas are so broad on 17 and 12 meters that little need for such pruning exists. The 40-meter frequency is nearly independent of adjustments to the capacitive stubs and outer radiating tip elements. Like the first antennas, this dipole is fed with a 75-Ω balun and feed line.

Figure 3 shows the schematic diagram of the traps. It explains the difference between the low and high-impedance modes of the traps. Notice that the high-impedance terminal is the output configuration used in most conventional trap applications. The low-impedance connection is made across only the inner conductor turns, corresponding to one-half of the total turns of the trap. This mode steps the trap's impedance down to approximately one-fourth of that of the high-impedance level. This is what allows a single trap design to be used for two different multiband antennas.

Figure 4 is a drawing of a cross-section of the coax trap shown through the long axis of the trap. Notice that the traps are conventional coaxial-cable traps, except for the added low-impedance output terminal. The traps are $8\frac{3}{4}$ close-spaced turns of RG-59 (Belden 8241) on a $2\frac{3}{8}$-inch-OD PVC pipe (schedule 40 pipe with a 2-inch ID) coil form. The forms are $4\frac{1}{8}$ inches long. Trap resonant frequency is very sensitive to the outer diameter of the coil form, so check it carefully. Unfortunately, not all PVC pipe is made with the same wall thickness. The trap frequencies should be checked with a dip meter and general-coverage receiver and adjusted to within 50 kHz of the 7150 kHz resonant frequency before installation. One inch is left over at each end of the coil forms to allow for the coax feed-through holes and holes for tension-relief attachment of the antenna radiating elements to the traps. Be sure to seal the ends of the trap coax cable with RTV sealant to prevent moisture from entering the coaxial cable.

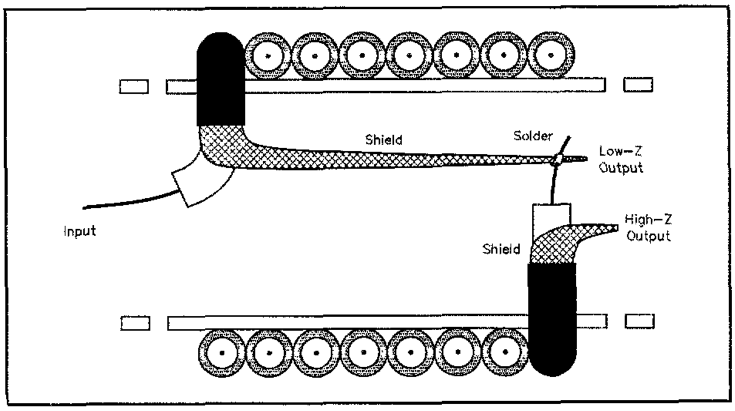

Figure 4—Construction details of the W8NX coaxial-cable trap.

Also, be sure that you connect the 32.3-foot wire element at the start of the inner conductor winding of the trap. This avoids detuning the antenna by the stray capacitance of the coaxial-cable shield. The trap output terminal (which has the shield stray capacitance) should be at the outboard side of the trap. Reversing the input and output terminals of the trap will lower the 40-meter frequency by approximately 50 kHz, but there will be negligible effect on the other bands.

The title-page photos show a coaxial-cable trap. Details of the trap installation are shown in Figure 5. This drawing applies specifically to the 80, 40, 20, 15 and 10-meter antenna, which uses the low-impedance trap connections. Notice the lengths of the trap pigtails: 3 to 4 inches at each terminal of the trap. If you use a different arrangement, you must modify the span lengths accordingly. All connections can be made using crimp connectors rather than by soldering. Access to the trap's interior is attained more easily with a crimping tool than with a soldering iron.

Antenna Patterns

The performance of both antennas has

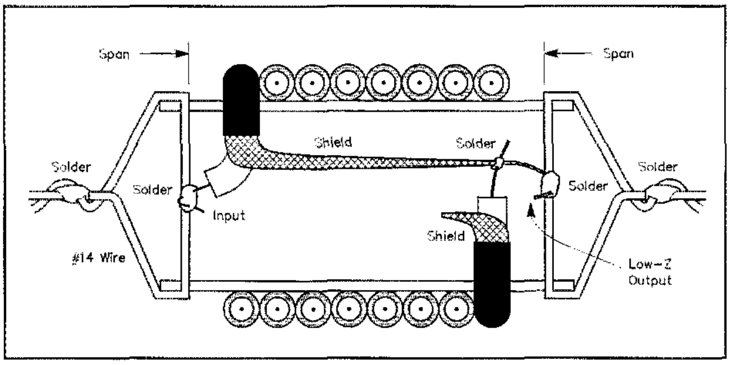

Figure 5—Additional construction details for the W8NX coaxial-cable trap.

been very satisfactory. I am currently using the 80, 40, 17 and 12-meter version because it covers 17 and 12 meters. (I have a tribander for 20, 15 and 10 meters.) The radiation pattern on 17 meters is that of 3/2-wave dipole. On 12 meters, the pattern is that of a 5/2-wave dipole. At my location in Akron, Ohio, the antenna runs essentially east and west. It is installed as an inverted V, 40 feet high at the center, with a 120° included angle between the legs. Since the stubs are very short, they radiate little power and make only minor contributions to the radiation patterns. The pattern has four major lobes on 17 meters, with maxima to the northeast, southeast, southwest, and northwest. These provide low-angle radiation into Europe, Africa, South Pacific, Japan and Alaska. A narrow pair of minor broadside lobes provides north and south coverage into Central America, South America and the polar regions.

There are four major lobes on 12 meters, giving nearly end-fire radiation and good low-angle east and west coverage. There are also three pairs of very narrow, nearly broadside, minor lobes on 12 meters, down about 6 dB from the major end-fire lobes. On 80 and 40 meters, the antenna has the usual figure-8 patterns of a half-wave-length dipole. I have some pattern distortion and input impedance effects from aluminum siding on my house. Nevertheless, DX is easily workable on either of these antennas using a 100-W transceiver, when the high-frequency bands are open.

Both antennas function as electrical half-wave dipoles on 80 and 40 meters with a low SWR. They both function as odd-harmonic current-fed dipoles on their other operating frequencies, with higher, but still acceptable, SWR. The presence of the stubs can either raise or lower the input impedance of the antenna from those of the usual third and fifth harmonic dipoles. Again, I recommend that 75-Ω, rather than 50-Ω, feed line be used because of the generally higher input impedances at the harmonic operating frequencies of the antennas.

The SWR curves of both antennas were carefully measured. A 75 to 50-Ω transformer from Palomar Engineers was inserted at the junction of the 75-Ω coax feed line and my 50-Ω SWR bridge. The transformer prevents an impedance discontinuity, with attendant additional undesired line reflections appearing at the 75 to 50-Ω junction. The transformer is required for accurate SWR measurement if a 50-Ω SWR bridge is used with a 75-Ω line. No harm is done to any equipment, however, if the transformer is omitted. Most 50-Ω rigs operate satisfactorily with a 75-Ω line, although this requires different tuning and load settings in the final output stage of the rig or antenna tuner. I use the 75 to 50-Ω transformer only when making SWR measurements and at low power levels. The transformer is rated for 100 W, and when I run my 1-kW PEP linear amplifier the transformer is taken out of the line. (I hope my absent-mindedness doesn't catch up with me some day!)

Figure 6 gives the SWR curves of the 80, 40, 20, 15 and 10-meter antenna. Minimum SWR is nearly 1:1 on 80 meters, 1.5:1 on 40 meters, 1.6:1 on 20 meters, and 1.5:1 on 10 meters. The minimum SWR is slightly below 3:1 on 15 meters. On 15 meters, the stub capacitive reactance combines with the inductive reactance of the outer segment of the antenna to produce a resonant rise that raises the antenna input resistance to about 220 Ω, higher than that of the usual 3/2-wavelength dipole. An antenna tuner may be required on this band to keep a solid-state final output stage happy under these load conditions.

Figure 7 shows the SWR curves of the

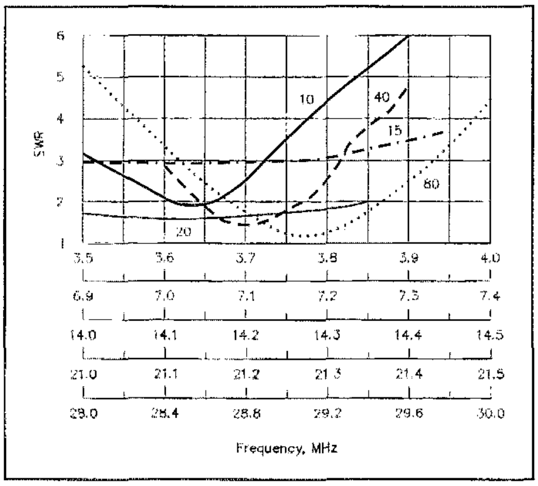

Figure 6—Measured SWR curves for an 80, 40, 20, 15 and 10-meter antenna, installed as an inverted-V with 40-ft apex and 120° included angle between legs.

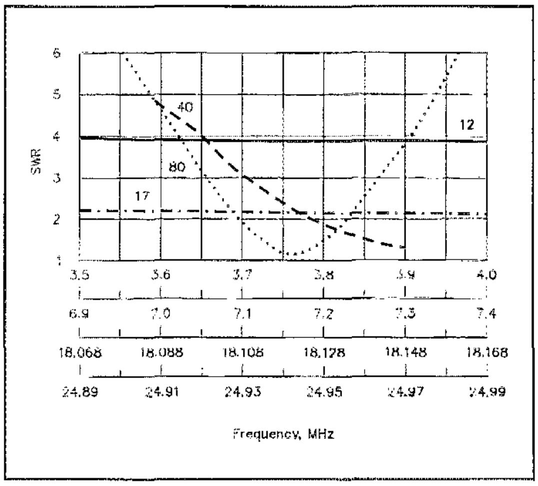

Figure 7—Measured SWR curves for an 80, 40, 17 and 12-meter antenna, installed as an inverted-V with 40-ft apex and 120° included angle between legs.

80, 40, 17 and 12-meter antenna. Notice the excellent 80-meter performance with a nearly unity minimum SWR in the middle of the band. The performance approaches that of a full-size 80-meter wire dipole. The short stubs and the very low inductance traps shorten the antenna somewhat on 80 meters. Also, observe the good 17-meter performance, with the SWR being only a little above 2:1 across the band.

But notice the 12-meter SWR curve of this antenna, which shows 4:1 SWR across the band. The antenna input resistance approaches 300 Ω on this band because the capacitive reactance of the stubs combines with the inductive reactance of the outer antenna segments to give resonant rises in impedance. These are reflected back to the input terminals. These stub-induced resonant impedance rises are similar to those on the other antenna on 15 meters, but are even more pronounced.

Too much concern must not be given to SWR on the feed line. Even if the SWR is as high as 9:1, *no destructively high voltages will exist on the transmission line.* Recall that transmission-line voltages increase as the square root of the SWR in the line. Thus, 1 kW of RF power in 75-Ω line corresponds to 274 V line voltage for a 1:1 SWR. Raising the SWR to 9:1 merely triples the maximum voltage that the line must withstand to 822 V. This voltage is well below the 3700-V rating of RG-11, or the 1700-V rating of RG-59, the two most popular 75-Ω coax lines. Voltage breakdown in the traps is also very unlikely. As will be pointed out later, the operating power levels of these antennas are limited by RF power dissipation in the traps, not trap voltage breakdown or feed-line SWR.

Trap Losses and Power Rating

Table 1 presents the results of trap Q measurements and extrapolation by a two-frequency method to higher frequencies above resonance. I employed an old, but recently calibrated, Boonton Q meter for the measurements. Extrapolation to higher-frequency bands assumes that trap resistance losses rise with skin effect according to the square root of frequency, and that trap dielectric losses rise directly with frequency. Systematic measurement errors are not increased by frequency extrapolation. However, random measurement errors increase in magnitude with upward frequency extrapolation. Results are believed to be accurate within 4% on 80 and 40 meters, but only within 10 to 15% at 10 meters. Trap Q is shown at both the high- and low-impedance trap terminals. The Q at the low-impedance output terminals is 15 to 20% lower than the Q at the high-impedance output terminals.

I computer-analyzed trap losses for both antennas in free space. Antenna-input resistances at resonance were first calculated, assuming lossless, infinite-Q traps. They were again calculated using the Q values shown in Table 1. The radiation efficiencies were also converted into equivalent trap losses in decibels. Table 2A summarizes the trap loss analysis for the 80, 40, 20, 15 and 10-meter antenna and Table 2B for the 80, 40, 17 and 12-meter antenna.

Table 1

Trap Q

Frequency (MHz)	3.8	7.15	14.18	18.1	21.3	24.9	28.6
High Z out (Ω)	101	124	139	165	73	179	186
Low Z out (Ω)	83	103	125	137	44	149	155

Table 2A

Trap Loss Analysis: 80, 40, 20, 15, 10-Meter Antenna

Frequency (MHz)	3.8	7.15	14.18	21.3	28.6
Radiation Efficiency (%)	96.4	70.8	99.4	99.9	100.0
Trap losses (dB)	−0.16	−1.5	−0.02	−0.01	−0.003

Table 2B

Trap Loss Analysis: 80, 40, 17, 12-Meter Antenna

Frequency (MHz)	3.8	7.15	18.1	24.9
Radiation Efficiency (%)	89.5	90.5	99.3	99.8
Trap losses (dB)	−0.5	−0.4	−0.03	−0.006

The loss analysis shows radiation efficiencies of 90% or more for both antennas on all bands except for the 80, 40, 20, 15 and 10-meter antenna when used on 40 meters. Here, the radiation efficiency falls to 70.8%. A 1-kW power level at 90% radiation efficiency corresponds to 50-W dissipation per trap. In my experience, this is the trap's survival limit for extended key-down operation. SSB power levels of 1 kW PEP would dissipate 25 W or less in each trap. This is well within the dissipation capability of the traps.

When the 80, 40, 20, 15 and 10-meter antenna is operated on 40 meters, the radiation efficiency of 70.8% corresponds to a dissipation of 146 W in each trap when 1 kW is delivered to the antenna. This is sure to burn out the traps—even if sustained for only a short time. Thus, the power should be limited to less than 300 W when this antenna is operated on 40 meters under prolonged key-down conditions. A 50% CW duty cycle would correspond to a 600-W power limit for normal 40-meter CW operation. Likewise, a 50% duty cycle for 40-meter SSB corresponds to a 600-W PEP power limit for the antenna.

I know of no analysis where the burn-out wattage rating of traps has been rigorously determined. Operating experience seems to be the best way to determine trap burn-out ratings. In my own experience with these antennas, I've had no traps burn out, even though I operated the 80, 40, 20, 15 and 10-meter antenna on the critical 40-meter band using my AL-80A linear amplifier at the 600-W PEP output level. I have, however, made no continuous, key-down, CW operating tests at full power purposely trying to destroy the traps!

Summary

Some hams may suggest using a different type of coaxial cable for the traps. The dc resistance of 40.7 Ω per 1000 feet of RG-59 coax seems rather high. However, I've found no coax other than RG-59 that has the necessary inductance-to-capacitance ratio to create the trap characteristic reactance required for the 80, 40, 20, 15 and 10-meter antenna. Conventional traps with wide-spaced, open-air inductors and appropriate fixed-value capacitors could be substituted for the coax traps, but the convenience, weatherproof configuration and ease of fabrication of coaxial-cable traps is hard to beat.

Notes

[1] L. Varney, "The G5RV Multiband Antenna... Up-to-Date," *The ARRL Antenna Compendium, Vol. 1*, p 86.

[2] M. Mims, "The Mims Signal Squirter," *QST*, Dec 1939, p 12.

[3] Available from Al Buxton, W8NX, PO Box 174, Columbus, OH 43216; Price: $24.95 plus $5 for shipping. Specify 3.5 or 5.25-inch floppy disk.

[4] "Five Band Antenna," *The ARRL Antenna Book*, 16th Edition, pp 7-10 to 7-11.

Al Buxton is no stranger to the pages of QST. *He was first licensed in 1937 as W7GLC, and has had a distinguished career in both industry and academia. Now retired, Al is active in Amateur Radio and computer application studies in antenna development. This article continues a series on transmission lines, antenna traps and trap-dipole antennas.* QST

Direct Digital Synthesis—An Intuitive Introduction

Transceiver ads hype DDS, but what is it? Here's where you'll find out.

By Howie Cahn, WB2CPU
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Boston, MA 02116
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Direct digital synthesis (DDS) is a relatively new technique that is a powerful and promising method for generating signals. When you tune a modern HF transceiver, you're probably using a DDS circuit. DDS output has high stability and low phase noise. It can switch frequencies very quickly, usually within microseconds. Its frequency can be set very precisely, often to within a fraction of a hertz. These properties make DDS a natural for transmission techniques such as spread spectrum, quadrature modulation, or frequency-shift keying, in addition to the more standard modes. Although DDS has limitations that I'll describe later, DDS should be the method of choice for many applications requiring a variable frequency source.

In spite of its advantages, DDS is not widely understood or used by hams in home-brew equipment. Part of the reason is the newness of DDS. There aren't many books that cover the subject. Also, many hams aren't aware of the ICs that have become available in just the last few years that make implementing a DDS system easy. Unlike virtually all other frequency generation methods, DDS is a *digital* process. It's based on logic circuits such as *storage registers* and *arithmetic units*. A DDS circuit is actually a very specialized computer. The output waveform is computed *on the fly* by digital hardware instead of being the result of the physical resonance of components such as quartz crystals, inductors and capacitors, or mechanical resonators.

In this article, I'll describe both the concepts and the circuitry for DDS. I'll keep things nontheoretical and use only as much math as needed. So, if you want to use DDS in home-brew gear, or if you just want to understand what's under the hood of your transceiver, this should get you started.

Back to School

Although the idea behind DDS is not inherently difficult to understand, its concept often takes a while to sink in. Eventually, you'll get it and the "light bulb" in your head will turn on. I'll try to shorten the process by metaphorically explaining how DDS works—by comparing it to something you may have already experienced.

In your high school trigonometry class, you may have been given an assignment such as: "Draw one cycle of a sine wave." Let's review how it's done. First, you must decide (or be told) how many points to plot to establish the curve. Let's plot, say, 20 equally separated points, corresponding to one point every 18 degrees: 360° ÷ 20 = 18°. In your trig table, you look up the sine of the angle of the first point and plot it. Then you add 18, look up and plot the next point and continue the process until a full cycle is completed. For instance, sin 0° = 0.000, sin 18° = 0.309, sin 36° = 0.588, etc. When you're finished, you should have a graph resembling that shown in Figure 1A. It's a familiar sine wave showing the sine value (on the Y axis) for an angle (on the X axis). In a real system, a cycle takes a specific amount of time—the cycle period. So the X axis represents not only the angle, but also the time.

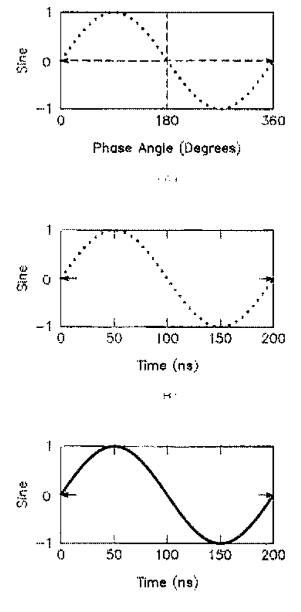

Figure 1—At A, a sine-wave plot; B, 5-MHz sine-wave plot; C, smoothed 5-MHz sine-wave plot.

To add the concept of frequency to our example, let's rephrase the assignment as: "Draw a 5-MHz sine wave based on points plotted every 10 nanoseconds." A frequency of 5 MHz corresponds to a period of 200 ns. Calculating a point every 10 ns means that there will be 20 points per 200-ns period. We can consider the interval between the points either as a time difference (10 ns), or as a specific fraction of the sine-wave cycle at the given frequency. This fraction is the *phase-angle* difference. In this case, the phase difference is again 360° ÷ 20 = 18°. Since this is the same phase difference as in the graph we originally plotted, we can relabel the graph as shown in Figure 1B. Connecting the points gives the sine curve shown in Figure 1C.

We've created, at least on paper, a 5-MHz sine wave. It was constructed from events that occur every 10 ns, which is the same as saying they have a frequency of 100 MHz. Let's call this the input frequency, f_{in}. The 5-MHz sine wave is the output frequency, f_{out}. Let's describe what we did a bit more precisely.

1. The input frequency determined the time interval between the plotted points. At the output frequency, this corresponds to a phase difference (in degrees) of $(f_{out} \div f_{in}) \times 360$.
2. At each point, we added this phase difference to the current phase angle.
3. Then we looked up the sine of this phase angle in a table.

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[1] Notes appear on page 32.

4. We continued repeating Steps 2 and 3 until we finished one cycle—although there's no reason we couldn't have gone on indefinitely.

5. Finally, we produced a smooth sine wave by connecting the dots.

From Process to Circuit

A direct digital synthesizer is simply *the hardware needed to carry out this process.* Most of its circuit elements are blocks of digital logic.

For Step 1, we need an input frequency source. This source determines the phase noise and stability of the system. A poor source may result in a system with noise and stability problems.

In Step 2, we must add the phase difference computed in Step 1 to the current phase angle. I'll refer to this phase difference as the *phase increment value.* Addition is performed by a circuit called a *digital adder.* To repeat: At each point, the phase increment value is added to the previous phase angle to yield the new phase angle. Now, we need a way to store this phase angle. In digital systems, values are stored in *registers.* The register used to hold the phase value is called the *phase accumulator.* A register can store some number, *n*, bits of data. As we'll see, in a DDS circuit, the greater *n* is, the more precisely the frequency can be controlled. The parts of the circuit I've described so far are at the top of Figure 2. You see that the phase accumulator is reloaded at each input clock pulse with the sum of the phase increment value and the previous value of the phase accumulator. Each of the lines shown that connect the blocks actually represents *n* digital signal lines.

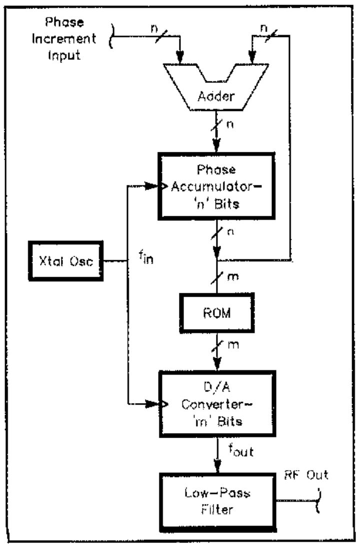


Figure 2—Block diagram of a DDS system.

In Step 3, we must convert the phase angle stored in the phase accumulator into its sine value. Again, it's just like in high school. In school, to get the sine of an angle, you looked it up in a table in the back of the book. (At least those of us who went to school before calculators did!) Here, it's the same routine—except that the look-up table is in a read-only memory (ROM) instead of a book. The output of the ROM is a binary number representing the sine of the phase angle at the current point in the waveform. The value is then converted to an analog voltage (or current) using a digital-to-analog converter (DAC).

Finally, in Step 5, the output is smoothed by a low-pass filter. You can see the full DDS circuit in Figure 2.

To make the example easier to understand, I picked the input and output frequencies so that $f_{in} \div f_{out}$ was a round number: 20. In fact, the only restriction on the relationship between these frequencies is that the output frequency can't be more than half the input frequency. This limit is imposed by something called the *sampling theorem*, otherwise known as the *Nyquist limit.* For a practical circuit this limit is actually about 0.4 instead of 0.5. Therefore, a DDS with a clock frequency of 30 MHz can produce an output no greater than 12 MHz.

To use a DDS circuit, you need two equations. First, you have to compute the phase increment value, the amount to add to the accumulator at each interval.

Phase increment value: The value to add to the phase accumulator, expressed in degrees is:

$$\Delta\Phi = (f_{out} \div f_{in}) \times 360 \qquad \text{(Eq 1)}$$

If you're designing or specifying a DDS circuit, you also need to know how large a phase accumulator is required to be able to set the frequency with the precision necessary for your requirement. The resolution of the system is based on the capacity of the phase accumulator and is given by the formula:

$$\Delta_f = f_{in} \div 2^n \qquad \text{(Eq 2)}$$

where n is the number of bits in the phase accumulator.

Frequency Resolution

The change in output frequency caused by a 1-bit change in the phase increment value is determined by: $\Delta_f = f_{in} \div 2^n$, where n is the number of bits in the phase accumulator.

For instance, for a system with a 24-bit phase accumulator and an input clock frequency of 33.5544432 MHz (again, picked to make the numbers come out evenly), changing the phase increment value by one bit changes the output frequency by $f_{in} \div 2^n = 33{,}554{,}432 \div 16{,}777{,}216 = 2$ Hz.

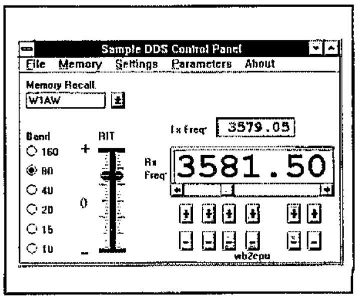
Interfacing

A DDS circuit is controlled by programming values into its registers. For example, to set the output frequency for the circuit in Figure 2, you enter parallel data representing the binary equivalent of the phase increment value into the Phase Increment Input line. Although this is different from the way traditional radio equipment works, having digital inputs and programmability can be an asset. For one thing, as you'll see below, a single DDS can be reconfigured to perform many different functions depending on how you program it. Another advantage is that you have options about how to connect, or interface, to a digital circuit. There are several ways to generate the binary data to program a DDS. I'll describe three of them.

The simplest method uses thumbwheel switches to produce the digital signals needed for the DDS inputs. The switches generate a binary version of the frequency represented by the numbers dialed into them. A ROM is used to convert this binary code into the code for the phase increment value needed by the DDS to generate this frequency. A variation of this technique was described recently by John Welch, N9JZW (see the References).

A second approach is similar to what's done in modern commercial transceivers. You adjust the frequency with a tuning knob. Instead of the knob being connected to a variable capacitor or potentiometer, however, here, the tuning shaft connects to a *digital shaft encoder.* With every revolution, the encoder produces pulses. Each pulse, when turning the knob clockwise, adds one to the number in a digital counter. Likewise, pulses from counterclockwise rotation decrement the counter. The binary output of this counter feeds the DDS. To allow you to see the current frequency, usually the counter also drives an LCD or LED display. Often a microprocessor is used to control the process.

A third method uses a personal computer to generate the digital data. Most ham shacks are equipped with a PC these days, so why not use it to operate your radio? You enter data using the computer's keyboard and/or mouse and get information from the display screen. The display can be



Memory Recall
W1AW

Band ● 80

Tx Freq 3579.05

Rx Freq 3581.50

Figure 3—Sample DDS radio panel.

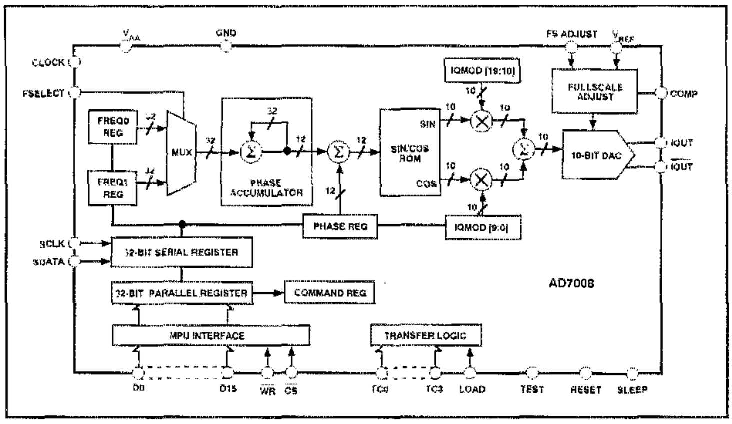


Figure 4—Functional block diagram of the Analog Devices AD7008 DDS modulator. *(Courtesy of Analog Devices)*

made to look like a familiar front panel of a radio. (This is referred to as creating a *virtual interface*—see Figure 3.) The advantage of this approach is flexibility. Software controls how you interact with your radio. There are no holes to drill or parts to buy to change the interface—you just load new software. In many cases, user-interface controls can be designed to be faster and easier to use than the physical radio panel controls they replace. Figure 3 is an example of a virtual radio interface screen showing several types of frequency adjustment controls. The software generates the data to program the DDS based on the frequency you enter. The data is sent from the computer to the radio using the machine's parallel or serial port, or by means of a bus-interface card.

Available ICs

What has made DDS practical in the last few years is the availability of ICs that incorporate many of the functions shown in Figure 2 into one chip and that can perform these operations fast enough to be useful for RF work. These chips are offered by several manufacturers including Qualcomm, Harris, and Analog Devices. The Analog Devices AD7008 is one of the easiest to use because it includes the DAC and features a flexible mechanism for programming the chip. (A block diagram of this device is shown in Figure 4.) The AD7008 can be used with an input clock frequency of up to 50 MHz. As we saw, this means it can produce frequencies up to 20 MHz. Its phase accumulator register is 32 bits wide. Using Eq 2, this means that we can set the frequency to within $(f_{in} \div 2^{n}) = 0.012$ Hz(!), although, in practice, the accuracy is limited by the accuracy of the input frequency source.

The AD7008 provides several features beyond those in the system shown in Figure 2. There are two registers for holding phase-increment values. In Figure 4, these registers are labeled **FREQ0 REG** and **FREQ1 REG**. Each can be set for a different frequency. You can switch between the registers with a logic-level signal on the IC's **FSELECT** pin. Being able to switch frequencies without having to program registers makes it easy to add equipment features such as frequency-shift keying or receive/transmit frequency offsets.

Notice the registers labeled **PHASE REG** and **IQMOD**. These let you add phase, frequency, or quadrature amplitude modulation suitable for SSB, to the signal being generated. This is done by feeding a stream of digitized values (representing the modulation signal) into the appropriate register.

Limitations

Although DDS technology has come a long way in a short time, it isn't suitable for every use. Two remaining problems are spurious output signals and a maximum frequency limit. A principal cause of spurious signals, or spurs, is quantization error, an inherent consequence of digital systems.

In a digital system, a value is represented by a specific number of binary data signals, or, bits. The number of bits determines how precisely a quantity can be represented. For example, with 8 bits, you can cover any of the integers from 0 to 255. You can't specify, say, 171½. You have to pick 171 or 172. Either way, there's an error of ½. In a DDS circuit, the limiting factor is the resolution of the DAC. For each point, there is going to be a difference between one of the exact steps the DAC can hit and the value needed to exactly represent the sine value of the angle at that point. This small difference shows up in the output signal as a spurious frequency component.

DACs currently available for RF DDS have resolutions in the range of 8 to 12 bits. The AD7008 contains a 10-bit DAC. There is no easy formula that will tell you where the spurious signals are going to fall. A rough prediction is that the strongest spurs will be approximately 20 log (2^{-n}) dB, or 6 dB per bit, down, where n is the DAC size in bits. Thus, for a 10-bit DAC, spurs will be about 60 dB below the desired signal. This is fine for many applications—amateur transmitters, for example. Spurs at this level may be undesirable in a high-performance receiver, however.

The DDS ICs that are currently available can produce maximum frequencies in the 20 to 40-MHz range. This prevents their use at VHF and above, and even in some HF applications like upconverting receivers. You can get around this by frequency multiplying or mixing the output to the needed frequency. Another approach, one that's used in commercial radios, is to combine DDS and phase-locked loop (PLL) techniques in a single circuit. This preserves the advantages of DDS—low phase noise and high resolution—and reduces the problems of spurious signals and maximum frequency. Both the frequency limit and the problem of spurious signals will be improved in future generations of components. It won't be long before DDS is the right tool for an even greater number of applications.

References

These references provide additional information on direct digital synthesis theory and applications.

John Welch, "The TW-1 Cheap Digital Front Panel," *73 Amateur Radio Today*, Mar 1994, pp 26-32.

Howie Cahn, "Connecting Computers to Radios—A PC Interface to the Ramsey 2-meter Transceiver," *Communications Quarterly*, Fall 1993, pp 13-29.

Bryan Bergeron, "Direct Digital Synthesis," *Communications Quarterly*, Summer 1993, pp 13-28.

Robert Schetgen, et al, *The ARRL Handbook for Radio Amateurs*, (Newington: ARRL, 1994), pp 10-17 to 10-23.

John Welch, "The Techno-Whizzy I, Part I," *73 Amateur Radio Today*, Dec 1992, pp 8-18; Part II, Jan 1993, pp 10-14 and 85.

Howie Cahn has been licensed since he was 12, receiving his Novice and General licenses in 1962, and his Amateur Extra in 1964. (WB2CPU is his original call although, at the time, Howie had no idea that CPU's had anything to do with computers!) When he realized the connection, Howie went to Cornell University to study electrical engineering, received his BSEE in 1970, and later did graduate study in EE and computer science. For the last 17 years, Howie has been designing hardware and software for Teradyne, Inc, a Boston-based manufacturer of electronics automatic test equipment. QST

COURTESY SRI

Simple Equipment for HF Fox Hunting

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J. M. Lomasney, WA6NIL
SRI International
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Figure 1—String along with the authors as they show you how to build a simple, yet effective HF direction-finder! This photo shows the complete setup: portable shortwave receiver on a metal plate, with a Faraday shield around the antenna for close-in work. An incoming signal is nulled when the whip is pointed directly at or directly away from it.

Glen Rickerd, KC6TNF, described simple and inexpensive equipment for VHF direction finding.[1] His method used a hand-held transceiver inside a hollow conducting cylinder held vertically near the operator's body. Our HF direction-finding system uses the same approach, even at wavelengths 10 times greater. In addition to tracking down transmitters operating from 3 to 30 MHz, you can use this project to locate local sources of radio-frequency interference (RFI). Transmitter ("fox") hunting is not commonly practiced at HF, probably because of the size and relative complexity of equipment for this frequency range, compared to that used at VHF.

Our equipment uses a portable HF receiver (a Sony ICF-7600 is shown) that is sensitive and has an internal BFO. When properly mounted on a small metal plate and with its whip antenna extended, this receiver forms a rotatable directional system with a null like that of a conventional loop. Figure 1 is a photo of the device in use; Figure 2 shows how the receiver and whip are mounted on the plate. This device works best with ground wave signals, but will also null sky wave signals under some conditions. (Sky wave signals don't always travel in direct paths, and may also be subject to scattering, which makes it difficult to achieve a well-defined null. —*Ed.*) For sky wave signals the level of performance is about equal to the time-honored shielded, single loop.

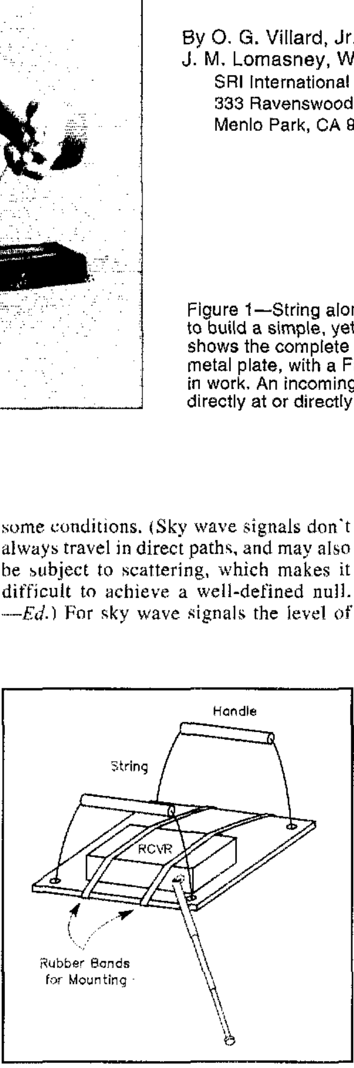

Figure 2—Receiver mounting on the conductive plate. The receiver whip is extended along the diagonal of the square plate, and should always be in the plane of the plate. Use rubber bands to hold the receiver in place. The conductive plate can be made from aluminum, brass, copper or copper-clad PC-board material.

Construction and Assembly

Figure 3 shows how the receiver's whip is pointed on end with respect to a radio wave. The wave has parallel fronts, so that the whip is aligned so as to be perpendicular to the electric field. No RF current is induced in the whip, except as a result of field distortion caused by the radio itself. If the reradiated energy is electrically symmetrical with respect to the whip, however, the antenna will still null the signal. Fortunately, the radio can be combined with a conductive plate in such a way that the reflected energy is symmetrical. In the null direction (the direction of whip, radio and plate), neither the incident wave nor backward-directed symmetrical receiver scatter induce signal energy into the whip and the radio. The result is a clean null if there are no additional signal sources present.

Modern portable SW receivers are thin enough to mount face up on the metal plate. When grounded to the plate, the radio effectively becomes part of the plate. (For best results the receiver ground should be connected directly to the plate. Of course, receiver-to-plate capacitance will—at least partially—provide a connection in any event.) Electromagnetically, the plate-

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[1] G. Rickerd, KC6TNF, "A Cheap Way to Hunt Transmitters," *QST*, Jan 1994, pp 65-66.

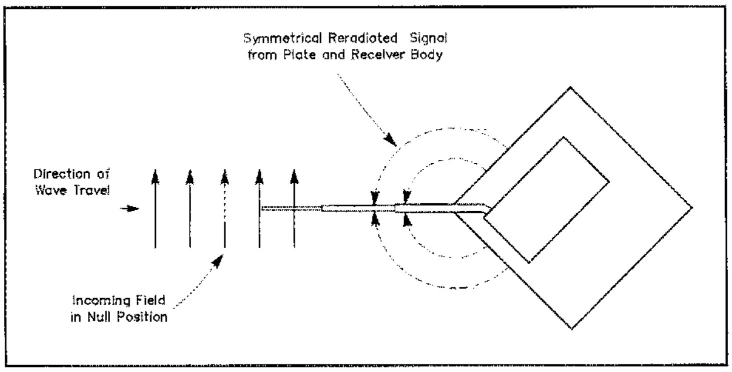


Figure 3—Field lines of the incoming wave are perpendicular to the whip and induce no voltage in it. When the receiver is mounted on the conductive plate, the receiver is symmetrical with respect to the whip. Parts of the end-on incoming signal wave are uniformly scattered and don't upset the antenna null by inducing voltage into the whip.

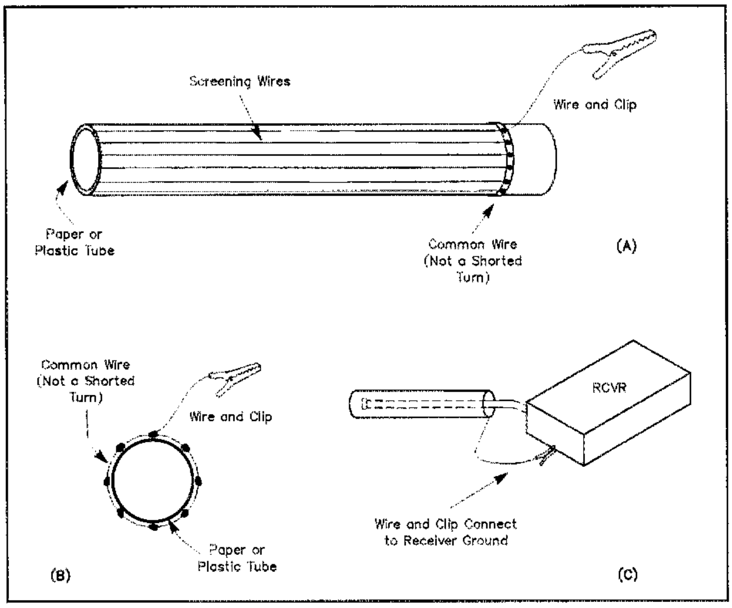


Figure 4—A Faraday shield over the whip reduces signal pickup and makes directional nulls more apparent. Wires in the shield are insulated from the antenna by the hollow paper or plastic tube (about 1-inch diameter) but connected to the receiver ground (through a clip connected to one side of an external power or antenna connector or headphone jack). Shield construction details are given in the text. A is a side view of the shield, B is an end view and C shows the shield in place.

plus-receiver of Figures 2 and 3 is essentially equivalent to the plate alone. A good null is obtained when the incident wave is aligned with the receiver whip, because the wave reradiated from the receiver also has a null in that direction. When DFing, the plate, receiver and whip must always maintain the same mechanical position with respect to one another. The whip should always lie along a line formed by an extension of a line drawn diagonally across the plate and in the plane of the plate.

Whip length is not critical. A longer whip will provide more signal to the receiver, however, the arrangement may become mechanically awkward. If the whip is too short, bearing accuracy suffers along with sensitivity. The size and thickness of the plate (which can be made of any convenient metal) is not critical. The plate can even be made of nonconducting material, such as cardboard or wood, and covered with aluminum foil. Arranging for the radio and plate to form a symmetrical structure is complicated by the unsymmetrical mounting of whips on most portable receivers. A plate area 20 to 50% larger than that of the radio usually permits a symmetrical layout.

The strings supporting the plate and radio (Figure 1) must be nonconducting. The device can be rotated back and forth like a puppet, while being held a foot or so from your body. Attach the radio to the plate with rubber bands attached to cup hooks. Most adjustments can be made from the top of the receiver despite the rubber bands, which easily can be removed when the receiver is needed elsewhere.

Operation

In the minimum-response position, the whip points in the direction of the station (or its reciprocal). In finding a null, accuracy can be improved by rotating the device back and forth across that bearing and forming a mental average of the null position. Although the plate can be tilted slightly if desired to improve results, horizontal orientation seems generally best. To average out local anomalies, we recommend you make measurements at various locations at least several feet apart.

One other circumstance needs to be taken into account. The human body produces an echo, just as does the receiver on its plate. Fortunately, body scatter falls off with distance just as rapidly as receiver scatter. (Both represent "near fields.") Depending on circumstances, scatter usually becomes unnoticeable at distances of 3 to 5 feet. Faraday screening of the whip reduces this distance appreciably. Otherwise, the effect of body scatter can be minimized by the same expedient used to minimize the effect of receiver scatter. Since the body is reasonably symmetrical, it will not appreciably affect bearing accuracy when located along a line oriented toward the distant transmitter and located either downstream or upstream of the receiver and its whip. A good procedure is to hold the receiver with its whip pointing perpendicularly outward from the body, and then to rotate body and whip together. If you place the device on a turntable, the bearing is readily found by trial and error, and by keeping your body in the right position and as far from the plate/receiver as possible.

The swinging plate device works particularly well when the signal is weak. The null direction is indicated either by an increase in noise, or by a change (normally a weakening) of any modulation. It is necessary to infer strength changes in this indirect way because the RF gain control of most portable shortwave receivers isn't accessible. By adjusting whip length, however, you can somewhat control sensitivity.

Sometimes the signal becomes so strong that, even with the whip retracted to the last section, the gain changes can no longer be distinguished. You can further reduce sensitivity by holding the equipment close to the ground. A more elegant way to attenuate the received signal is to cover the whip with a partial Faraday shield. Make the shield in the form of a concentric cage of equally spaced parallel wires grounded at one end to the plate or the receiver chassis (Figure 4). Cage diameter and length are not critical. About ½ to 1 inch diameter is about right. The shield reduces the amount of RF reaching the whip, but doesn't significantly alter directionality, because the whip and the shield are essentially parallel and concentric. You can make the shield from hookup wire taped to a paper cylinder that slips over the whip without electrical contact. A shield of eight wires introduces an attenuation of about 15 dB. With more wires, the attenuation is proportionally increased, as the shielding becomes more effective. Perhaps another 10 to 15 dB of attenuation is possible with more wires.

A different technique is to make use of body absorption, in the manner described by KC6TNF in Note 1. Retract the whip antenna to the last section, install the Faraday shield and hold the entire assembly vertically in front of your torso. It seems to be important to orient the radio perpendicular to the body. As you and the equipment turn, you'll find the null. This technique is useful for checking results obtained with radio, plate and whip horizontal. This technique will also help determine which null actually points toward the oncoming wave.

In some cases the received signal may become so strong that it leaks into the radio, thus spoiling the directional response. When this happens, place the radio inside a metal cylinder for further attenuation.

Conclusion

Fox hunting with simple, lightweight equipment is possible at HF, with standard, unmodified battery-powered receivers. Battery-powered QRP transmitters would be ideal "foxes." (This equipment will also be helpful when nearby hams' vacuum cleaners are accidentally transmitted on 20 meters!—*Ed.*) QST

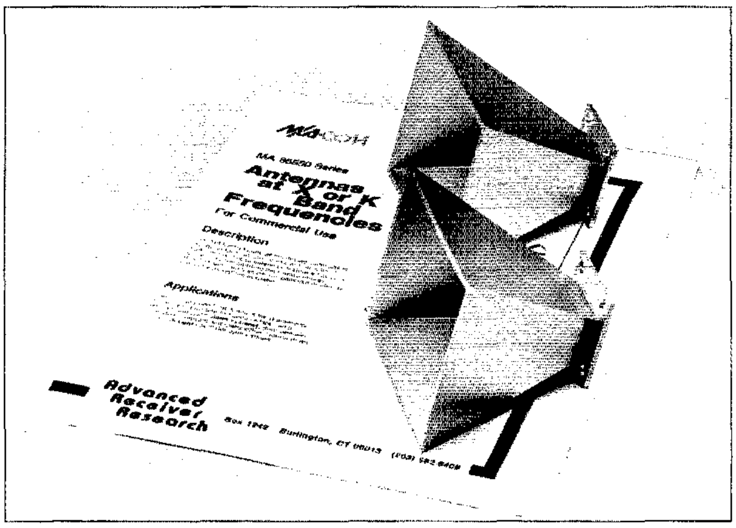
New Products

TEN-TEC KITS

◊ The hams at Ten-Tec present their new line of T-Kits. These include a 2-meter power amplifier for hand-held transceivers that offers 1 to 5 W input with 20 to 35 W output ($74); the Argonaut II RF deck for HF QRP operation (5 W at 160 to 10 meters) ($95); desk ($49) and hand-held ($29) replacement microphone kits for a variety of transceivers; a microprocessor-controlled, 10-memory 2-meter FM transceiver with LED display and 5 or 35 W output ($195); HF/VHF SWR bridge and wattmeter for 1.8 to 30 and 144 to 148 MHz ($49); module boards; antenna products; enclosures; SWL equipment; and more. Most kits are also available factory wired and tested at an extra charge. T-Kit, a Division of Ten-Tec Inc, 1185 Dolly Parton Pkwy, Sevierville, TN 37862-3710; tel 800-833-7373 or 615-453-7172; fax 615-428-4483.

RF-SHIELDED BOXES

◊ Frustrated by home-built devices that leak stray RF or clock signals, disrupting the operation of other circuits? Keep the RF from getting into or out of your project with an RF-tight, hot tin-plated steel box. SB-series shielded boxes come in many compact sizes, from 2.1×1.9×1.0 inch to 6.4×2.7×1.1 inches, at retail prices from $4.50 to $13.20. Feedthrough capacitors cost 85 cents to $3.50 each. There's also a complete line of modular construction boxes, rack chassis, metal cabinets, dual-slope cabinets, and "Rackem 'N' Stackem" boxes and racks. Sescom Inc, 2100 Ward Dr, Henderson, NV 89015; tel (orders only) 800-634-3457, fax (orders only) 800-551-2749; info 702-565-3400; tech line 702-565-3993; fax 702-565-4828.



MICROWAVE ANTENNA

◊ It's easier than you think to experiment with operation at 10 GHz, especially with the popular, low-cost MA86551 X-band antenna (pictured above). The horn antenna has been retooled for better performance and value. It's made of injection-molded ABS, and electrodeposited with 0.0004-inch copper and a 0.0003-inch nickel final plating for a bright, mirror-like finish. It covers 8 to 12.4 GHz, with a center frequency of 10.525 GHz, and a 25° E- and H-plane beamwidth. The integral flange mates with UG-39/U. Retail price is $20. Jay Rusgrove, W1VD, Advanced Receiver Research, Box 1242, Burlington, CT 06013; tel 203-485-0310.

SATELLITE MAGAZINE

◊ The publisher of *Monitoring Times* announces the forthcoming publication of the world's first full-spectrum satellite monitoring magazine, *Satellite Times*. Managing editor Larry Van Horn, N5FPW, says that the bimonthly magazine will cover all phases of satellite communication, including amateur, military, commercial, broadcasting, scientific, government, personal communication and private systems. Charter subscriptions are (US funds) $16.95 (US), $23.95 (Europe), $26 (Europe via air mail). Grove Enterprises, 300 S Hwy 64 W, PO Box 98, Brasstown, NC 28902; tel 800-438-8155 or 704-837-9200; fax 704-837-2216. QST

The 160-Meter Sloper System at K3LR

Sure, doesn't everyone have a 190-foot tower in his backyard? With help from his friends, K3LR made his 160-meter dream come true.

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Jim Breakall, WA3FET
308 Lutz Ln
Port Matilda, PA 16870

During the spring of 1992, we began the development work on a 160-meter antenna system using half-wave slopers. This antenna was to be installed on the 190-foot tower belonging to Tim Duffy, K3LR. Our goal was to build an array that would provide forward gain in any one of several switch-selectable compass directions. Good rejection of high-angle signals was an important requirement, as was overall efficiency. We describe in this article a quick review of the theoretical design process, and then discuss the construction, testing and operation of the actual array.

Background

Perhaps the best-known directional antenna using sloping dipoles is that of Dave Pietraszewski, K1WA.[1] His design utilizes five identical λ/2 slopers spaced uniformly around a mast tall enough so the dipoles descend toward the ground at an angle of 60° below horizontal. All five radiators are fed with equal lengths (slightly over 135°) of coax. Only one element is driven at a time, and the other four open-circuited transmission lines function as loading inductors so that all of the passive elements act as reflectors.

A second type of directional array using slopers was described by Dennis Mitchell, K8UR.[2] This antenna system requires only four dipoles, and the lower half of each radiator is "pulled in" (see Figure 1) so that it slopes back toward the base of the tower. Changing the geometry this way produces a signal whose polarization is almost entirely vertical, depending on the exact disposition of the wires. All of the elements in the K8UR antenna are driven with equal-amplitude currents phase-shifted by either 0°, 90°, or 180°, just like the classic "4-Square" phased-vertical array. As with all such phased arrays, the challenge is to get the feed currents into each element exactly right, so the full potential of the array can be realized, particularly for front-to-back ratio.

Design

The initial design of the K3LR antenna combines some of the best features from both the K1WA and K8UR arrays—the mechanical simplicity of the K8UR design, with the straightforward feed system of the parasitic K1WA array. The physical appearance of the K3LR system is close to

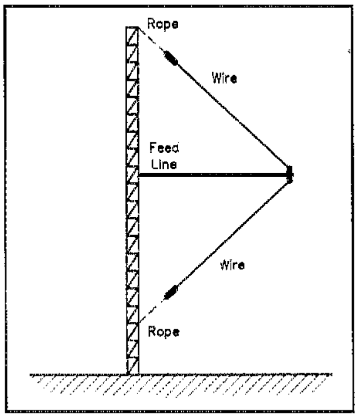


Figure 1—The K8UR Sloper System uses four identical half-wave sloping dipoles spaced uniformly around a tall mast. The lower half of each dipole is pulled in toward the tower.

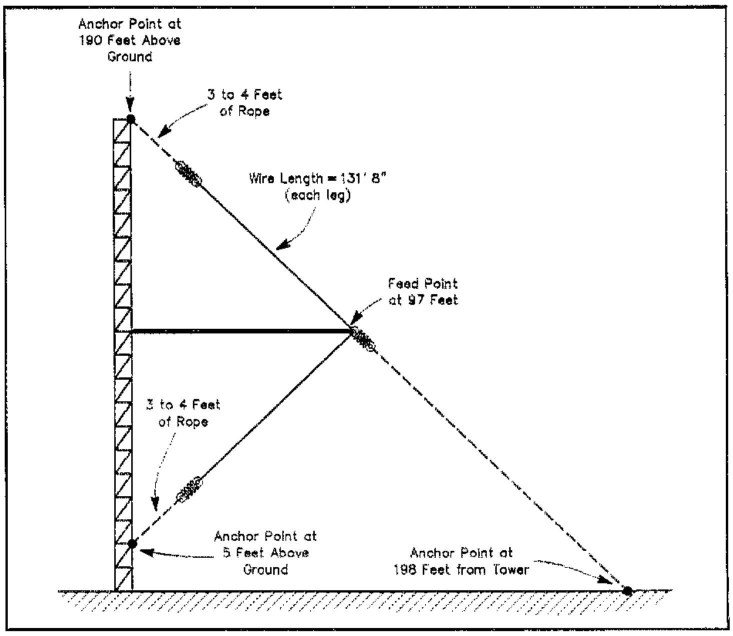


Figure 2—Detail of one element (of four) of the K3LR Sloper System, after tuning adjustments to compensate for insulation on element wires.

[1] Notes appear on page 38.

that of K8UR's, consisting of four identical bent dipoles spaced at 90° intervals around a tall mast. Figure 2 shows the layout of a single element, including all of the final dimensions. This array is electrically similar to that of K1WA, because only one element at a time is driven. The remaining three dipoles are inductively loaded by open-circuiting the far end of each individual feeder, so that all three act as parasitic reflectors. Only four elements are used (rather than five as described by K1WA) because modeling with *ELNEC*[3] indicated that there were no performance advantages to be gained by using the extra dipole. Figure 3 illustrates the principal-plane radiation patterns of this early-stage K3LR antenna system.

For the 160-meter array, experimentation with *ELNEC* indicated that the best combination of gain and front-to-back ratio would occur when the parasitic elements were loaded with about 100 Ω of inductive reactance. To produce this amount of loading, an electrical length of 153.45° of open-circuited *lossless* 50-Ω line was needed at the center of each element. Because of the long cable lengths, we chose to use RG-8X (instead of RG-213) for the feeders in order to reduce the suspended weight. With a velocity factor of 78%, the calculated physical length of each transmission line is 177.74 feet at the design frequency of 1.840 MHz.

Since the four RG-8X feeders actually have a small amount of loss, even at 160 meters, the impedance at the center of each reflector due to the open-circuited lines was not purely inductive but had become complex, with a value of 25.76 + *j*93.45 Ω. When this corrected loading impedance was substituted into *ELNEC*, the radiation patterns that resulted were rather surprising. Figure 4 shows that there is a small reduction in forward gain—but a dramatic improvement in front-to-back ratio!

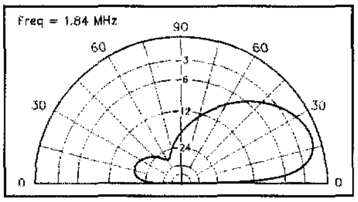


Figure 3—K3LR Sloper System, early stage modeling (elevation plot). The outer ring is 6.0 dBi, maximum gain of array is 5.06 dBi.

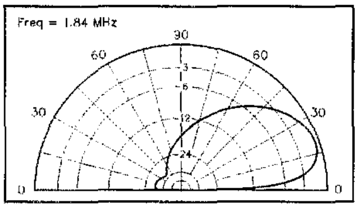


Figure 4—K3LR Sloper System, including feed-line losses in computations (elevation plot). The outer ring is 5.0 dBi, maximum gain of array is 4.57 dBi.

Table 1

***NEC*-Predicted Performance of the 160-Meter Sloper System in Various Configurations**

Description	*Forward Gain (dBi)*	*Front-to-Back Ratio (dB)*
No HF beams, no radials	2.88	18.59
No HF beams, 4 radials	3.85	15.59
4 HF beams, no radials	3.27	14.42
4 HF beams, 4 radials	3.74	19.10

We wanted even more gain, so a decision was made to add four elevated quarter-wave radials to the antenna system. These radials would be horizontal, mounted about 10 to 15 feet above the ground, and all four were to have their inner ends connected directly to the tower. Since the radials were close to the ground in terms of wavelength, we decided to use *NEC*[4] with its Sommerfeld/Norton ground model (rather than *ELNEC*, with its simplified Fresnel reflection coefficient ground model) for the remainder of our computer analysis. In order to be really precise, we felt that it would also be necessary to include all of the small antennas (relatively speaking, of course) which were already mounted on the K3LR 190-foot tower. So, two full-size 3-element 40-meter beams (at 100 and 190 feet), a 6-element 10-meter beam (at 198 feet), and a 3-element 20-meter beam (at 60 feet) were added to the *NEC* computer model, along with the tower, the four slopers and the four elevated radials.

Figure 5 is a drawing of the *NEC* model for the complete tower, including the entire top-band sloper system and the four HF beams. The principal elevation and azimuthal-plane radiation patterns for the 160-meter array are shown in Figure 6. *NEC* indicates a maximum gain of 3.74 dBi at a take-off angle of 20°, and a front-to-back ratio of 19.10 dB. The driving-point impedance predicted by the computer (for any one of the four elements) is 80.4 + *j* 65.2 Ω at a frequency of 1.84 MHz.

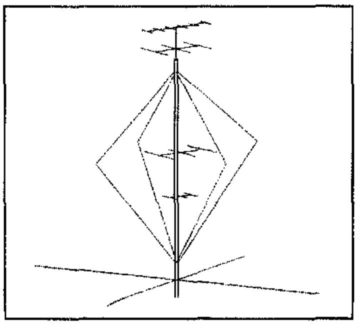


Figure 5—*NEC* model of the 160-meter array as mounted on K3LR's 190-foot tower.

Table 1 shows the *NEC*-generated values for the forward gain and front-to-back ratio when the top-band antenna is mounted on the 190-foot tower, both with or without the HF beams, and with or without the four elevated horizontal radials.

Construction

The four bent dipoles were made from standard #14 insulated electrical wire. The insulation necessitated an increase in the final element lengths in order to achieve resonance at the desired frequency. All of the elements were fed with home-brew ferrite-bead current baluns, with the length of the balun taken into account when

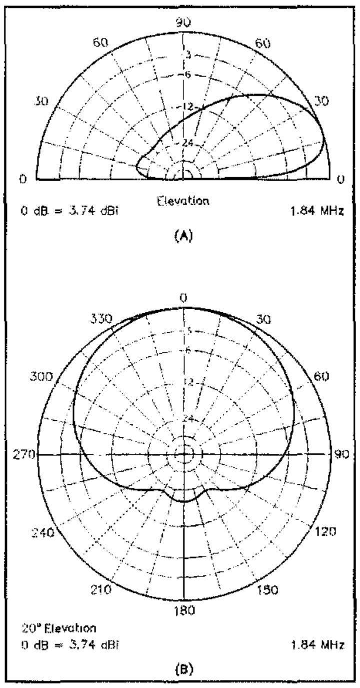


Figure 6—At A, elevation-plane radiation pattern for the K3LR 160-meter array. At B, azimuthal-plane radiation pattern. Maximum gain is 3.74 dBi. *NEC* models ground losses more accurately than *MININEC*-based programs when wires are located close to ground. The elevated radials are 13 feet high. The four antennas have their maximum radiation in the NE, SE, SW and NW directions.

determining the electrical length for each of the Belden RG-8X feed lines. These slopers are oriented at 45°, 135°, 225°, and 315°.

We used a fiberglass box to house the antenna-switching circuitry, so that the coax connectors would be insu-lated from each other without having to use special hardware. Four double-pole double-throw Deltrol relays (from Surplus Sales of Nebraska[5]) were used for driven-element selection. Five UHF chassis-type female coax connectors were installed on the outside of the fiberglass enclosure, with the 12-V relays mounted inside the box immediately adjacent to the SO-239 fittings. A conventional antenna switch will not work for this project, because both the center conductor and the shield must float, to allow the coaxial feeders to act as loads.

Four 15-foot pressure-treated 4×4-inch wooden poles were installed at azimuth angles of 0°, 90°, 180° and 270° to support the 133-foot elevated radials attached to the tower at a height of 13 feet. These horizontal radials are also made of #14 insulated wire.

Testing

When initially constructed in accordance with the original design dimensions, the radiators were a little bit short, since minimum SWR occurred at about 1.86 MHz. Each element was then lengthened by 1.3 feet (for a total of 131 feet, 8 inches) to bring the resonant frequency to 1.84 MHz. At that point, the SWR in the directional mode was so low that no impedance-matching was required.

The array seemed to perform somewhat erratically at first, but it became much more stable after the "common point" was grounded. This was done by running a short, heavy wire from the chassis-mounted SO-239 coax connector for the main feed line (at the Hoffman enclosure) to the base of the tower, which is set in cement and grounded via three 8-foot rods. In this manner, the outer shield of the main coaxial feeder is tied directly to a good earth ground immediately adjacent to the relay switch-box.

The antenna system was tested both with and without the four elevated radials, and the front-to-back ratio appeared to be at least 20 dB in either case. The difference in front-to-back ratio, which had been predicted by *NEC* was not noticed, either because it was too small to be discerned, or because other environmental factors in the real world were involved. The array seemed to play slightly better with the radials (perhaps because of some extra forward gain), so they were left in place.

Since the wire elements and the support ropes are very long, there is a fair amount of sag in the system. As a result, the lower ends of the slopers actually overlap at the base of the tower, and the wires extend rather close to the ground. Thus, all four ends are spaced well apart, in order to avoid arcing, which can occur if the wire elements should accidentally touch each other (or the tower itself). This problem could be avoided completely if the anchor-points for the support ropes could be raised off the ground, or moved farther away from the tower base. In addition, the array would fit somewhat better if the tower itself were slightly taller.

Table 2

Approximate Dimensions for the Sloper System on Various Amateur Bands

Freq (MHz)	*Tower Attach (Feet)*	*Element Length (Feet)*	*Anchor Distance (Feet)*	*Feed Line Length* (Feet)*	*Radial Length (Feet)*	*Radial Height (Feet)*
1.8	200	131	208	176.2	133	13
3.7	100	65.5	104	87.8	66.5	10
7.1	60	34	62	44.9	34.5	6

* These lengths assume that Belden RG-8X is used, and that a 14-inch length of RG-142 is added at the feedpoint for construction of a ferrite-bead current balun.

Operation

The sloper system was used for the first time during the CQ World-Wide SSB DX Contest in October 1992. K3LR was operated in multi-multi class, and the new 160-meter antenna worked very well. The operators (Alan, N3BJ, and Scott, WR3G) felt that the array was "loud" and that there were no problems hearing or working any station. The transmitting setup was a bit compromised by the use of an old amplifier that put out only 800 W. Even at that power level they completed 124 QSOs, working 13 zones and 32 countries. In the 1993 SSB Contest, the score was 102 QSOs, 12 zones, and 36 countries. All of these numbers stack up very well against other top multi-multi entries.

The sloper system was again put on the air during the ARRL 160-Meter Contest in early December 1992. This time K3LR was entered in the multi-single category with WR3G, W3YQ and K3LR as operators. The final total was 1333 QSOs and 99 multipliers, a new all-time record for this category. During the contest, the big antenna was compared to an inverted V (apex at 150 feet) located 750 feet away from the 190-foot tower. The parasitic array was always one S-unit better than the inverted V in a desired direction, as long as the station was at least 500 miles away. However, there were times when close-in stations were better on the inverted V.

During the 1993 CQWW CW Contest, the top-band sloper system really shined. With nearly 200 QSOs in 72 countries, the array kept Tim's station close to the top on 160 meters. K3LR worked 104 Europeans, compared with the 111 European stations worked by K1AR. Since K3LR is in Western Pennsylvania (only half a mile from Ohio), the new antenna is the "secret weapon" that enables Tim to be competitive with the East Coast stations.

Interactions

K3LR has noticed some *minor* fluctuations in the SWR readings taken while rotating the lower 40-meter beam (which is mounted at 100 ft), and he attributes these variations to the presence of the 160-meter sloper system. Otherwise, there have been no discernible effects on the remaining HF beams due to the array.

Using the Array on a Different Band

The design can be scaled easily to other frequencies, and suggested *initial* dimensions are given in Table 2. We are looking forward to receiving comments from others who build or modify this antenna system.

Acknowledgments

We'd like to thank Scott Jones, WR3G, and Tim Jellison, W3YQ, for their assistance in building the antenna, and for their operational observations.

Notes

[1] D. Pietraszewski, K1WA, "7-MHz Sloper System," *The ARRL Antenna Book*, 1991, pp 4-12 to 4-14.

[2] D. C. Mitchell, K8UR, "The K8UR Low-Band Vertical Array," *CQ*, Dec 1989, pp 42 to 46.

[3] *ELNEC* is available from Roy Lewallen, W7EL, P. O. Box 6658, Beaverton, OR 97007.

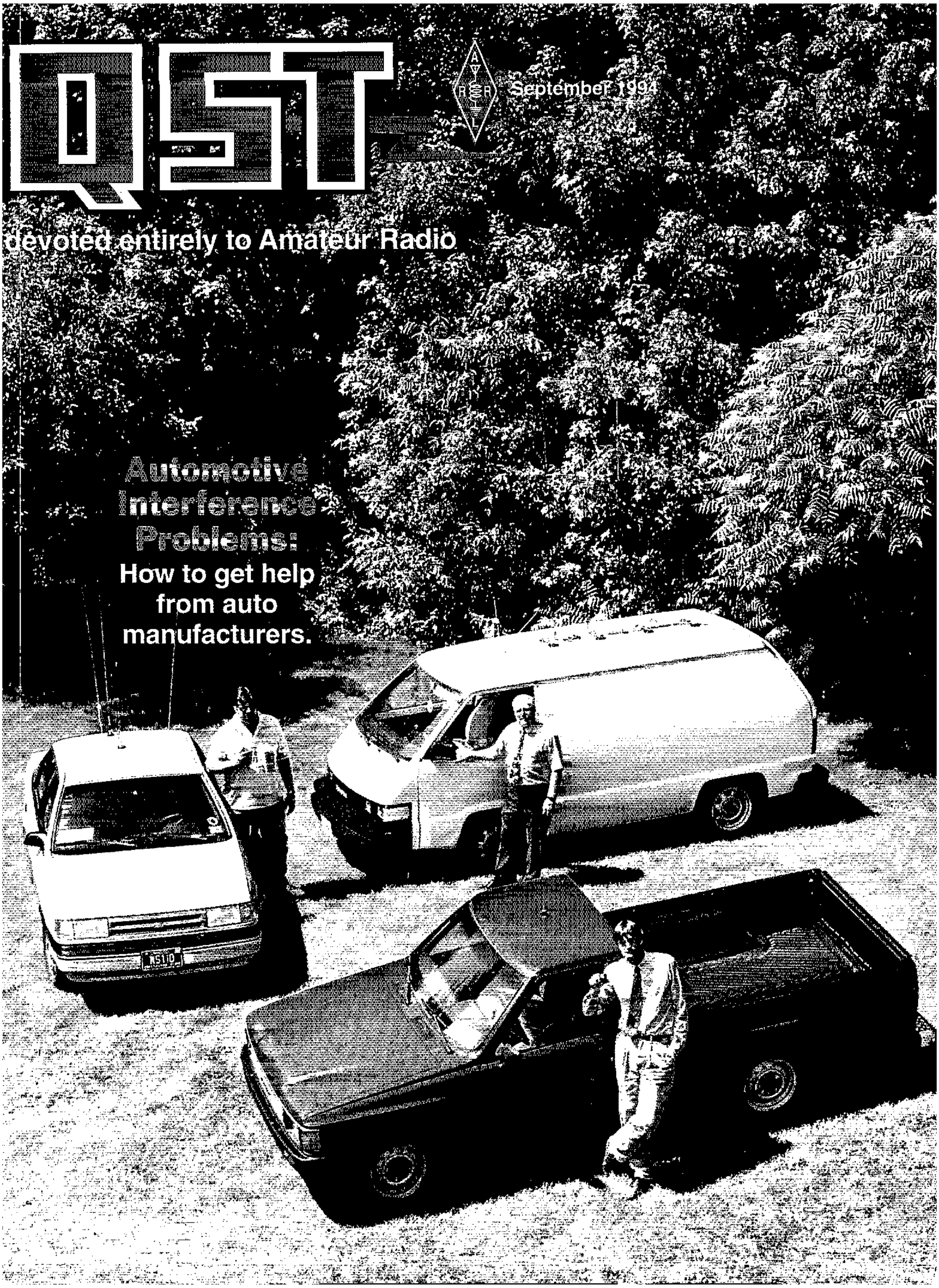
[4] G. J. Burke and A. J. Poggio, *Numerical Electromagnetics Code (NEC)—Method of Moments*, Naval Ocean Systems Center, San Diego, CA, Jan 1981.

[5] Surplus Sales of Nebraska, 1502 Jones St, Omaha, NE 68102. Tel 402-346-4750.

Al Christman was first licensed in 1974 as WA3WZD. He received his Doctorate in Electrical and Computer Engineering from Ohio University, and is Professor of Electrical Engineering at Grove City College in Western Pennsylvania. When not modeling antennas by computer for fun, Al chases SSB DX on 20 and 80 meters, or enjoys ragchewing and contesting.

Tim Duffy was first licensed as WN3SZX at the age of 12, receiving his Extra Class two years later. He graduated from Penn State University and is Director of Engineering for a communications company. Tim has a long and distinguished string of contest awards to his credit, and is a member of the ARRL Contest Advisory Committee.

Jim Breakall was first licensed in 1965 as WN3FET. Dr Breakall is a professor of Electrical Engineering at Penn State University, specializing in antenna theory and design. Like Al Christman, Jim has been very involved with the development and validation of the NEC computer modeling program. Jim is very familiar to Dayton attendees because of his well-received lectures in the Antenna Forum. He is now building a "large station in central Pennsylvania." Watch out, K3LR!



QST

September 1994

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How to get help from auto manufacturers.

QST

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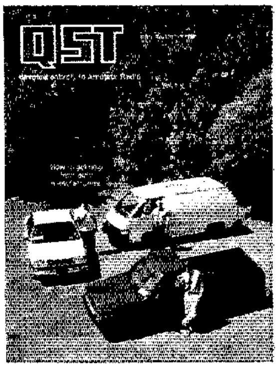


OUR COVER

Features Editor Brian Battles, WS1O (left); Field Services Manager Rick Palm, K1CE (bottom); and Managing Editor Al Brogdon, K3KMO (top) agree that mutual interference between automotive systems and ham rigs can be a royal pain! While they were on the side lawn at HQ discussing this important topic, ARRL Lab Supervisor Ed Hare, KA1CV, was inside the building, finishing his article on that same subject (page 51).

CONTENTS

September 1994
Volume LXXVIII Number 9

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

So What's New in *The ARRL Antenna Book?*

The first edition of *The ARRL Antenna Book* came off the press in 1939. Since then, more than 800,000 copies of its various editions have provided amateurs with a wealth of practical information on antennas, transmission lines and propagation. The 17th edition appeared on bookstands last month. Along with much new written material, the book includes a disk full of great software.

By R. Dean Straw, N6BV
Senior Assistant Technical Editor

I have a long-standing passion for Amateur Radio. I've been known to operate contests with abandon, but my first love in Amateur Radio has always been antennas. As a kid back in Hawaii in the 1950s, I daydreamed a lot about *really big* antennas, fantasizing about how I would crack monster pileups with a single call. At the Saturday afternoon movies, while the other kids were wishing they could be like Superman, leaping effortlessly over tall buildings, I wanted to fly over tall steel towers—putting huge antennas on top of them!

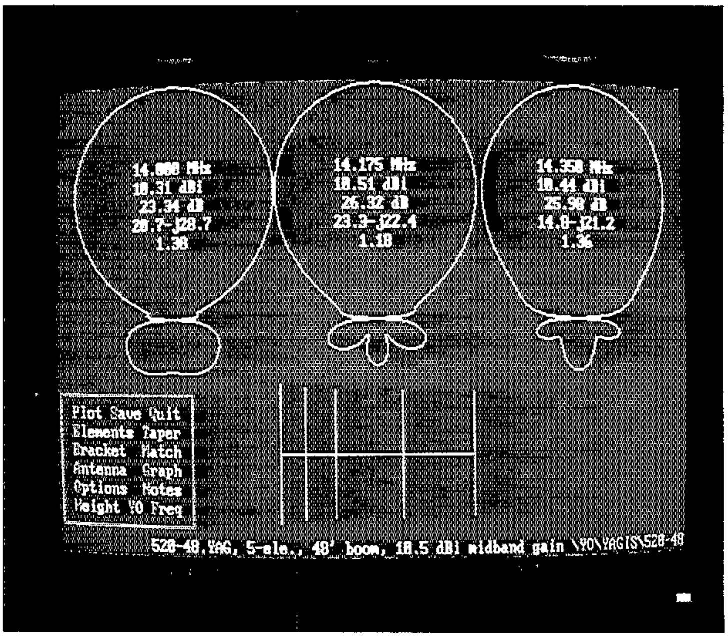
You can imagine my delight when I came to work at League HQ and was handed the assignment of working on the next edition of *The ARRL Antenna Book*, known to all amateurs simply as "the Antenna Book." Along with many other amateurs, I have always held this venerable book in high esteem, almost in reverence.

The dog-eared first edition of the *Antenna Book* in the HQ library is tiny compared to today's jumbo model. That first book contained only 142 small-format pages, while there are more than 700 full-size pages in the 17th edition. Over the last 55 years there have been some monumental changes in radio, and in most everything else in the world too. Still, I find it comforting that many sentences are very familiar in that old book. After all, some things—such as fundamental principles—*shouldn't* change with time.

Of course, many things do benefit from modern analytical tools. I approached my new task with a mixture of excitement and awe. After all, some legendary engineers and writers had labored over the last half-century putting together the knowledge in the *Antenna Book*. People like George Grammer, Byron Goodman and Jerry Hall had laid a solid foundation. My task was to use modern computing power to further illuminate those areas that could benefit from such treatment.

So, you ask, what really is new in the *Antenna Book*? The main changes came in the chapters dealing with Yagis (both HF and VHF/UHF), HF propagation, and to a smaller extent, transmission lines. Throughout the new book there are additions and modifications, but only where computer analysis could provide additional insight or clarification. For example, modern software makes it especially easy to generate detailed azimuth and elevation polar plots, directly comparing one antenna with another. When it comes to gaining a visceral understanding of how antennas really work, the old adage "one picture is better than a thousand words" accurately describes the impact of a printed pattern plot.

However, there is another, modern adage: "One computer screen can be better than 10 printed plots." The reason is simple—you can interact with properly designed software to explore, in great detail and at your own pace, particular areas that interest you most about an antenna's performance.



```
14.000 MHz     14.175 MHz     14.350 MHz
10.31 dBi      10.51 dBi      10.44 dBi
23.34 dB       26.32 dB       25.98 dB
20.7-j28.7     23.3-j22.4     14.8-j21.2
1.38           1.18           1.36

520-48.YAG, 5-ele., 48' boom, 10.5 dBi midband gain \YO\YAGIS\520-48
```

Figure 1—*YA* display for a 5-element 20-meter Yagi on a 48-foot boom in free space. *YA* shows all pertinent performance parameters for this antenna at the low end, middle frequency, and high end of the band.

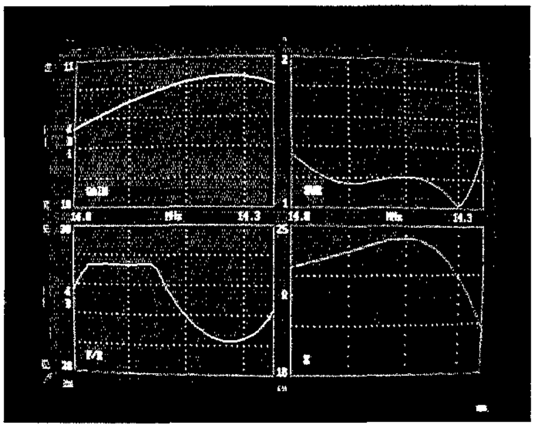


Figure 2—*YA* display across the 20-meter band for the antenna of Figure 1, showing forward gain, front-to-rear ratio, SWR, and feedpoint impedance. It is important to characterize an antenna across its entire useful frequency range to avoid severe peaks and valleys in its response.

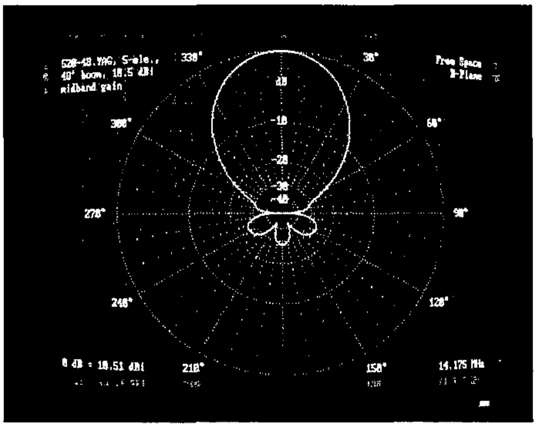


Figure 3—Detailed *YA* E-plane plot for the antenna of Figure 1. The worst-case rearward lobes are about 24 dB down at the mid-frequency point, and are at 135° and 225° in azimuth (with the main lobe at 0°).

Bundled Software in the *Antenna Book*

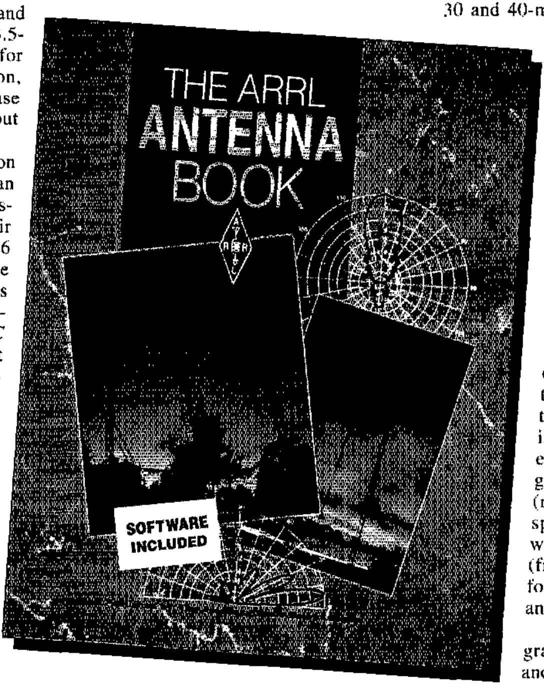
So, for the first time ever, the ARRL has bundled some really useful, interactive software in the back of the *Antenna Book.* This software is designed strictly for the IBM PC or fully compatible computer and comes on a 1.44-MB high-density 3.5-inch diskette. Included are programs for Yagi analysis, propagation prediction, and transmission-line analysis. Please note that this software is copyrighted, but not copy-protected.

The software programs will work on an older computer, such as one using an 80286 or even an (ancient) 8088 processor. However, they really strut their stuff on a more modern 80386, 80486 or Pentium-based computer. The graphics will work on older computers that have at least a CGA or Hercules display adapter (be sure to run *MS-HERC* in the \ELEVAT subdirectory first), but they look *much* better on a machine using a VGA display adapter, especially in color.

The programs will work anywhere from 10 to 20 times faster if a math coprocessor is in the system. 80486DX computers have the math coprocessor inside the main chip, as do Pentium systems. Neither the 80486SX nor the vast majority of 80286 or 80386 computers have the math coprocessor inside. Math coprocessors are pretty inexpensive nowadays and they are a *very* worthwhile investment if you intend to do any sort of serious modeling work. The quality and versatility of this ARRL-supplied software may well spur you to go ahead and upgrade your computer hardware to take full advantage!

Yagis

I mentioned that I am an HF contester—thus, almost by definition, I am interested in Yagis. Over the years Brian Beezley, K6STI, has done a commendable job creating innovative antenna-modeling software. The HF Yagi chapter in the 17th edition has a detailed discussion about the trade-offs that can be made between forward gain, peak front-to-rear ratio and SWR, all computed over a full amateur band. Much of this was derived from voluminous computations using Beezley's *YO* (Yagi Optimizer) program. *YO* was used to create a stable of more than 80 different Yagi designs, optimized to meet stringent performance criteria. There are seven large tables in the book, covering the 10, 12, 15, 17, 20, 30 and 40-meter bands, for boom lengths between 8 and 80 feet.



The data files for these 80 individual designs are located on the diskette too. Further, K6STI has provided a special program to analyze Yagis. This is called *YA*, for "Yagi Analyzer," a slick bit of programming!

Figure 1 shows the main screen for a 5-element 20-meter Yagi on a 48-foot boom in free space. The response at the low end, middle and high end of the 20-meter band is shown in the three on-screen patterns for this antenna. The second line (under the frequency for each pattern) is the forward gain, in dBi, referenced to isotropic, although gain may also be shown in dBd (referenced to a dipole in free space). The third line is the worst-case peak rearward lobe (front-to-rear ratio), in dB. The fourth line is the feedpoint impedance and the last line is the SWR.

YA can generate a set of four graphs showing the performance across the whole band. See Figure 2, created for the same 20-meter Yagi. For this particular design, the gain rises slightly over the whole frequency band, while the worst-case front-to-rear ratio (F/R) stays better than 20 dB over the whole band, rising to a peak level of 24.4 dB at about 14.150 MHz. The SWR

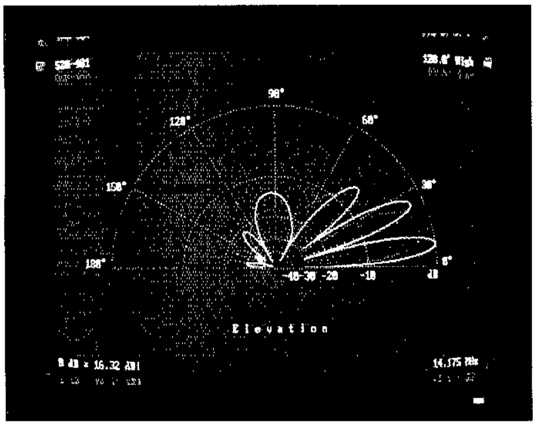
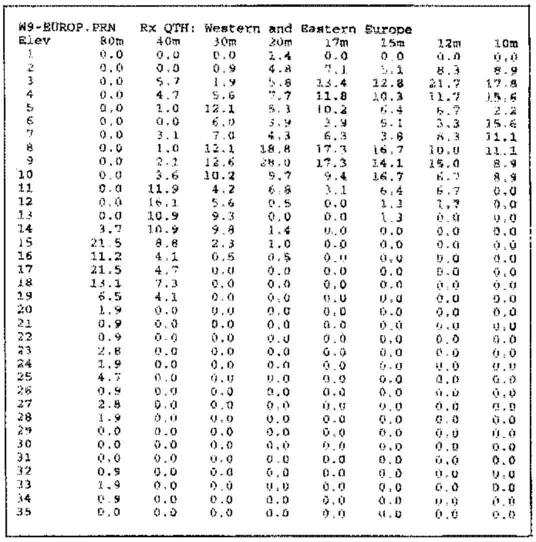


Figure 4—Polar elevation plot by *YA*, showing the effect of height on the pattern. If two antennas of the type illustrated in Figure 1 were placed on a tower 69 feet and 120 feet above flat ground, the null in the pattern of the higher antenna at 15° elevation will be covered nicely by the lower antenna, as would be expected.



```
W9-EUROP.PRN    Rx QTH: Western and Eastern Europe
Elev       80m     40m     30m     20m     17m     15m     12m     10m
 1         0.0     0.0     0.0     1.4     0.0     0.0     0.0     0.0
 2         0.0     0.0     0.9     4.8     7.1     5.1     8.3     8.9
 3         0.0     5.7     1.9     5.8    13.4    12.8    21.7    17.8
 4         0.0     4.7     5.6     7.7    11.8    10.3    11.7    15.6
 5         0.0     1.0    12.1     5.3    10.2     6.4     6.7     2.2
 6         0.0     0.0     6.0     3.9     3.9     5.1     3.3    15.6
 7         0.0     3.1     7.0     4.3     6.3     3.8     8.3    11.1
 8         0.0     1.0    12.1    18.8    17.3    16.7    10.0    11.1
 9         0.0     2.1    12.6    28.7    17.3    14.1    15.0     8.9
10         0.0     3.6    10.2     9.7     9.4    16.7     6.7     8.9
11         0.0    11.9     4.2     6.8     3.1     6.4     6.7     0.0
12         0.0    16.1     5.6     0.5     0.0     1.3     1.7     0.0
13         0.0    10.9     9.3     0.0     0.0     1.3     0.0     0.0
14         3.7    10.9     9.8     1.4     0.0     0.0     0.0     0.0
15        21.5     8.8     2.3     1.0     0.0     0.0     0.0     0.0
16        11.2     4.1     0.5     0.5     0.0     0.0     0.0     0.0
17        21.5     4.7     0.0     0.0     0.0     0.0     0.0     0.0
18        13.1     7.3     0.0     0.0     0.0     0.0     0.0     0.0
19         6.5     4.1     0.0     0.0     0.0     0.0     0.0     0.0
20         1.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
21         0.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
22         0.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
23         2.8     0.0     0.0     0.0     0.0     0.0     0.0     0.0
24         1.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
25         4.7     0.0     0.0     0.0     0.0     0.0     0.0     0.0
26         0.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
27         2.8     0.0     0.0     0.0     0.0     0.0     0.0     0.0
28         1.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
29         0.0     0.0     0.0     0.0     0.0     0.0     0.0     0.0
30         0.0     0.0     0.0     0.0     0.0     0.0     0.0     0.0
31         0.0     0.0     0.0     0.0     0.0     0.0     0.0     0.0
32         0.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
33         1.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
34         0.9     0.0     0.0     0.0     0.0     0.0     0.0     0.0
35         0.0     0.0     0.0     0.0     0.0     0.0     0.0     0.0
```

Figure 5—A portion of the data file from the *Antenna Book* diskette that shows the percentages of time versus angle when each frequency band is open from the Chicago area to all of Europe. For example, on 20 meters the peak percentage is 28.7%, at a takeoff angle of 9°.

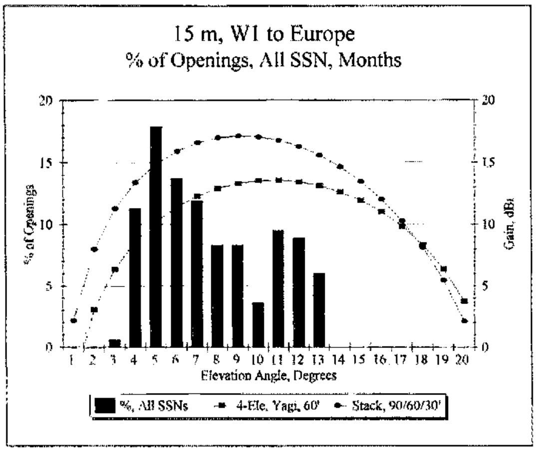
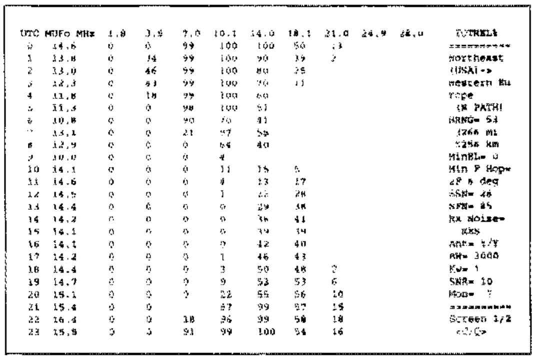


Figure 6—A graph from the *Antenna Book* that shows an overlay of computer-generated elevation patterns for two different antenna systems, compared to the statistical histogram of the percentage of all openings versus elevation angle for the 15-meter band from New England to Europe. One antenna is a 4-element Yagi 60 feet high, and the second is a stack of such Yagis at 30, 60 and 90 feet. The stack can be expected to outperform the single antenna by at least 5 dB at the peak statistical elevation angle of 5°.



```
UTC MUFo MHz  1.8   3.5   7.0  10.1  14.0  18.1  21.0  24.9  28.0    TOTREL%
 0   14.6     0     0    99   100   100    50    13                ==========
 1   13.8     0    14    99   100    90    39     2                Northeast
 2   13.0     0    46    99   100    80    25                      (USA)->
 3   12.3     0    43    99   100    70    11                      Western Eu
 4   11.8     0    18    99   100    60                            rope
 5   11.3     0     0    98   100    51                             (S PATH)
 6   10.8     0     0    90    70    41                            BRNG= 53
 7   13.1     0     0    21    57    59                             3266 mi
 8   12.9     0     0     0    64    40                             5256 km
 9   10.0     0     0     0     4                                  MinBL= 0
10   14.1     0     0     0    11    15     5                      Min F Hop=
11   14.6     0     0     0     4    13    17                      2F 5 deg
12   14.5     0     0     0     1    22    28                      SSN= 26
13   14.4     0     6     0     0    29    38                      SFI= 85
14   14.2     0     0     0     0    36    41                      Rx Noise=
15   14.1     0     0     0     0    39    39                        RES
16   14.1     0     0     0     0    42    40                      Ant= 1/Y
17   14.2     0     0     0     1    46    43                      BW= 3000
18   14.4     0     0     0     3    50    48     2                Kw= 1
19   14.7     0     0     0     9    53    53     6                SNR= 10
20   15.1     0     0     0    22    55    56    10                Mon= 7
21   15.4     0     0          67    99    57    15                ==========
22   16.4     0     0    18    96    99    58    18                Screen 1/2
23   15.9     0     5    91    99   100    54    16                 <1/2>
```

Figure 7—The total reliability screen from the *ION* propagation-prediction software. This is for the path from New England to Western Europe for the month of July, with a solar flux level of 85. The 30-meter band would be the most consistent performer on this path during the late afternoon and early evening.

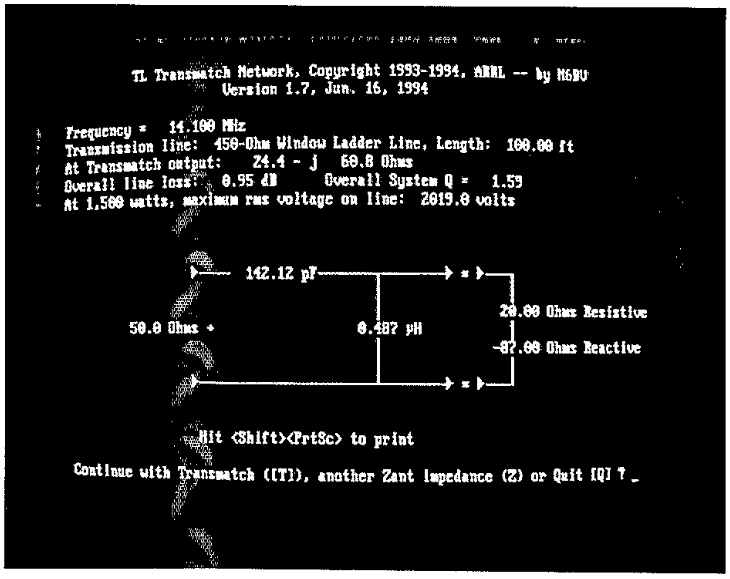
stays below 2:1 over the whole band also, while the magnitude of the feedpoint impedance gradually falls as the frequency is raised.

The patterns may be seen in more detail in *YA* by invoking the "Plot" function. Figure 3 shows a detailed free-space E-plane plot of the 5-element Yagi we have been considering at 14.175 MHz. Figure 4 shows the pattern of this antenna at 69 feet, compared with its pattern at 120 feet. You can see that the two heights are complementary—where the high antenna has a null about 15°, the lower antenna has a peak.

Propagation

Another subject near and dear to the heart of the amateur operator is propagation. HF DXers and contesters are particularly attuned to the vagaries of the ionosphere. The elevation angles needed to launch signals toward distant QTHs has been for many years a subject of considerable interest, but there has been precious little hard data available for the amateur. In the mid 1980s in Northern California, I well remember struggling to find elevation angle information when I was designing the 80-meter quad arrays at the N6RO contest station. I resorted to extrapolating from the only source of information available to amateurs at that time—the single table labeled "Measured Vertical Angles of Arrival of Signals from England at Receiving Location in New Jersey," which had been in the *Antenna Book* since the early 1970s.

It turns out that the information in that table (which covered only 7 to 28 MHz) was derived from measurements taken by Bell Labs in 1934. That was a year, incidentally, when the sunspot cycle was at its very lowest. The likelihood that this information measured for England to New Jersey was also valid for a station about 3000 miles to the west was pretty slim. But as I said, it was the only information around!



```
TL Transmatch Network, Copyright 1993-1994, ARRL -- by N6BV
Version 1.7, Jun. 16, 1994

Frequency =   14.100 MHz
Transmission line:  450-Ohm Window Ladder Line, Length:  100.00 ft
At Transmatch output:    24.4 - j   60.8 Ohms
Overall line loss:   0.95 dB      Overall System Q =   1.59
At 1,500 watts, maximum rms voltage on line:  2019.0 volts
```

```
Hit <Shift><PrtSc> to print

Continue with Transmatch ([T]), another Zant Impedance (Z) or Quit [Q] ? _
```

Figure 8—Computed matching network from *TL* (transmission line) analysis program for 100 feet of RG-8 or RG-213 coaxial cable at 14.1 MHz, with a load of 25 – *j* 40 Ω. *TL* computes a number of parameters for a variety of types of transmission lines at frequencies up to 5 GHz.

So in preparing material for the *Antenna Book*, once again I turned to the computer. I used a very sophisticated ionospheric-modeling program called *IONCAP*. This program has been under almost continuous development over the last 30 years by various agencies of the US government. It is used here at HQ each month to generate the propagation graphs in *QST*'s How's DX column. I ran thousands of computations, creating huge databases from which statistically significant data could be extracted about elevation angles for locations in the US to important DX locations throughout the rest of the world.

The *Antenna Book* itself contains detailed summaries of all this data, customized for all 10 US geographic call areas, W1 through W0, to six important areas of the world: Europe, the Far East, South America, Southern Africa, South Asia, and the South Pacific. On the disk, however, are 60 ASCII files, which show considerably more detail.

A sample from one of the 60 files, from Chicago to all of Europe, is shown in Figure 5. The file covers all levels of solar activity and all hours, months and years. For each amateur HF band and at each takeoff angle up to 35°, the percentage of all possible openings is shown. For example, on 40 meters, signals arrive (or depart) at a 12° elevation angle 16.1% of all the time the 40-meter band is open to Europe from Chicago. On 10 meters, the peak percentage occurs at 3° for 17.8% of all times this band is open. There is a second peak angle at 6° elevation, occurring 15.6% of the time.

Figure 6 shows a graph from the 17th edition. Statistical propagation data for New England to Europe has been overlaid with elevation pattern responses created by the *NEC* computer program. This shows the 15-meter percentages versus elevation angle, overlaid with the elevation response for a single 60-foot-high 4-element Yagi, and for a stack of three such Yagis at 90, 60 and 30 feet height over flat ground. Such information should help you plan the heights at which to place your antennas.

Besides the wealth of statistical elevation-angle data, the *Antenna Book* disk also includes a nifty propagation-prediction program by noted software author Jake Handwerker, W1FM. The software is called *ION_HDX*, or *ION* for short. With *ION*, you have the ability to make short-term forecasts of what the HF bands should be doing, for any given level of solar flux in any month and year. Radio station WWV broadcasts solar flux numbers regularly, and many amateur packet networks list them on request.

Figure 7 shows the total reliability predicted by *ION* from Boston to Western Europe for July 1994, with an assumed solar flux of 85. This prediction assumes Yagi antennas at each end of the circuit, 3000-Hz receiver bandwidth, 1-kW transmitter power, and a required signal-to-noise ratio (S/N) of 10 dB. For this relatively low level of solar activity in midsummer, the 15-meter band is not very likely to be open to Europe, even from the East Coast! There are corresponding screens for predicted signal level, angle of radiation, modes of propagation, and other parameters of interest. The *ION* program is very full-featured and useful, indeed.

Other Software

Another useful program on the diskette included with the *Antenna Book* is called *TL*, for "Transmission Line." *TL* can be thought of as a sort of Smith Chart without the chart. *TL* can analyze any length of any type of transmission line, terminated with any desired impedance, at any frequency up to 5 GHz. It computes not only the impedance at the input of a line, but also the overall loss, the SWR at input and output, and the maximum voltage on the line for 1.5 kW of power. It can generate an on-screen schematic of a network to match the impedance at the input of a line to 50 Ω.

See Figure 8 for an example, where 100 feet of RG-213 is operating at 14.1 MHz with a termination of 25 Ω in series with a capacitive reactance of 40 Ω. The overall loss in the transmission line at this frequency is 1.34 dB, and there is a maximum of 417.9 V on the line. *TL* is useful to analyze a variety of transmission-line problems.

Another useful program for working with Yagi antennas is called *SCALE*, which allows the operator to scale a *YA* Yagi design to another frequency, another taper schedule (for telescoping tubing), or for input to another program, such as *NEC2* or *MN*.

The Bottom Line

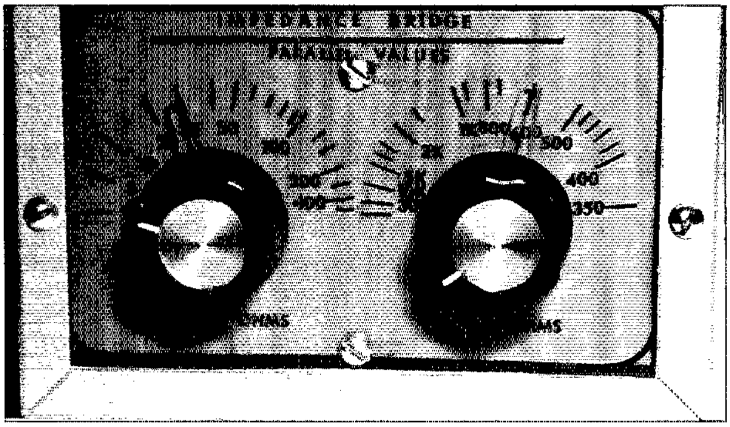
Many amateurs have told me how delighted they are with the software bundled with the 17th edition of *The ARRL Antenna Book*. Several have commented that they would gladly have paid the full price for just a single program like *YA* or *ION*. And they also get a 700-page book and a multitude of other useful programs! Other hams, who are not so deeply into computers, have told me that the printed book is still worth every penny it has been for almost 60 years now. I would love to hear further feedback, negative as well as positive, from readers.

[Software updates are available on HQ's Hiram BBS at 203-666-0578.—*Ed.*]

A Reevaluation of the Caron RF Impedance Bridge

When you encounter capacitive reactance with this bridge, a bit of fine tuning is required to obtain accurate results.

By Charles Camillo, W6TGK
6150 Locust Ln
La Grange, IL 60525



When making antenna impedance measurements at the end of a long transmission line (usually from within the shack) it's always necessary to translate the measured impedance back to the antenna feed point, either by the use of a Smith Chart, or any of a number of computer programs designed for this purpose.[1] Because the measured impedance at any point on the transmission line can be very different from the antenna impedance, we must be prepared to measure a wide range of resistance and reactance. This is especially true if the measurements are made over a range of frequencies above and below the resonance point of the antenna. We must also be able to accurately determine the cable's electrical length. My favorite method of measuring the electrical cable length is to measure its input impedance with a short at the antenna end, if it is easily accessible. Because many typical noise bridges[2] used for this work are poorly calibrated and are noisy at the low end of the resistance scale, one can be misled by substantial errors. With John Grebenkemper's noise bridge modifications,[3] it's possible to improve the low-resistance accuracy for most measurements. However, we're still left with the noisy resistor and the frustration of adjusting the dials for a good balance without going past the point of best balance because of the directly driven shafts.

After several years of working with my noise bridge, I felt I needed something better, yet cheaper than a laboratory bridge costing hundreds of dollars. *The ARRL Antenna Book*, 15th Edition, describes an RF impedance bridge[4] that can be used to read resistance values as low as several ohms and uses a capacitor to balance the unknown resistance (see Figure 1). This circuit caught my interest several years ago and I built one. In order to satisfy my desire for a smooth and easy adjustment of the resistance dial—and at the same time improve the resolution of the readings—I used a National Velvet Vernier drive and dial on that shaft.

Resistance calibration was done using 1%-tolerance metal-film resistors obtained from Radio Shack (part number 271-309). Using a digital multimeter, I selected values of 5.8, 10.5, 46.6, 100 and 470 Ω. I soldered the resistors into PL-259 connectors in much the same way as described by Grebenkemper.

I calibrated the reactance dial as detailed by Wilfred Caron in *The ARRL Antenna Book*, simply by calculating the reactance at 1 MHz for various values of capacitance, assuming a straight-line relationship be-

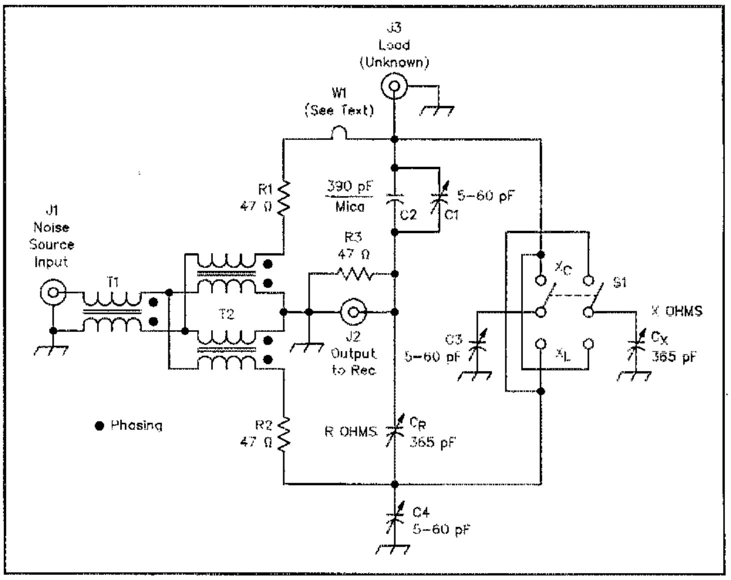


Figure 1—Schematic of the Caron Parallel Impedance Bridge described in *The ARRL Antenna Book*, 15th edition, pp 27-19 to 27-22. Refer to the original text for a complete explanation of the circuit.

[1] Notes appear on page 31.

tween capacitance and shaft rotation, using the maximum and minimum values of the capacitor to determine the graph line. Being a skeptic, I wanted to check the accuracy of the reactance dial calibrated this way.

To accomplish this, I made up more loads consisting of various parallel resistor-capacitor combinations using PL-259 connectors, mica capacitors and suitable resistors. These checks indicated the bridge was reading low by a factor of approximately 2 to 1—more in some cases—as shown in Table 1. Naturally, my first conclusion was I had made a wiring error or the parts arrangement was causing a substantial amount of stray capacitance to be added in parallel across C_X (see Figure 1). When I found no errors, I built a second, smaller unit, to keep wiring and wire lengths to a minimum.

This version gave the same results. Now I began to suspect the loads were at fault, either because the PL-259s were introducing excessive reactance, or the mica capacitors were not as good as expected, which wasn't very plausible. At this point, I dropped the project. I purchased a good laboratory standard bridge—a General Radio 916A RF Impedance Bridge[5]—and got caught up experimenting with it. My thoughts never returned to the parallel-impedance bridges until one day when a friend asked how they had worked out. I then decided to determine what the problem was.

Calibrating the Loads

The GR-916A RF impedance bridge is a classic four-arm Wheatstone bridge that measures the equivalent series impedance of the unknown. Calibrating the loads required measuring the equivalent series impedance and converting the results to parallel values using the equations:[6]

$$R_P = \left(X_S^2 + R_S^2\right) \div R_S \qquad \text{(Eq 1)}$$

$$X_P = \left(R_S^2 + X_S^2\right) \div X_S \qquad \text{(Eq 2)}$$

When I had made my original capacitor/resistor loads, I chose four values of capacitance and 33 or 47-Ω resistors for the combinations. (As you'll see later, the choice of low-value resistors was a serendipitous event.) The results of the measurements are shown in Table 2. There is close agreement in the resistance values obtained with all methods, but the reactance values—and thus the capacitance values—of the Caron bridge are substantially different from those obtained using the GR bridge and the multimeter. Because the multimeter and the GR bridge were in fairly close agreement, I was forced to assume a circuit or calibration error. Before going any further, I made another thorough check of the wiring, solder connections and grounds. Then, since some time had passed—and I assumed I wouldn't repeat any calibration errors of the past—I performed the calibration again—from scratch. A repeat of the data taken in Table 1 indicated no change. At this point, I decided it was necessary to review the operation and theory of the bridge. Then I might have a better chance of understanding the source of the errors.

The ARRL Antenna Book article shows the balance conditions in the simplified form using only C_R and a resistive load. There is no equation or mathematical relationship covering the case when the load is reactive and C_X is included in the circuit, only the statement that C_X tunes out the reactance. Although the rigorous mathematical analysis of circuits is not one of my strong points, I decided to see if I could develop the proper equations.

I began by deriving, from scratch, the balance equation for the simplified resistive load, to see if I could duplicate the results of *The ARRL Antenna Book*.

Purely Resistive Loads

The simplified schematic in Figure 4A shows the current paths for three complete circuits or meshes.[7] These current paths yield three circuit equations:

$$e = (I_1 + I_2)R + I_1 R_u \qquad \text{(Eq 3)}$$

$$e = (I_1 + I_2)R + I_2 X_1 + (I_2 - I_3)R_D \qquad \text{(Eq 4)}$$

$$e = I_3 R + I_3 X_R + (I_3 - I_2)R_D \qquad \text{(Eq 5)}$$

where R_u is the unknown resistance and $R_D = R$.

At balance, $I_3 = I_2$ to give a null in the detector R_D. To solve for R_u in terms of X_R, we equate Equations 3 and 5, combine terms and rearrange to get:

$$I_1 R + I_1 R_u = I_2 R \qquad \text{(Eq 6)}$$

Because the voltage drops across R_u and X_1 must be equal to null the current in R_D, then, $I_1 R_u = I_2 X_1$

$$I_1 R_u = I_2 X_1 \qquad \text{(Eq 7)}$$

and,

$$I_2 = \frac{R_u}{X_1} I_1 \qquad \text{(Eq 8)}$$

Substituting Equation 8 in Equation 6:

$$I_1 R + I_1 R_u = R_u \frac{X_R}{X_1} I_1 \qquad \text{(Eq 9)}$$

which simplifies to

$$R_u = \frac{RX_1}{X_R - 1} \qquad \text{(Eq 10)}$$

Table 1

Measurement Results of Calibration Loads Using the Caron Bridge at 3 MHz

Marked Values		*Caron Bridge*			
Resistance (ohms)	*Capacitance (pF)*	*Parallel Z_u*	*R_p (ohms)*	*C_p (pF)*	*Ratio C_p/C_{marked}*
33	47	34-*j*4800/3	34	33.2	0.45
35	320	35-*j*1050/3	35	151.7	0.47
47	470	49-*j*510/3	49	261.0	0.56
33	820	35-*j*430/3	35	370.3	0.45

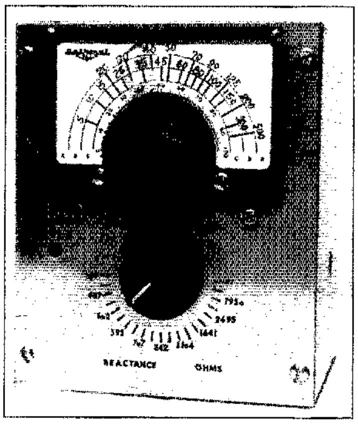


Figure 2—My first model of the Caron bridge uses a National vernier drive equipped with a logging scale. My second—more compact—version of Caron's bridge is shown in the title photo.

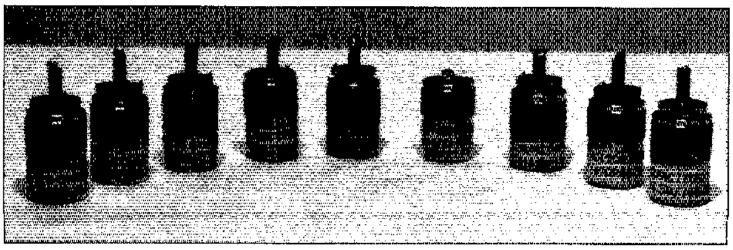


Figure 3—Various loads used to calibrate the bridge.

This converts to

$$R_u = \frac{R}{\frac{X_R}{X_1} - 1} \quad \text{(Eq 11)}$$

Since Equation 11 is identical to the one in *The ARRL Antenna Book*, I proceeded to the next step, adding an unknown reactance and the capacitor, C_X.

Complex Impedance Load

Assuming the load is inductive, S1 places C_X directly across the load, which puts C_X and L_u in parallel with R_u. Working out the parallel combination of the three elements, we find that C_X cancels the unknown inductance, L_u. Thus, this case presents no problems.

In the capacitive case (Figure 4B), I have redrawn the bridge circuit in order to simplify the analysis. The addition of the one connecting wire from the junction of Z_u and C_X to the center tap of the matching transformer means we cannot assume the currents in Z_u and C_X are equal.

From the circuit, we can again write three mesh equations as:

$$(I_1 + I_3)R + I_1 Z_u = e \quad \text{(Eq 12)}$$

$$(I_2 + I_3)R + I_2 X_x = e \quad \text{(Eq 13)}$$

$$(I_1 + I_3)R + (I_2 + I_3)R + (X_1 + X_R)\,I_3 = 2e \quad \text{(Eq 14)}$$

To solve for Z_u, we equate Equations 12 and 13:

$$I_1\,(R + Z_u) = I_2(R + X_x) \quad \text{(Eq 15)}$$

Also, at balance,

$$\frac{I_1 Z_u}{I_2 X_x} = \frac{I_3 X_1}{I_3 X_R};\; Z_u = \frac{X_x X_1 I_2}{X_R\, I_1} \quad \text{(Eq 16)}$$

Solving Equation 15 for I_2/I_1 and substituting the answer into Equation 16, gives

$$Z_u = \frac{X_1 X_x\,(R+Z_u)}{X_R\,(R+X_x)} = \frac{X_1 X_x R}{RX_R + X_R X_x - X_1 X_x} \quad \text{(Eq 17)}$$

Again, substituting the reactance equations

$$X_1 = \frac{1}{j\omega C_1};\; X_x = \frac{1}{j\omega C_x};\; X_R = \frac{1}{j\omega C_R} \quad \text{(Eq 18)}$$

into Equation 17 gives

$$Z_u = \frac{RC_R}{(C_1 - C_R) - j\omega RC_1 C_x} \quad \text{(Eq 19)}$$

Since Z_u is a parallel combination of resistance and capacitance, then Z_u can also be expressed as:

$$Z_u = \frac{R_u\left(\frac{1}{j\omega C_u}\right)}{R + \frac{1}{j\omega C_u}} = \frac{R_u}{1 + j\omega R_u C_u} \quad \text{(Eq 20)}$$

Equating Equations 19 and 20, and simplifying, we get

$$R_u(C_1 - C_R) + j\omega\, R_u\, RC_1\, C_x = RC_R + j\omega\, R_u\, RC_u\, C_R \quad \text{(Eq 21)}$$

Separating real and imaginary parts on each side and equating them gives the two balance equations necessary to fully characterize the unknown impedance,

$$R_u = \frac{RC_R}{C_1 - C_R} \quad \text{(Eq 22)}$$

$$C_u = \frac{C_1}{C_R} C_x \quad \text{(Eq 23)}$$

As you can see, Equation 22 is identical to the original—resistance only—balance equation. This is a considerable encouragement to trust the validity of Equation 23, which says that the indicated value of the reactive component is reduced by the ratio of C_1 divided by C_R. Since the parallel combination of C_1 and C_2 is fixed at about 400 pF, a small value of C_R (which is determined by the load resistance) leads to a substantial reduction in the indicated reactive component. This is exactly what I got as a result of my choice of the small values of resistance for the loads. Had I used a 470-Ω resistor, I might never have seen the problem! The C_R capacitor would have been very nearly at its maximum value, giving a value of approximately 1 for the ratio of C_1 to C_R.

Recalibration and Checking the New Equation

The final step in this analysis was to test the new balance equation to determine the

Table 2

Data of R_p and C_p from Various Measurement Methods at 3 MHz

Marked Values		*Multimeter*		*GR-916A Bridge*		*Caron Bridges First Model*		*Caron Bridges Second Model*	
R (ohms)	*C (pF)*	*R (ohms)*	*C (pF)*	*R_p (ohms)*	*C_p (pF)*	*R_f (ohms)*	*C_p (pF)*	*R_p (ohms)*	*C_p (pF)*
33	47	32.8	46	32.5	41.4*	37	20.3	32	33.2
35	320	35.0	318	34.9	315	39	145	36	152
47	470	46.6	480	46.5	489	50	249	49	261
33	820	32.6	822	32.5	768	37	339	32	370

*Used the lead reactance correction methods described in the General Radio 916A Bridge manual as the reactance for this capacitor is almost at the zero point on the scale.

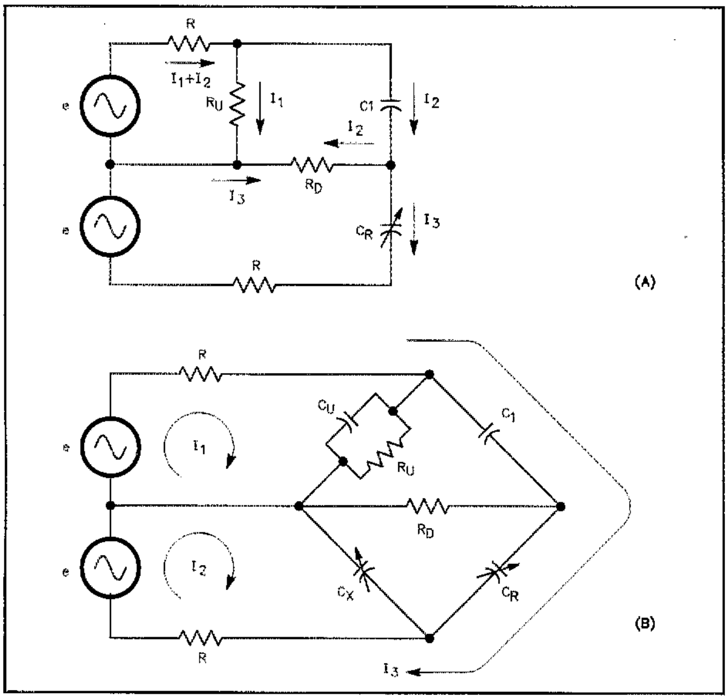


Figure 4—Simplified bridge circuits (see text).

Table 3

Data of Final Bridge Measurements with C_1/C_R Corrections Applied (at 3 MHz)

Load (ohms/pF)	R_P	X_X	LOG	C_R	C_1/C_R	C_X	C_P	STD	Devn %
33 47	37	7950/3	45.0	173	2.312	20.0	46.2	41.4	+11.7
35 320	40	1104/3	47.5	183	2.186	144.2	315.2	315.1	+0.03
47 470	50	651/3	53.5	208	1.923	244.6	470.4	489.0	−3.8
33 820	37	470/3	45.0	173	2.312	338.8	783.3	768.2	+2.0

proper reactance of the loads. But before I could do that, I needed a means of determining the value of C_R for any measured value of R_u on the bridge dial. I was fortunate that my first bridge model with the National vernier dial is equipped with a logging scale. I measured the capacitance versus the logging scale using a digital capacitance meter, after temporarily disconnecting the capacitor from the circuit. The short leads to the meter added no more than 1 pF to the measurements, so that was neglected.

Now, I could measure X_u and R_u and determine C_R from the logging scale and the graph. I remeasured all four loads and made a comparison chart, Table 3, which clearly confirms the validity of the correction to the reactance readings. These new values obtained for the unknown capacitance may be even more accurate than the General Radio bridge, the accuracy of which is diminished at the low end of the reactance dial. At least, this seems to be the case for the 47-pF capacitor where the Caron bridge value is closer to the digital capacitance meter than the GR bridge. Thus, the new equation seems to work well, as it gives the correct values for the loads.

Summary

This analysis shows that the Caron bridge can be used without corrections—as long as you're measuring a pure resistance or a complex load consisting of resistance and inductance. When you encounter a capacitive reactance, you *must* use the correction factor presented here to properly establish the reactance of the unknown load.

Although the new procedure is more cumbersome than simply taking two readings directly from the instrument scales, the parallel bridge should be evaluated in terms of its original purpose: to measure—with a higher degree of precision and accuracy—small values of impedance (both resistance and reactance are small). As stated at the outset, our measurements frequently require us to deal with these low values, and it seems appropriate to have another means of checking the accuracy of our measurements. The smoothness and lack of noise in the resistance dial is much preferred to the potentiometer in the typical series impedance reading noise bridges. With the addition of the reduction drives on both shafts, this instrument is a pleasure to use.

Finally, we're left with two choices in calibrating the R dial:

- Draw a graph of R and C_R versus an arbitrary logging scale and use the graph with only one reading taken from the C_R dial, or,
- Calibrate the C_R dial with two scales, one for R and a second for C_R.

Even though you have to take three readings for each point, it seems a small price to pay considering the advantages of using this device to measure network or antenna impedances.

Notes

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[1] W. Hayward, *ARRL Microsmith*, V2.0 (available from the ARRL, 225 Main St, Newington, CT 06111-1494, tel 203-666-1541; 5.25-inch diskette, no. 4076; 3.5-inch diskette, no. 4084; price $39 plus shipping. Connecticut residents add 6% state sales tax; Canadians add 7% GST.)

[2] "A Noise Bridge For 1.8 Through 30 MHz," *The ARRL Antenna Book*, 17th edition, 1994, pp 27-23 to 27-26.

[3] J. Grebenkemper, "Improving And Using R-X Noise Bridges," *QST*, Aug 1989, pp 27-32.

[4] W. Caron, "An Accurate RF Impedance Bridge," *The ARRL Antenna Book*, 15th edition, 1988, pp 27-19 to 27-22.

[5] D. Sinclair, "A New R-F Bridge for Use at Frequencies up to 60 Mc," *General Radio Experimenter*, Vol. XVII, No. 3, Aug 1942.

[6] *The ARRL Antenna Book*, 15th edition, p 27-19.

[7] W. Hayt, J. Kemmerly, *Engineering Circuit Analysis*, (New York: McGraw-Hill Book Co, Inc. 1962).

Charles Camillo, W6TGK, was first licensed in 1949 as W9GZJ. He has also held the calls W1RBO and W9ISV. Charles received his degree in engineering physics from the University of Illinois in 1948. After graduation, he joined the Research Department of Amphenol. In 1956, Charles was one of the first to recognize the advantages of foamed polyethylene as a cable dielectric and developed the first cables using this material.

In 1962, Charles was appointed the US technical expert on RF connectors and cables for the International Electro-technical Commission meetings in Switzerland.

Charles became the General Manager of Amphenol's Cable and Wire Division in Chicago, where he remained until 1970.

From 1971 until his retirement, Charles was with the electronics group of Illinois Tool Works.

Currently, Charles is retired and enjoying experimenting with wire antenna configurations and measurements.

New Products

ATV SCAN CONVERTER, TNC

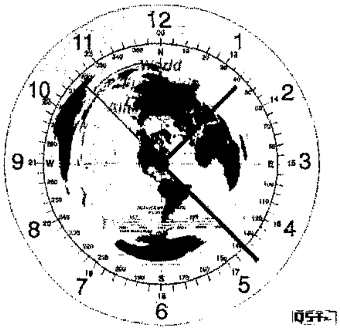
◊ The AOR TSC-100 Still Image Color Scan Converter uses advanced digital signal-processing (DSP) technology to operate amateur television (color SSTV) in a standalone configuration, with no computer necessary. It's fully compatible with Robot color 36- and 72-second, and Amiga AVT 90- and 94-second modes. Its compact size and 12-V dc (500 mA) power requirement make it practical for portable and mobile operation. It has standard NTSC outputs and connects to most image sources: camcorders, 8-mm video sources, Composer, VTR, laserdisk and more. The TSC-100 interfaces to almost any analog-grade voice radio or audio circuit, and has a built-in microphone relay and monitor speaker. The compact unit measures just $8\frac{1}{2} \times 5\frac{1}{2} \times 1\frac{1}{2}$ inches and weighs $1\frac{1}{2}$ pounds. Retail price is $749.95

The AR-210 is a compact ($2\frac{1}{2} \times 1 \times 1\frac{1}{4}$ inches), lightweight (3.3-ounce), TAPR TNC-2-compatible terminal node controller (TNC) for operating VHF packet radio. It features a built-in real-time clock with lithium battery back-up, an internal mailbox system, enhanced command set, diagnostic calibration, TCP/IP-capable KISS mode and runs at 1200 bauds (AFSK). The AR-210 runs on 10 to 13.8 V dc at just 200 mA, and can be equipped with an optional NiCd battery with quick charge circuit. The serial port is a standard RS-232C interface with a DB-9 connector, and the radio connections use its mini-phone jacks for hand-held rigs or a DIN-type jack for other radios. Retail price is $229.95.

For information contact EDCO, 325 Mill St, Vienna, VA 22180; tel 703-938-8105, fax 703-938-4525.

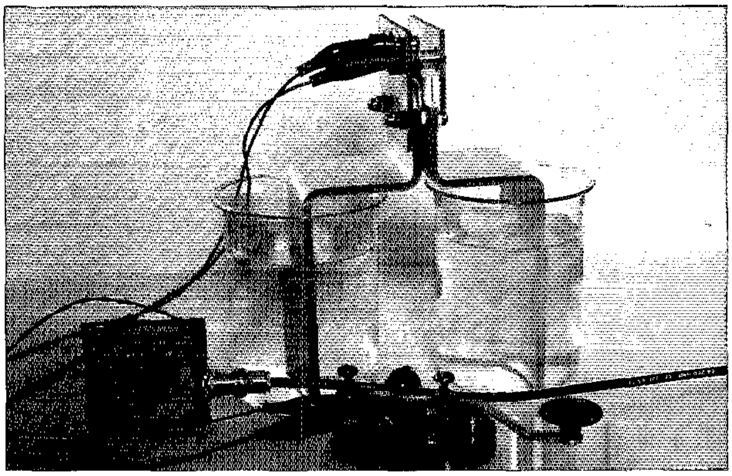
RADIO AMATEUR WORLD CLOCK

◊ Tim Thirst, G4CTT, has introduced an attractive, colorful analog-dial quartz clock for radio amateurs that features a global map centered on your "home" and true compass bearings from the center to any point in the world. Order with your location specified and your clock is customized with the map face centered on your home. (Choices include Eastern, Central and Western North America, Eastern and western Europe, Australia, the Far East and more. The scale of miles/kilometers on the face helps to measure the distance from the center to a DX country. The clock is marked in 12- and 24-hour formats, with compass readings in 10° increments. The 3-color, 9-inch face is easy to read from across the shack. An AA battery provides at least a year of operation. Retail price £29.95 Sterling (about $46 US). Eastern Communications, Cavendish House, Happisburgh, Norfolk, England; tel 44-692-650-077, fax 44-692-650-925.



Thermoelectric Power for QRP Transmitters

By Arnold Sayre, W8WVM
Route 7, Box 14
Buckhannon, WV 26201
(Photos by the author)



Peltier-device power provides plenty of potential to permit QRP operation.

Supplying Amateur Radio equipment with power generated by solar panels or wind-operated devices has become somewhat commonplace over the last few years. More recently, vegetables and fruit ("green machines?"—*Ed.*) have been uniquely tapped as power sources. Remember the lemon-powered QRP transmitter?[1] Now you can be the first kid on your block to power a transmitter with...hot and cold water!

Introduction

Solid-state heat pumps have been known since the discovery of the Peltier (Pel'-tyay) effect in 1834. The devices became practical only recently, however, with the development of semiconductor thermocouple materials. The modules I've used and describe here contain PN junctions made of bismuth telluride. The PN junctions are assembled between two flat, ceramic plates (see Figure 1). When current flows through the attached leads, one plate gets hot and the other gets cold. (In the majority of applications, the cold plate is used to cool something; the hot side, attached to a heat sink, dissipates the heat being pumped from the cold side.) In this application, the process is used in reverse: By heating one plate and cooling the other, we produce current to power a QRP transmitter.

Construction

The heart of this power supply is a Frigichip[2] thermoelectric module. You'll also need two ⅛-inch-thick aluminum plates[3] that can be formed to the shape shown in Figures 2 and 3, with the aid of a shop vise. Be careful not to distort or scratch the surfaces to which the cooler *module will be attached. Use a small* amount of heat-sink compound between the aluminum and ceramic surfaces to ensure good thermal conductivity.

To maintain thermal isolation between the hot and cold plates, the two aluminum plates are held together by nylon bolts and nuts. *Use caution when tightening the bolts!* Plates must be drawn up evenly to prevent cracking the ceramic. Do not overtighten them! Be gentle with the leads—constant flexing will cause them to break off at the module end. Solder tip or banana jacks to the lead ends to make connections to your transmitter.

Operation

I used the transmitter employed in the "Lemonized QSO" (see Note 1), although you can use any similar design. To power the transmitter, immerse the two device legs in plastic cups—styrofoam is best (see Figure 4). Fill one cup with hot water and the other with crushed ice and water. Although either device leg can be placed in the hot or cold water, electric polarity depends on leg placement. For proper polarity, the red lead should be positive, the black lead negative. Before connecting the power supply to your transmitter, allow a few minutes for the aluminum plates to conduct the heat and cold to the Peltier module. Connect a voltmeter across the leads to monitor the voltage rise. I measured a maximum potential of 1 V, although the transmitter operated at 0.5 V or less. This power source and transmitter were

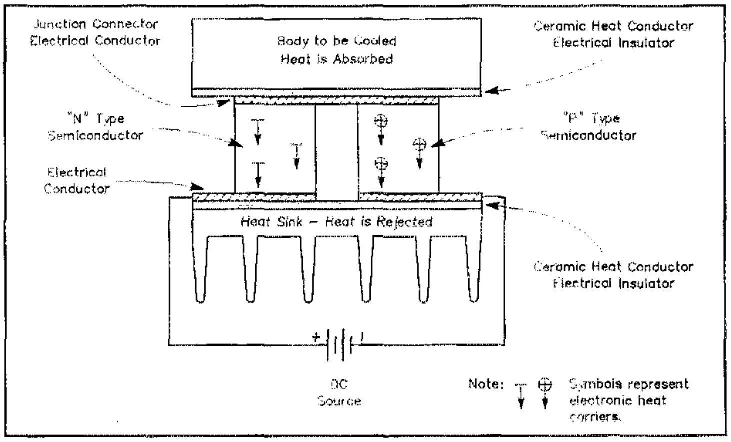


Figure 1—Cross section of a typical Peltier device being used as a cooling device. For the sake of clarity, only one junction is shown. The device used in this project has a total of 127 junctions.

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[1] Notes appear on page 33.

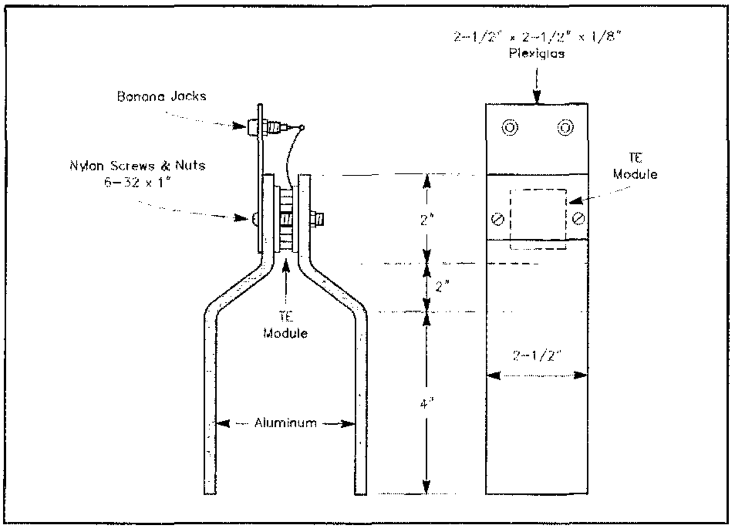


Figure 2—The mounting assembly for the Peltier device is made of aluminum and Plexiglas secured by nylon bolts and nuts . See Figure 3.

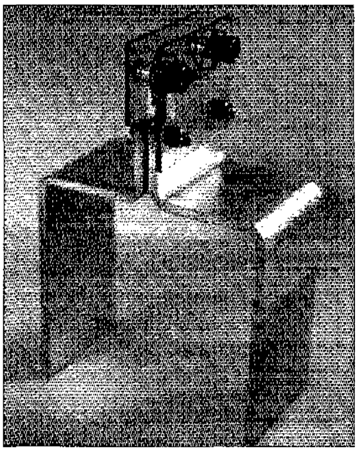


Figure 3—A Peltier device secure in its mounting assembly.

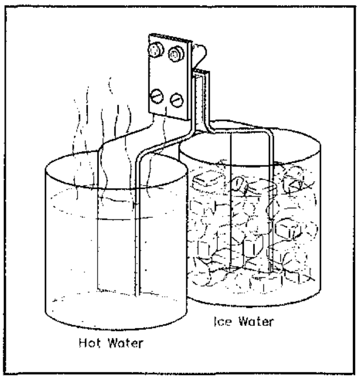


Figure 4—One leg of the Peltier device mounting assembly is placed in a container filled with hot water; the other leg is submerged in ice water. Although the title photo shows the use of glass beakers, styrofoam cups provide better thermal insulation and extend the operating period.

The Peltier Effect—Curiosity to Treasure

Jean Charles Athanase Peltier (1785-1845), was a French watchmaker turned physicist. He'd retired from watchmaking at the age of 30, and devoted the rest of his life to the study of physics. One of Peltier's discoveries was that when a current from an external source passes through a junction formed by two metals, it cools the junction. Further, when current is passed through the junction in one direction, it absorbs heat; when the current flow is in the opposite direction, the junction produces heat. We now know this as the Peltier Effect.

For many years a laboratory curiosity, Peltier devices have now become quite useful. This is due entirely to the development of new alloys for use in the junctions. The following list of current applications attest to the Peltier device usefulness:

- Consumer products: refrigerators for recreational vehicles (RVs) and mobile homes; aquarium coolers; picnic coolers.
- Scientific: coolers for transistors, ICs, diode lasers and infrared detectors.
- Aerospace and military: electronic equipment cooling and parametric amplifier coolers.
- Medical: Opthomological cornea freezers; tissue preparation and storage.
- Mobile refrigerators: medical supplies and food services.

The quality that sets this heat pump apart from all others is that it has absolutely no moving parts; its heat transferring process is completely electronic. (See Figure 1 in the article body.)—*Arnold Sayre, W8WVM*

used to complete a two-way QSO over a two-mile distance. The initial hot and cold water supplies lasted long enough to complete exchanging call signs, signal reports and brief comments.

Summary

Don't become overenthusiastic, throw away your NiCd cells and call your utility company to tell them you won't be needing their services any longer! It isn't likely that Peltier devices will replace other power sources in the immediate future. However, the Peltier device is experimentally interesting and practical. When used as described here, it is certainly an attention getter at club meetings and ham fests.

I present this article with two goals in mind. First, it's a novel demonstration of an alternative method of producing electric power. Secondly—and perhaps more importantly—this should give radio amateurs a chance to learn about a space-age device—albeit one that's 160 years old!

Peltier devices can be connected in series to increase the available voltage and in parallel to increase the current. Therefore, it is possible to thermally couple multiple devices to hot and cold tanks with, perhaps, hot water supplied by a solar heater and cold water piped from a nearby stream.

Right now, Peltier devices are more practical for cooling than for producing electricity. With mass production of the devices, perhaps their cost will drop and the use of Peltier-based thermoelectric generators will become commonplace.

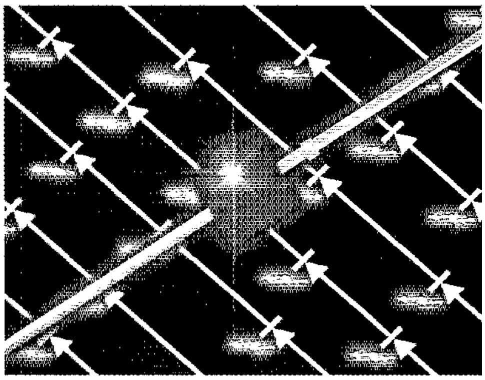
Notes

[1] B. Culter and W. Hayward, "Lemonized QSO," *QST*, Mar 1992, pp 18-19.

[2] Frigichips are manufactured and distributed by Materials Electronic Products Corp (MELCOR), 1040 Spruce St, Trenton, NJ 08648-4587, tel 609-393-4178, fax 609-393-9461. The preferred device size for this project is the No. CP 1.4-71-10L; price is $18.75 plus $10 shipping and handling. Similar units are available from All Electronics Corp, PO Box 567, Van Nuys, CA 91408-0567, tel: 800-826-5432, 818-997-1806; fax: 818-781-2653, and H&R Corp, PO Box 122, Bristol, PA 19007-0122, tel 800-848-8001, 215-788-5583, fax 215-788-9577. Check for availability and current prices before ordering.

[3] One source of aluminum sheet and nylon screws is Small Parts Inc, 13980 NW 58th Court, PO Box 4650, Miami Lakes, FL 33014-0650, tel 305-557-8222 (orders only), fax 800-423-9009, 305-557-7955.

Under the Hood: *Lamps, Indicators and Displays*



This sixth installment of Under the Hood illuminates the components that tell us what our rigs are doing.

By Bryan Bergeron, NU1N
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Our latest portable, dependable, and easy-to-use equipment places more emphasis on ergonomics and less on the inner workings of individual components. Communications equipment manufacturers focus on effective, useful and efficient controls, indicators and displays. Indicators for audio and RF level, frequency and mode, emergency conditions, power output, and signal levels give you real-time information about what's happening inside your equipment. Follow along for an introduction to the indicators—including LCDs, LEDs, neon, and incandescent bulbs—that you're likely to find in your communications and test equipment.

Display and Indicator Types

Analog Meters

No discussion of displays and indicators would be complete without mentioning the old standby—the once-ubiquitous analog meter (Figure 1). Even with the digital displays for frequency, mode, and channel used in most modern transceivers, there's something comforting about a mechanical, analog panel meter for monitoring rapidly changing signals, such as power output, SWR, and signal strength.

Analog panel meters generally rely on the magnetic interaction between a suspended coil of wire and the poles of a permanent magnet. Since the basic analog meters generally have a full-scale range of less than 1 mA, shunt resistors are required for measuring larger currents. Similarly, resistors can be added in series with the meter to create a voltmeter. Many analog meters have integral series or shunt resistors, with a faceplate calibrated according to the full-scale voltage or current indicated by the meter.

Despite their long-standing popularity, analog meters are becoming increasingly rare. Analog meters are generally bulkier; less accurate; less shock-, vibration-, and moisture-resistant; and, perhaps most significantly, much more expensive than the newer digital indicators and meters. You may find that a new panel meter doubles the cost of your project. Luckily, there are solid-state alternatives to the mechanical panel meter, such as liquid crystal displays, that are much more cost-effective.

Liquid Crystal Displays

Liquid crystal displays or LCDs are found in everything from digital watches and thermometers to hand-held digital multimeters and VHF transceivers. LCDs are popular because they're inexpensive, require very little power, don't flicker, produce minimal electromagnetic fields, have a fast response time and long life span, and provide a pleasing, high-contrast display that can be viewed in direct sunlight. If you have a modern mobile, portable, or hand-held rig, it probably uses an LCD.

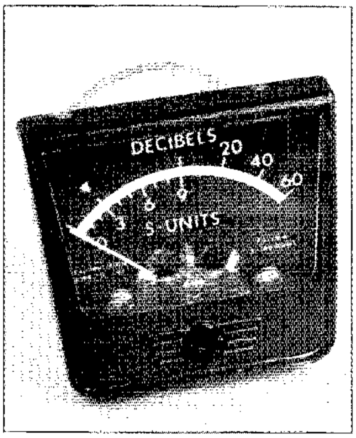

Figure 1—Analog panel meters are popular in ham gear, but their fragile construction has led to the adoption of solid-state substitutes.

LCDs take a variety of forms, from digital numeric panels to mock analog panel meters. Whatever the form, they are constructed of a thin film of viscous liquid crystal sandwiched between two glass plates with transparent etched electrodes. When opposing etched electrodes are energized, the optical properties of the liquid crystal matrix changes, rendering the area defined by the electrodes visible.

LCDs have come a long way in the past few years, in part because of pressure from the portable TV and laptop computer market. The main distinction in LCDs is between *passive matrix* displays and the more capable (and more expensive) *active matrix* displays.

Passive-matrix LCDs, like those found in inexpensive wristwatches, have poor contrast and limited viewing angles.

Active-matrix displays aren't commonly used in communications equipment (unless you consider a laptop computer part of your packet station), in part because of their great expense. You're much more likely to find this type of display in your laptop, at least until the technology matures. In active-matrix LCDs (AMLCD) designs, a thin-film transistor and capacitor are used at every pixel. In effect, the panel is a giant IC. Although they're expensive, AMLCDs provide full color, a wide viewing angle, good response time, and impressive contrast.

Electroluminescent Lamps

Electroluminescent (EL) lamps are used to backlight LCDs for many of the same reasons LCDs are so popular. EL lamps are inexpensive, lightweight, require very little power, have long life spans, and generate little heat. Other key attributes of EL lamps is their uniformity of illumination and availability in a variety of colors, including white, red, green, blue-green, and yellow. If your mobile transceiver has an illuminated LCD, odds are it's illuminated by an EL lamp.

EL lamps are constructed with transparent conductive electrodes that are separated by a dielectric containing luminescent phosphor. Application of an ac voltage to opposing electrodes causes the dielectric between them to glow. Brightness can be easily controlled by varying the magnitude and/or frequency of the applied voltage.

Incandescent Lamps

When a wide viewing angle, great brightness, and a variety of colors are required, it's difficult to compete with the incandescent lamp (Figure 2). Incandescent lamps, which are brighter than either LCD or neon lamps (typical output for an incandescent lamp is on the order of 0.5 to 2.8 lumens), consist of a thin tungsten filament mounted between two electrodes in an evacuated glass bulb (a tungsten filament would fail within a few seconds if it were operated in the presence of oxygen). When current flows through the filament, the electrical energy is transformed to light. A small amount of inert gas may be added to improve efficiency and color; any color is possible with the use of filters.

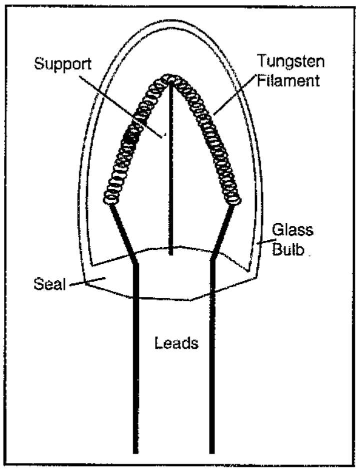

Figure 2—Incandescent bulbs are constructed with a thin tungsten filament, often stabilized by a molybdenum support, encased in an evacuated glass envelope. Operating life and light output can be increased by filling the envelope with argon, nitrogen, krypton, or other gas. Halogen bulbs use a quartz envelope that is filled with a halogen gas such as iodine or bromine.

Tungsten is a popular filament material because of its low rate of evaporation and high melting point. Due to the limited space inside the lamp envelope, the tungsten is coiled and mounted into various filament shapes, often supported by molybdenum posts.

Typical operating voltages (from 5 to 115 V) and currents (from a few mA to an ampere or more), are determined by the size, length, and composition of the filament. Lower-current lamps have a smaller diameter tungsten filament and therefore higher resistance. However, power requirements also influence an incandescent lamp's susceptibility to shock and vibration. Low-voltage, high-current lamps have thicker filaments and are therefore less fragile; eg, a 28-V, low-current lamp will be more fragile than a 5-V lamp operated at higher current.

In addition to their relatively fragile tungsten filaments, a major limitation of incandescent lamps is their relatively short life-spans. Even when operated away from the shock and vibration common in portable equipment, incandescent bulb life spans are commonly in the hundreds or, at best, thousands of hours. As a result, the incandescent lamps require the added cost of a socket that facilitates bulb replacement.

Incandescent lamp life span is usually rated at 60-Hz ac. Operating an incandescent lamp on dc decreases lamp life by approximately 50%. Dc alters the molecular structure of the tungsten, resulting in hot spots on the filament, which accelerates the rate of tungsten evaporation and embrittlement.

Neon Lamps

Neon glow lamps are popular in ac-power-supply circuitry, both as indicators and as light-duty surge suppressors, in part because they are line-voltage compatible, require very little power, and are relatively inexpensive (Figure 3). Neon lamps are less prone to shock and vibration failures than are incandescents; typical life spans are about 25,000 hours. Compared to LEDs, neon lamps have a much shorter life span. Unlike LEDs, though, neon lamps can be operated at higher temperatures (up to 150° C) and are not damaged by voltage transients or high-voltage static discharges.

Neon glow lamps are constructed with a pair of oxide-painted nickel electrodes mounted within an envelope containing neon gas under low pressure. When a high voltage is applied through a current-limiting resistor, the gas ionizes and produces the characteristic reddish-orange glow. A small amount of mercury can be added to produce a bluish tint; other colors are made possible by coating the inside surface of the envelope with phosphor.

An important concept in the operation of neon lamps is breakdown voltage—the voltage at which a lamp starts to glow steadily. Breakdown voltage is a function of the gasses use, electrode spacing and geometry, and the geometry of the tube. It is also influenced by ambient light (harder to fire in the dark), and the time since the last discharge (the longer the time, the harder to fire). In most neon lamp designs, the breakdown voltage is normally between 55 and 95 V ac. Dc breakdown voltage is roughly 50% greater than the ac breakdown voltage. Similarly, neon lamp ac breakdown voltage increases with increasing frequency, in part due to the capacitance of the lamp—about 0.5 pF.

Typical current for a neon lamp is between 0.5 and 3.0 mA. Since the internal resistance of a neon lamp is from about 3 to 30 kΩ, a series resistor (150 kΩ is typical) must be used to limit lamp current. Since life span increases considerably as operating current decreases, a small decrease in operating current results in a considerable increase in life span. Interestingly, light output is directly proportional to current, so a small decrease in operating current results in little output loss. Lamp life is decreased significantly (about 50%) by operating a lamp at dc versus the same rms ac voltage.

Unlike incandescent lamps and LEDs, neon lamps fail rather gracefully. As neon lamps age, the electrodes evaporate and condense on the inside of the glass envelope. The firing voltage increases until it reaches the value of the supply voltage. At this point, the lamp flickers and exhibits a slight decrease in brightness.

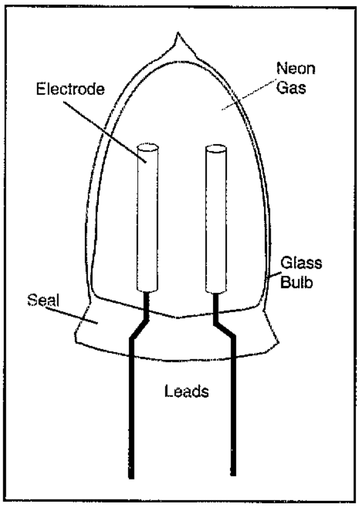

Figure 3—Neon glow lamps are constructed of two parallel electrodes encased in a glass envelope filled with neon gas under low pressure. When the voltage across the electrodes exceeds the breakdown voltage of neon, the ionized neon gas emits a characteristic reddish-orange glow.

Light-Emitting Diodes

Light-emitting diodes or LEDs are rapidly replacing both incandescent and neon lamps in communications equipment. The reasons are clear: LEDs consume less power, are smaller, have better than a 50-fold improvement in life expectancy over incandescent and neon lamps (typical life span is about 11 years), are resistant to shock, vibration, moisture, and temperature extremes, provide good visibility even in direct sunlight, and are available in virtually any conceivable configuration, from discrete lamps, 7, 14, and 16-segment digital displays, to high-output clusters, high-density surface-mount chips, bar graph, dot-matrix, and alphanumeric displays.

When LEDs were first introduced in the late 1960s, they were available in only low-output, single-color (red) configurations. Advances in technology gave us higher-intensity LEDs in a variety of colors, including warm whites, in bases that match those of the incandescent lamps LEDs will replace.

LEDs are simple PN junction semiconductor devices. When the LED is forward biased, electrons in the N region combine with holes near the PN junction, and light is generated (Figures 4 and 5). An LED may be forward biased from a battery or other steady dc source, or from a pulsed dc source up to about 100 MHz. The energy level associated with the recombination, and therefore the wavelength of the light produced, is a function of the semiconductor material used.

Like many semiconductor devices, when LEDs fail, they do so catastrophically. LEDs are rated in terms of maximum and typical forward current, typical forward voltage, maximum reverse voltage, viewing angle, output color, packaging, diffused or nondiffused, and intensity. A typical LED rating would be 30/20 mA max/typical forward current, 1.4-V forward voltage, and 5-V maximum reverse voltage. Output intensity is generally specified in millicandelas (mcd) at specified current; 3 mcd at 20 mA is typical, although high-intensity LEDs (eg, greater than 1000 mcd) are available.

LEDs, like neon lamps, are generally designed to be used with a series resistor that limits the forward current to a safe value (some LEDs designed for a specific supply voltage have built-in series resistors). The value of the series resistor can be determined by the formula: $R_s = (V_{in} - V_{LED})/I_{LED}$, where V_{in} is the supply voltage, V_{LED} is the LED voltage drop, and I_{LED} is the desired maximum current. V_{LED} varies from about 1.3 to 2.5 V, depending on the LED construction and output wavelength (green LEDs tend to have a greater V_{LED} than do red LEDs). For example, with a 6-V supply and a LED with a 1.7-V voltage drop, and a desired current of 20 mA, the optimal resistor would be about 220 Ω.

LED package configurations greatly influence light output and appropriateness for a given application. For example, there are cylindrical, inverted cone, arch, rectangle, and square package designs. An LED molded into the shape of a lens can produce roughly 10 times the light output of the same semiconductor LED housed in a

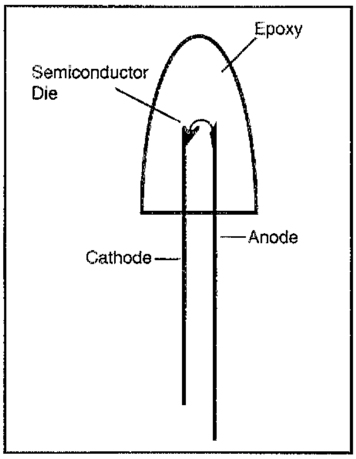

Figure 4—LED construction. Unlike incandescent and neon lamps, LEDs are polarized. The anode (+) lead is longer than the cathode (−) lead. In the design shown, the transparent epoxy dome acts as a lens to focus the light output. LEDs are current dependent, in that light output is directly proportional to the forward current.

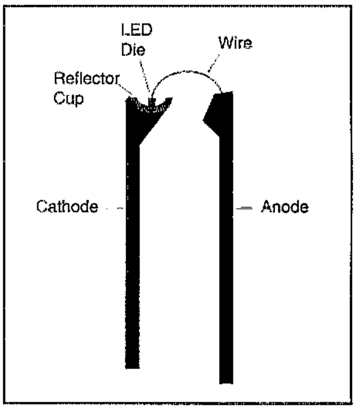

Figure 5—Close-up of LED construction. The semiconductor LED die sits atop a concave reflector that directs and concentrates the light output.

Light Output Terms and Factors

Look at any good electronics supply catalog and you'll see lamps described in terms of optimal operating voltage and current, estimated life span, and brightness. Operating current and voltage, specified in typical and maximum amperes and volts, respectively, are familiar concepts to all of us. Estimated lifespan, generally specified as average lamp life, is the time before half of an entire lot of lamps can be expected to fail. For example, a lamp rated at 500 hours would be expected to burn out in one half of all of its applications by the 500th hour. That is, a given lamp could fail after 5 or 5000 hours, but odds are it will fail at around 500 hours.

Unless you've worked with laser communications or other applications that deal with optics, the terminology associated with brightness or luminous intensity may seem a bit foreign at first. When quantifying the light output of a device, the candela is the standard unit of measurement, replacing the once popular candlepower. Whereas luminous intensity, measured in candelas, is a measure of the light output of a device, the lumen is a measure of the amount of energy in a beam of light shining on a surface. These units are tied together in the following relationship: a 1-candela light source produces a 1-lumen beam of light, which results in 1 foot-candle illumination on a 1-square-foot area located 1 foot from the source.

Large light sources, such as high-powered halogen lamps, are measured by total light output. In such cases, the usual unit of light output, the Mean Spherical Candlepower (MSCP) is a measurement of total visible light being emitted from the lamp. MSCP can be converted to a lumens equivalent by multiplying the MSCP value by 12.57. In contrast, the intensity of LEDs and other small light sources is measured in millicandela (mcd). The intensity of small, point sources, such as LEDs, are generally measured on-axis.

It can be helpful to think of light source as a small RF antenna. For example, the light intensity must double for the eye to detect a noticeable difference, eg, from 10 to 20 mcd or from 100 to 200 mcd. Similarly, light intensity is angular-dependent. That is, a 10-mcd device with a viewing angle of 35° may have the same total light output as a 200-mcd device with a viewing angle of 8°. Like adding elements to a beam antenna, the light intensity produced by a lamp can be effectively increased an order of magnitude by using lenses. Lenses, whether an integral part of the lamp or an external device, collect the light emitted from a naked lamp and project it into a useful direction.

transparent rectangular epoxy mold; internal reflectors can raise the output even further.

One of the more interesting LED designs is the bicolor LED, which can be switched from one color to another (typically from red to green). Bicolor LEDs are formed by putting two LED chips on the same package as a single LED, connected in reverse parallel. When biased in one direction, one LED lights; when biased in the other, the other generates light. By varying the duty cycle of each color, a third state can be displayed; eg, a mixture of red and green produces amber. Some bicolor LEDs are made in three-LED packages, which simplifies the drive circuitry needed to display three states. Because of the three states available, bicolor LEDs can be used to pack a lot of information in a small space. Applications for bicolor LEDs range from dc polarity indicators to front-panel function indicators.

Miscellaneous Devices

In addition to the major indicators and displays described above, a number of miscellaneous display devices deserve mention. For example, the ubiquitous cathode-ray tube (CRT) is found in or around most communications stations. The ICOM IC-781 transceiver and R-9000 receiver, for example, use a CRT to display the frequency spectrum of received signals as well as operating frequency and mode information. Many ham stations feature modulation monitor scopes and even panoramic adaptors, that allow limited spectrum analysis when used with a receiver. Finally, CRTs are used in nearly all computer monitors.

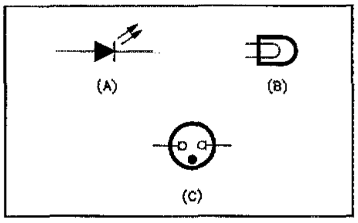

Figure 6—Schematic symbols of LEDs (A), incandescent lamps (B), and neon-glow lamps (C).

Incandescent digital readout tubes, which are bright, operate at a low current, and provide a wide viewing angle, are also found in some communications and test equipment. Numerical neon glow tubes, often used in test equipment, provide a pleasing display, albeit with a somewhat restricted viewing angle because of their multi-plane construction. Microprocessor-controlled vacuum-fluorescent display modules are used by a number of equipment manufacturers.

Summary

Display technology is undergoing rapid evolution. Economic forces are bringing to the market devices that are more compact, more cost effective, more reliable, and provide more functionality than their predecessors. LEDs, neon lamps, and other discrete display components can be expected to be with us for some time. There is, however, considerable movement toward virtual interfaces, in which your computer screen becomes a powerful and easily customized user interface to a "black box" communications unit.

New Books

MASTERING RADIO FREQUENCY CIRCUITS THROUGH PROJECTS AND EXPERIMENTS

By Joseph J. Carr, K4IPV

Tab Books, Division of McGraw-Hill Inc, Blue Ridge Summit, PA 17294-0850; tel 717-794-2191; fax 717-794-2103. Softcover, 7¼x9 inches, B&W artwork, 411 pp, $19.95

Reviewed By Jim Kearman, KR1S
QST *Assistant Technical Editor*

It's no secret that you can pass every Amateur license exam through Extra Class without having a clue about electronics. When you get down to it, the goal is to pass the exam by getting the right answers to the questions. Most of us promise to go back and really learn the theory, but simply studying the books can put you to sleep. This book by the prolific Joe Carr seems oriented toward those of us with ham licenses and guilty consciences. Starting with a refresher of the simple math you need to know to understand how radio stuff works, Joe follows with details on home construction techniques and simple test equipment. But the bulk of this hefty book consists of schematic diagrams and descriptions of practical RF circuits, from VLF through the microwaves. Even working fast, it would take you a lot of evenings and weekends to build every project.

Most of the circuits are oriented toward receivers. Because this book may fall into the hands of unlicensed students, that's probably a good idea. I couldn't find any power amplifier circuits, for example, even in the extensive index. Most receiver circuits, such as mixers, IF amplifiers and the switching of IF filters, are covered. Some coverage of automatic gain control (AGC) circuits would have been nice, though, and audio circuits are dealt with only in passing, with the 386 IC. Much better audio chips are available now and should have been included. A poorly conceived audio stage can make an otherwise excellent receiver sound like a piece of junk.

If I ever set out to write a book like this, I think I'd have the reader construct a series of modules that would eventually end up as a complete transceiver or separate receiver and transmitter. Although each circuit is described thoroughly and is interesting by itself, when you finish the book you're just going to have a pile of circuits. Nowadays, knowing how to integrate modules is as important as designing the modules themselves.

To excuse the lack of focus, you could say this book wasn't written exclusively for hams, although we'll probably be its biggest market. But if Joe intends the book to appeal to "professionals," why the heavy emphasis on direct-conversion receivers? An entire chapter is dedicated to them; nothing approaching equal coverage of superheterodyne receivers is to be found. It's obvious that Joe hopes to sell a lot of copies of this book to hams, but some of amateurs build superhets, too. So why not include some transmitter circuits? And although the title says the book covers "projects and experiments," I couldn't find any experiments.

Amateur Radio definitely needs a project-oriented text. The ARRL's *Solid State Design* is in many ways out of date. When I first saw *Mastering Radio Frequency Circuits*, I hoped it would be a fitting replacement for *Solid State Design*, but it's not. If you want to learn by doing about radio circuits and have some useful pieces of equipment when you're finished, *Solid State Design* is still the best value. *Mastering Radio Frequency Circuits* has value as a compendium of circuits, and the text contains valuable information. In fact, compared to other offerings on the market, this book seems underpriced. But I don't think you'll master RF by using it.

Nickel-Metal-Hydride Batteries in Amateur Radio Applications

Meet the newest power source for portable equipment: the NiMH battery

By Gary Kuusisto, N6TCF
333 Holly Creek Court
Apartment 2014
The Woodlands, TX 77381

Most hams are familiar with nickel-cadmium (NiCd) batteries, used in our hand-held transceivers. NiCd batteries offer the possibility of hundreds of charge-discharge cycles, and reasonable energy capacity. Nickel-metal-hydride (NiMH) batteries have voltage profiles and usable lifetimes similar to NiCd batteries, but offer 30 to 40% greater energy capacity. What's more, when a NiMH battery has reached the end of its career, you can dispose of it without risk of polluting the environment with cadmium, a toxic metal. Because the two battery types are so much alike, it is easy to substitute a NiMH battery where a NiCd is used now.

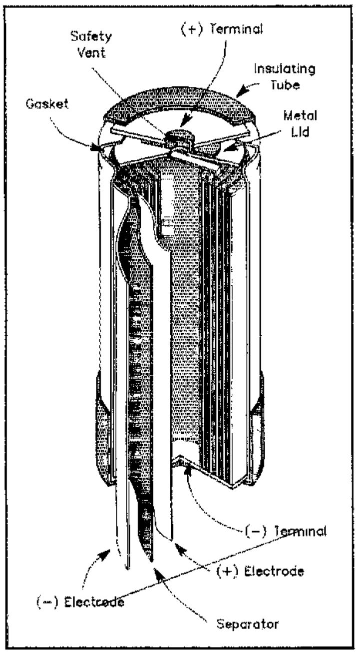

Figure 1—Cutaway drawing of a nickel-metal-hydride (NiMH) cell. Except for the cathode, which is made of hydrogen-storage metal instead of cadmium, construction is very similar to that of the nickel-cadmium (NiCd) cell. Hydrogen-storage metal is preferred because it is less toxic than cadmium, and gives the NiMH cell a greater energy storage capacity.

Similar Construction/Better Performance

NiMH cell construction is similar to that of NiCd cells. In fact, they use many of the same components. In either a NiCd or NiMH cell, the positive plate is made of nickel and the electrolyte is potassium hydroxide. The only difference lies in the choice of metal used for the negative plate (Figure 1). NiCD cells have cadmium negative plates, while hydrogen storage metal is used in NiMH cells. During a charging cycle, the hydrogen storage metal plate absorbs hydrogen, which it releases when the cell is discharged, allowing current to flow.

NiMH batteries operate over the same temperature range as NiCd batteries (–20° to 60°C (–4° to 140°F). .

Hydrogen storage metal is a compound capable of storing a quantity of hydrogen gas hundreds of times its own volume, at less than atmospheric pressure. This feature gives NiMH batteries up to 40% more capacity than NiCD batteries of the same size.

Figure 2 demonstrates the higher energy density of NiMH batteries over NiCd and lead-acid batteries. Higher energy density is especially important for portable equipment like hand-held transceivers. NiMH batteries of equivalent capacity will weigh less than NiCd counterparts. Of perhaps more importance, you'll be able to operate for longer periods without discharging your batteries. Hams active in public-service communication will appreciate the longer discharge cycle.

Discharge Performance of NiMH Cells

Figure 3 graphs the voltage curves for NiMH and NiCD cells over a typical discharge cycle. At full charge, a single NiMH cell has a terminal voltage of 1.35 V, which quickly drops to a nominal 1.2 V. The recommended low-voltage cutoff for both cell types is 1.0 V. As you can see in Figure 3, the greatest difference between the two curves is that the NiMH cell has a much greater capacity.

One difference between NiCd and NiMH cells is their maximum discharge current. At high discharge rates (3 or more amperes), an AA NiMH cell has less ampere-hour capacity than its NiCD counterpart. The reason is the higher internal resistance of the NiMH cell. Fortunately, this characteristic isn't a problem for most devices, and the discharge voltage remains stable.

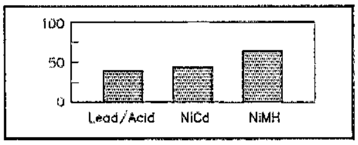

Figure 2—Energy density and specific energy of common rechargeable cells, in watt-hours/kg. Higher energy density means that, for a given battery weight, you get more hours of operation.

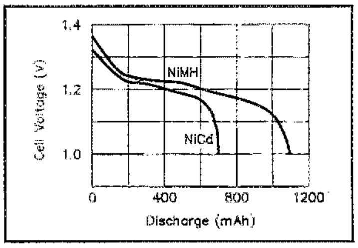

Figure 3—Discharge curves for NiCd and NiMH cells. Both cells behave similarly near the end of their discharge cycles, making NiMH cells compatible with low-voltage cutoff circuits in battery-powered equipment.

Charging NiMH Cells

NiMH cells are very sensitive to overcharging. Although this is also true for NiCd cells, a NiMH cell will have drastically reduced capacity after a couple of abusive charging cycles. A constant-current charger coupled with an end-of-charge controller will provide more than 500 charge/discharge cycles from a NiMH cell.

At full charge, the terminal voltages of both NiCd and NiMH cells drop slightly. This *negative delta V (voltage)* point is sensed by many advanced charge-control systems and used to terminate charging. At full charge, the state of charge of the cell no longer changes. The cell begins to act like a resistor and dissipates the charging current as heat. This rise in cell temperature causes a drop in both the cell's internal resistance and voltage.

The standard charging technique for NiMH batteries uses a constant current. If the battery is fully discharged, energy equal to 150% of its capacity is applied to fully charge it. For example, NiMH cell rated at 1100 mAH capacity is charged at 110 mA for 15 hours, for 1650 mAH of charge. The allowable temperature range during charging is 0° to 45°C (32° to 113°F). Charge control, which is essential for NiMH cells, can be accomplished either by sensing the cell voltage or temperature.

Storing NiMH Batteries

Allowable storage temperatures for NiMH batteries are from –20° to 35°C (–4° to 95°F). NiMH batteries return to service quickly when stored in a discharged state (terminal voltage less than 1.0 V per cell). After a couple of charge/discharge cycles they will deliver full capacity. What about when NiMH batteries are stored charged? In this type of service, NiMH batteries don't perform as well as NiCd batteries. When stored at room temperature, a fully charged NiMH battery will be completely discharged after 15 days. NiCd batteries, on the other hand, will retain their charge for 30 to 60 days. If you use and recharge your batteries frequently, this characteristic isn't a disadvantage.

Long-term overcharge and shallow discharge of NiMH batteries may lead to slightly lower terminal voltage over the life of the battery.

Conclusion

NiMH batteries offer higher energy density and an environmentally safer chemistry. On the other hand, NiMH batteries have poorer charge retention, higher internal resistance and more specialized charging needs than NiCds. Making the transition from NiCd to NiMH batteries requires users to accommodate slightly different charge, discharge and storage characteristics, but the two types are essentially interchangeable.

Acknowledgment

I referred to *NiMH Technical Notes* (Bethel, CT: Duracell, Inc) while preparing this article.

New Products

CT KEYBOARD TEMPLATES

◊ In the heat of a contest, who has time to concentrate on remembering all the commands in K1EA's *CT* logging software? When the QSOs are rolling in, it's no time to press Alt-H for a help screen or search through the manual. Contest Assistant is a set of precut templates for most standard 101-key keyboards (a set for Gateway AnyKey keyboards is also available). They show the commands available on all the keys programmed by *CT*, including all shifted (Shift, Alt, Ctrl) key combinations. They're made of heavy, plastic-laminated material (in a manila-envelope color to stand out on most keyboards), in a two-piece design for easy storage or uncut for custom use. Retail price is $7.50 for the ready to use version, $4.50 for the uncut kit (s/h $1.25). VCH Products, 7433 Popp Rd, Ft Wayne, IN 46845; tel 219-627-2604.

COMPUTERIZED ROTATOR CONTROLS

◊ When trying to focus a Yagi on a distant station, it can be slow and confusing to mentally convert a beam heading to a direction on the face of a rotator-control box. A new unit from Pro-Search provides digital conversion of beam headings (easy to note in your log) with a 360° readout in continuous 1° increments on a 1½-inch-wide, ½-inch-high display. It works with all 8-wire CDE rotators. Retail price is $59.95 plus s/h.

The shack of the '90s can be automated further with the PSE-1 series computerized rotator controllers, with automatic braking and programmable features. It even "talks" and scans directions. A 16-button control panel lets the operator directly enter the desired antenna azimuth or turn the antenna manually with directional control buttons. There are presets for five continents and users can assign favorite beam headings to memories. It's compatible with HAM-M, HAM-II, HAM-III, HAM-IV, Tailtwister, HDR-300 and other popular rotators with no modifications. Retail prices range from $189.95 to 469.95. Pro-Search Electronics, 1350 Baur Blvd, St Louis, MO 63132; tel 800-325-4016 or 314-994-7872.

VHF/UHF ANTENNAS

◊ Extra range comes in handy when you try to punch your signal through the increasingly busy bands from 144 to 1300 MHz. A new family of rugged, optimized, high-performance Yagi antennas is available for FM, ATV and packet radio applications. The 2-meter antennas for FM and packet include the 4-element COY2M4EL for $59 and the 3-element COY2M3EL "Stealth" for $65. FM and packet antennas for 223 to 225 MHz include the 4-element COY2234EL for $65 and the COY2235EL for $59. The 70-cm antennas include the 3-element COY4393EL for packet and ATV on 427 to 441 MHz at $49, and two 7-element models: the COY4347EL for ATV, repeater links and SSB on 420 to 440 MHz for $65 and the COY4407EL for FM on 440 to 450 MHz for $65. The 16-element COY23CM16EL covers ATV, links and FM on 1240 to 1300 MHz at $110. The COY2M440 is a dual-band vertical folded monopole for 2 meters and 70 cm at $49. Swiech Communication Systems, 12218 Greentree Rd, Poway, CA 92064; tel 619-748-2286.

WATTMETER CONVERSIONS

◊ You can go crosseyed trying to follow some average-reading analog wattmeteers, such as those used on Collins 312B-4/5 consoles. C. J. Hawley, KE9UW, designed the PDC-1 as a simple peak-detector circuit to convert almost any averaging-type wattmeter to a peak-reading wattmeter with an adjustable needle hang time. At only 2×1½ inches, it's been installed inside Drake, MFJ, Daiwa, Ten-Tec, Heathkit, Nye Viking, Bird and other wattmeters. Retail price is $19.99 for the ready to assemble kit. Floyd Soo, KF8AT, Hi-Res Communications Inc, 18464 Ash Creek Dr, Macomb, MI 48044; voice or fax 810-228-1600.

CODE KEYS

◊ True-blue CW fans and collectors will be impressed with the fine craftsmanship and styling of a new series of imported German code keys, including the $149.95 mini, the $164.95 mobile key, the $169.95 Portable Warbler, the $199.95 Profi and the $209.95 mahogany base straight key. Schurr Morse Keys, Electronic Switch Co Inc, Suite E-6, 4343 Shallowford Rd, Marietta, GA 30062; tel 404-518-4634.

A Smart Charger For Nickel-Cadmium Batteries

Recharge your hand-held's battery pack FAST, with this easy weekend project.

By Steven Avritch, WB1EOB
PO Box 9303
Forestville, CT 06011-9303

Wouldn't we all like to thank the people who invented NiCd rechargeable batteries? NiCd batteries save us from spending countless dollars on throw-away (and expensive) conventional (nonrechargable) alkalines. Everyone familiar with NiCds also wishes they were easier to charge. Trickle charging takes forever, but speeding up the process can damage batteries by overcharging. We all know that NiCds must be charged correctly or battery life is significantly reduced. Since Mother Nature isn't about to change the rules for charging NiCds, a Smart Charger is needed. This article describes a simple yet sophisticated NiCd battery "quick" charger. This Smart Charger is based on the MAX713 single-chip battery charger. The unit continuously monitors the battery during fast charge and automatically switches over to trickle charge when the battery is full—eliminating any possibility of overcharge.

MAX713 Single-Chip Battery Charger

The MAX713 single-chip charger can fast charge and trickle charge any NiCd battery pack containing from 1 to 16 cells (ie, battery pack voltages ranging from 1.2 to 19.2 V). The charger in this article, however, is limited to packs having a maximum of nine cells, because of the 12-V wall adapter power supply. Charge rates are programmable from 4C (20-minute charge) to C ÷ 3 (4-hour charge), where C is the capacity of the battery in milliamperehours (mA-h). The trickle charge rate is fixed at C ÷ 16. The MAX713 uses three different techniques to determine when to end the fast charge and switch over to trickle. These techniques are:

- Voltage peak detection
- Battery temperature
- Time.

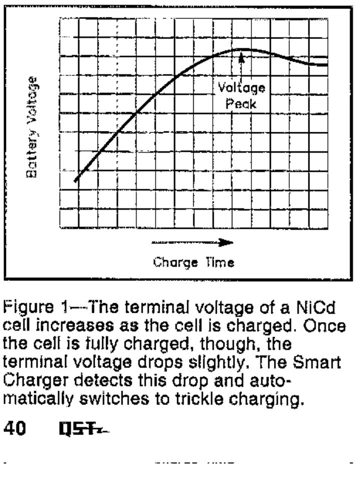

Figure 1—The terminal voltage of a NiCd cell increases as the cell is charged. Once the cell is fully charged, though, the terminal voltage drops slightly. The Smart Charger detects this drop and automatically switches to trickle charging.

Voltage peak detection is the best method to determine when a NiCd battery is fully charged. As NiCd batteries charge, the voltage across the battery slowly increases. This voltage peaks when the battery is fully charged and then drops off slightly (see Figure 1). The MAX713 continuously samples the battery voltage and shuts down when the voltage peak is detected. Also, if the battery voltage does not peak within a preset timeout period, fast charge is terminated.

The MAX713 also has provision for thermally protecting the battery during the charge cycle. A high-temperature shutoff will terminate the charge cycle if the battery temperature exceeds a preset limit. Also, a low- (cold) temperature shutoff will not allow fast charging until the battery temperature reaches another preset limit. The temperature-detection features are optional and require the addition of three thermistors and a few other parts. One of the thermistors must be in contact with the battery, which is not practical with sealed battery packs. To simplify the project, the Smart Charger does not use temperature detection. Note that temperature detection *is* recommended if the charge rate selected is greater than 2C.

Design Overview

This charger is capable of charging battery packs from one to nine cells (1.2 to 10.8 V). The number of cells in the pack to be charged must be programmed into the MAX713 by connecting the PGM pins of the 713 as defined in Table 1. This design assumes that the battery pack(s) to be charged will always have the same number of cells. If you wish to build a charger capable of charging battery packs with varying numbers of cells, switches must be added to change the PGM pin connections according to the battery to be charged.

The charging section is simply the MAX713 and a few external components (see Figure 2). The entire circuit is powered by an inexpensive 12-V dc wall adapter. The current capacity of the adapter should be at least 50 mA greater than the preprogrammed fast-charge rate current (ie, if the fast charge rate current is 250 mA, then the wall adapter should be rated for at least 300 mA).

Designing a Charger

A few decisions must be made before the design can begin:

- Wall adapter voltage (minimum)
- Number of cells in pack to be charged
- Fast charge rate
- Time-out period (maximum time in fast charge).

The following charger design is for a 7.2-V, 500-mAh battery pack, the kind that come with many hand-helds. The wall adapter voltage should be at least 1 V higher than the battery being charged. The number of cells in the pack is determined by dividing the pack voltage by 1.2; the 7.2-V pack contains 6 cells. The charge rate is selected

Table 1
Maxim MAX713 Programming for 1- to 16-Cell Packs

No. of Cells	*PGM1*	*PGM0*
1	V+	V+
2	OPEN	V+
3	REF	V+
4	BATT–	V+
5	V+	OPEN
6	OPEN	OPEN
7	REF	OPEN
8	BATT–	OPEN
9	V+	REF
10	OPEN	REF
11	REF	REF
12	BATT–	REF
13	V+	BATT–
14	OPEN	BATT–
15	REF	BATT–
16	BATT–	BATT–

OPEN	Not connected
BATT–	Connect to pin 12
REF	Connect to pin 16
V+	Connect to pin 15

to be C + 2 (250 mA), which will result in a charge time of approximately 2.5 hours. The charge time is calculated by:

Charge time = C ÷ (charge rate × 0.8). (Eq 1)

The 0.8 accounts for the fact that the charging process is only about 80% efficient.

Therefore, for this design:

- Number of cells = 6
- Charge rate = C ÷ 2 (250 mA)
- Time-out = 2.5 hours (150 minutes)
- The wall adapter is 12 V dc, rated at 300 mA.

There are four steps in the charger design:

1. Calculate the power dissipation in Q1
2. Calculate the value of R1
3. Calculate the value of R_{SENSE}
4. Determine the PGM pin connections.

Step 1

Check to ensure that the power dissipated in the PNP transistor (Q1) does not exceed the maximum power rating of the transistor. The power dissipated by Q1 is calculated by:

Power (Q1) =
(max adapter V – minimum battery V) × charge rate (A) (Eq 2)

The minimum battery voltage will be 1.0 × number of cells, or 6.0 V.

In our case, the power dissipated by Q1 is:

Power_{Q1} = (12 – 6.7) × 0.250 = 1.50 W

Make sure that the power dissipation rating of Q1 exceeds 1.5 W.

Table 2
Maxim MAX713 Programming for Various Time-Out Periods

Time-Out Period (Minutes)	*PGM3*	*PGM2*
22	V+	REF
33	V+	BATT–
45	OPEN	REF
66	OPEN	BATT–
90	REF	REF
132	REF	BATT–
180	BATT–	REF
264	BATT–	BATT–

See Table 1 for explanation of programming pin connections.

Step 2

The value of R1 (refer to Figure 2) is calculated by:

R1 (kΩ) = (minimum adapter V – 5) + 5 (Eq 3)

In our case, R1 = (12 – 5) + 5 = 1.4 kΩ

Step 3

The value of R_{SENSE} (refer to Figure 2) is calculated via the following equation:

R_{SENSE} (Ω) = (0.25) + charge rate (A) (Eq 4)

In our case,
R_{SENSE} = (0.25 ÷ 0.250) = 1 kΩ

Step 4

The last thing to do is determine how to connect the four PGM pins on the MAX713. PGM0 and PGM1 determine the number of cells in the pack (6 in our case) and should be connected according to Table 1.

PGM2 and PGM3 determine the max charge time, which should be connected according to Table 2 (150 minutes in our example; note that 150 minutes is not in the table, therefore we'll select the next greater

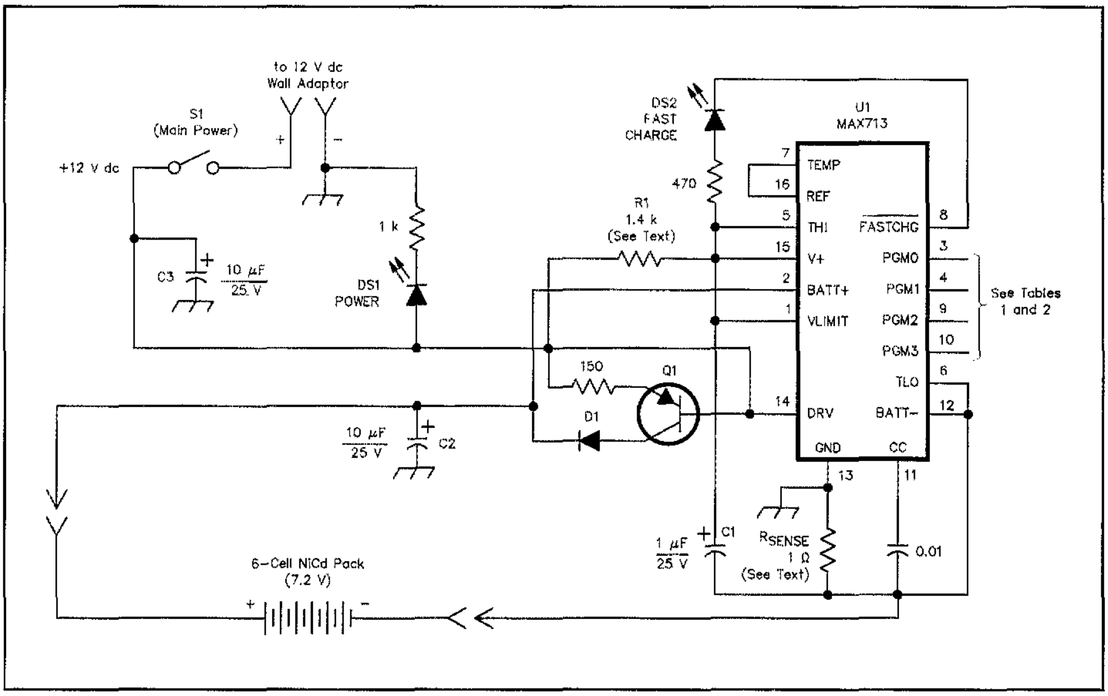

Figure 2—Schematic diagram of the Smart Charger, as set up to charge a 7.2-V pack. All parts except the Maxim MAX713 are available from Radio Shack stores. See text for details.

D1—1N4001 or eqivalent.
DS1—Green LED (Radio Shack 276-022).
DS2—Yellow LED (Radio Shack 276-021).
Q1—MJS2955 PNP power transistor (Radio Shack 276-2043 or equivalent; see text).
R1—1.4 kΩ (see text).
R_{SENSE}—1 Ω (see text).
S1—SPST switch (Radio Shack 275-612).
U1—Maxim MAX713CPE single-chip battery charger.
Misc—The wall-cube power supply used in this project is a Radio Shack 273-1652. The charger is housed in a plastic project box (Radio Shack 270-627).

value of 180 minutes).

The PGM pins for our design need to be connected as follows (6-cell pack with time-out at 180 minutes):

Pin	Connect To
PGM0	NO CONNECTION
PGM1	NO CONNECTION
PGM2	REF
PGM3	BATT–

When the Smart Charger is turned on, the MAX713 charger circuit is connected to the battery, and fast charge begins. When the voltage peaks or the time-out is reached the charger reverts to trickle charge.

Construction and Checkout

Simply wire the circuit according to the schematic.

Connect a milliammeter in line at the positive terminal of the battery. Also, connect a voltmeter across the positive and negative terminals of the battery. Place a half-charged battery in the charger and turn on the power; the **FAST CHARGE** indicator should light.

The voltage will slowly rise and the milliammeter should indicate the maximum charge rate as designed (250 mA in our example). After the battery voltage peaks or the fast charge time-out (2.5 hours) has passed, whichever comes first, the **FAST CHARGE** light should go out and the charge rate should decrease to approximately C ÷ 16 (approximately 500 ÷ 16, or 31 mA).

Getting the Parts

All parts except the MAX713 can be obtained from Radio Shack. The 1-Ω resistor is available from Radio Shack in a package with many resistors of different values. Some parts were selected only because they are standard Radio Shack stock. For example, the 100-W PNP transistor (Q1) is overkill; it could be replaced with a smaller device.

I used a 6-cell battery holder, which cannot be obtained at Radio Shack. Radio Shack sells a variety of cell holders, though. To charge six cells in series, use a 4-cell and a 2-cell holder. Like my 6-cell holder, these holders use a 9-V battery connector, also available from Radio Shack.

The MAX713 chip or a complete set of parts including the chip are available from the author.[1]

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[1] The following parts are available from:

Simple Design Implementations, PO Box 9303, Forestville, CT 06011-9303 203-582-8526

1. Maxim MAX713CPE single chip battery charger: $10 + $1.50 shipping and handling.
2. Complete kit of parts, including perforated board, case, wall adaptor: $30 plus $3.00 shipping and handling.

New Books

EMERGENCY RADIO!

By Norm Schrein, KA8PGJ

Published by Index Publishing Group Inc, 3368 Governor Dr, Suite 273F, San Diego, CA 92122; tel 800-546-6707 or 619-281-2957; fax 619-281-0547. First edition, 1994, paperback, 8½×5½ inches, 214 pp, B&W photographs. $14.95 plus $3 s/h.

Reviewed By Brian Battles, WS1O
QST *Features Editor*

"*It's a dessert topping...It's a floor wax...*" The old *Saturday Night Live* "Shimmer" ad-parody skit comes to mind when I try to think of how to describe this book. "*It's a frequency directory...It's a public service communication guide....*" Surprise! It's a little of both, but it doesn't cleanly fit either category. When I first saw it, I expected it to be another tourist guide to scanning police and fire radio, with lists of frequencies, monitoring tips, etc. I was wrong. It's not a typical scanner book in the traditional sense. Author Norm Schrein, KA8PGJ, takes us to the other side of the scanner hobbyist's world. He profiles law-enforcement and public safety officers in a sort of "you are there" style. The radio portion of the book is, in fact, sometimes underplayed.

The subtitle is *Scanning news as it happens*...and that's an appropriate description. This is the printed counterpart to real-life action TV shows like *Cops*, *Rescue 911* and so on. Norm rides with the police officers, firefighters, emergency medical technicians and volunteers as they patrol our communities and respond to calls. Each chapter focuses on one particular local agency, with titles that include Headshot in a Heartbeat—The Cincinnati SWAT Team, The Gangs of Ft Worth, Hams and the Miamisburg Disaster, EMS Indy Style, Women in Police Work, Saginaw Saturday Night, Working Side-By-Side, A Tampa Fire and Rescue Story, Ham Radio—Nothing Amateur About It, Hoover Dam—The "Dam Police," Security at Both Ends of Pennsylvania Avenue, and America's Front Line.

There's a fair amount of radio-related info and lists of interesting frequencies, but this is primarily a book about the work these people do. Radio communication is discussed mainly in terms of how it's used as a part of general operations of various agencies, not as the principal topic. Some topics are fascinating to read about. It's not all cops and robbers—although there's a fair share of them here—but Norm also brings you along on visits with traffic reporters for broadcast stations, US Border Patrol agents, Major League baseball teams, Las Vegas casinos, the US Postal Inspection Service, federal prison personnel, marine services and more. This is unusual and interesting stuff that you just don't often get to read about.

Any book has shortcomings, however, and this one has a few. First, packing 36 chapters into a 214-page book means that the chapters are all short—most only two or three pages—and you'll find yourself wishing some of them were longer and delved into more detail. There's also a distressing lack of an index—surprising in a book as thoughtfully researched as this. I'd have liked an appendix or two giving frequency listings and specific data on what to monitor. This information is sprinkled throughout the text, so you do get the info, it's just not all in one convenient place. Besides, there are dozens of "complete" (and redundant) scanner frequency guides available from other sources, so I guess the publisher felt that most scanner enthusiasts will go elsewhere if they want books chock-full of lists, directories and tables.

There's no doubt that the author is an established authority on the subject of monitoring emergency and public service communications. Billed as "Mr Scanner," Norm edits the *Betty Bearcat Frequency Directories* and the *Fox Scanner Radio Listings Directories*, and his credits include serving as president of the Bearcat RC, as a former police officer and as a firefighter. Even former US Senator Barry Goldwater, K7UGA, who's profiled in a chapter entitled A Radio Pioneer, provided a quote for the book's back cover: "An excellent job..."

Well, Senator, I do believe I agree.

Exploring the Internet—*Part 1*

By Steve Ford, WB8IMY
Assistant Managing Editor
(Internet address: sford@arrl.org)

WELCOME TO CYBERSPACE

Someone once said that you can examine the most intricate details of the human brain and never find the mind. You can also disassemble every computer in the world and never find *cyberspace*.[1] Like the mind, cyberspace is more than simply the sum of its parts. It's a vast community with millions of inhabitants. Cyberspace is an alternate reality, a kind of global consciousness that our computer networking technology has created. It's intangible—just like the mind—but very real nonetheless.

The denizens of cyberspace are people like you and me. Their thoughts, feelings and knowledge enter cyberspace at their keyboards and are available to every other "cybernaut" in the world. You could be "jacked into" (more *cyberspeak*) a huge Cray mainframe or a Commodore 64—it doesn't matter. The computer type is unimportant as long as you can use it to access whatever network is available to you.

Information flows through cyberspace at up to 45 *million* bits per second. (Soon it will flow at rates exceeding 2 *billion* bits per second.) The never-ending data stream travels through an electronic nervous system composed of telephone lines, fiber-optic circuits, satellite links and so on. If cyberspace is the global mind, the physical network is the brain and body, better known as the *Internet*.

The Information Superhighway

When you hear politicians and media personalities speak of the Information Superhighway, they're often talking about the Internet. Thinking of the Internet as a highway isn't a bad idea. Imagine eight-lane freeways connecting large cities. From these large cities you find smaller freeways linking small towns.

The eight-lane highways comprise the Internet *backbone*. Connected to this are computers that transfer data at high speeds. Connected to the backbone are smaller networks serving particular geographic regions (exits off the highway). Feeding off these are even smaller networks or individual computers. The ARRL Headquarters local area network (LAN) is one of these.

There's no central computer at the core of the Internet. Instead, the Internet is a composite creature made of thousands of systems scattered throughout the world. How many computers and networks comprise Internet? No one knows. Some people say that there are as many as 5000 networks connecting nearly 2 million computers and more than 15 million people.

How Does it Work?

Data moves through the Internet using *packet-switching* techniques. That is, pieces of data are transported in packets of a specific size. The packets are handed off from one network and computer to another like batons in a relay race. Like amateur packet radio, Internet packets are occasionally retransmitted if the data is corrupted at any point along the way.

Every computer connected to the Internet has a unique address. This address is included in every data packet sent to a particular destination. By analyzing the address, the network "knows" how to route each packet efficiently. In addition, the flow of data between networks and computers is automatically managed according to how busy the system is at the moment, speeding up or slowing down as necessary. The system used to move data through the Internet in this fashion is known as Transmission Control Protocol/Internet Protocol, or *TCP/IP*.

By tapping into Internet, you open a world of amazing possibilities. You can:

✓ Exchange electronic mail (e-mail) with friends around the world. Internet e-mail moves at high speed and is relatively reliable. You can use Internet e-mail to contact anyone on the ARRL Headquarters staff. I'll tell you how next month in Part 2 of this series.

✓ Enter the USENET *newsgroups* and participate in discussions on almost any topic imaginable. There are several newsgroups devoted to Amateur Radio topics.

✓ Tap into thousands of information data bases and libraries worldwide. The ARRL maintains an *Infoserver* with a large inventory of valuable information. In Part 2 of this series I'll give you a step-by-step description of how to use this service via the Internet.

✓ Retrieve thousands of documents, journals, books, computer programs, images and sound files.

✓ Operate distant computers by remote control. One of the Internet's earliest capabilities was to allow researchers, students and scientists to run programs on powerful processors at distant computers to obtain results beyond the capabilities of their own local data-processing equipment. Clever applications have been developed that allow you to play games, retrieve call sign directory information and even determine a geographic location's latitude and longitude.

Amateur Radio TCP/IP

Packet operators who use TCP/IP are operating on a pseudo-Internet, although they may not know it. The group of programs hams call TCP/IP is actually an adaptation of Internet TCP/IP. If an Internet user saw an amateur TCP/IP system in

[1] *Cyberspace* is a word derived from the term *cybernetics* (which uses the prefix *cyber-*, Greek for *to steer* or *govern*), the science of augmenting living organisms with mechanical or automatic devices to perform specialized or enhanced control functions. Cyberspace is, therefore, a fanciful expression that describes humans operating in an "electronic universe" where they use electronic computers to communicate and control remote devices by transmitting data via digital links (over telephone wires, fiber-optic cables, RF circuits, etc).

operation, he or she would recognize it right away.

As a group, the amateur TCP/IP networks are known as *AMPRNET* (Amateur Packet Radio Network). Not all AMPRNET networks connect to the Internet, however. Many provide coverage throughout a state or region, but do not tap the vast global network.

On the other hand, some AMPRNET networks *do* connect to the Internet superhighway. For example, you might be able to connect to a packet station that's operating a *gateway* to Internet. Depending on how it is configured, you can use a gateway to send e-mail, transfer files or even enjoy live discussions with hams across the globe. Everything sent to and from the gateway is dispatched via the Internet backbone. The result is a tremendous increase in speed and coverage that would be impossible on normal packet-radio networks.

Getting started on amateur TCP/IP isn't difficult. If you own a VHF or UHF transceiver and a Terminal Node Controller (TNC), you're almost there. You only need TCP/IP software for your computer and an AMPRNET address. The software is available free from many sources and there's a coordinator in your state or region who issues addresses. The details of amateur TCP/IP go beyond the scope of this article, so I recommend you pick up a copy of *NOSIntro* by Ian Wade, G3NRW. It's available from your favorite dealer or directly from the ARRL. (See the *ARRL Publications Catalog* elsewhere in this issue.)

Amateur Radio TCP/IP is fun, but it's not the best way to access the Internet. Ham networks are slower, so transferring even moderate amounts of information can be a painful exercise. In addition, there are restrictions on how Internet gateways may operate according to FCC third-party traffic rules. If you want to experience the full benefits of the Internet, you need a more direct connection.

Merging onto the Information Superhighway

Your journey through cyberspace begins with your computer. Although you can use just about any computer to access the Internet, I recommend a modern machine with a large hard disk. You'll need the storage capacity for all the information you're going to grab off the network. An efficient, up-to-date computer makes the job much easier.

The next item you must consider is a *modem*. The modem is the interface between your computer and your telephone line. Speed is essential. I consider 2400 bit/s to be the *minimum* requirement. A 9600 bit/s modem is much better. If you can push the throttle to 14400 bit/s or 28800 bit/s, go for it. When you're connected, time often equals money! The faster your modem, the less time you'll need to spend on line.

The only software you'll need is a terminal program to communicate with your modem.

You can spend as little as a few dollars for a shareware program or hundreds of dollars for sophisticated software. (Some Internet access services require special software. The software is usually provided by the service.)

You'll find Internet connections at colleges and universities, corporations, government agencies, nonprofit organizations, military installations and elsewhere. The place where you work or go to school may have Internet access you can use right now. Talk to your computer department and see what's available.

There are public access sites that offer Internet services for a fee. See the sidebar, "Internet Public Access." If you can't tap a public access site, you can enjoy Internet activities through commercial on-line services. CompuServe, GEnie, Prodigy and MCI offer access to Internet e-mail only. America On-Line and Delphi provide full Internet access, allowing you to transfer files, chat "live" with other users and so on.

You can reach the Internet through BIX (the *Byte* Information Exchange). BIX offers access to the Internet as part of its basic service. For information, call 800-695-4775. PSI, based in Reston, Virginia, provides nationwide access to Internet services to owners of IBM-compatible computers. Special software is required, but it's available free from PSI. For information, call 800-82PSI82 or 703-620-6651.

Internet Public Access

The following sites provide public access to Internet (usually requiring a monthly or yearly charge). You may not need to live in or near the cities shown to use many of these services. Call the numbers shown for information

Alabama

Nuance, tel 205-533-4296

Alaska

Anchorage. University of Alaska Southeast, *Tundra Services*, tel 907-465-6453

Alberta

Edmonton, *PUCNet* Computer Connections, tel 403-448-1901

Arizona

Phoenix, *Internet Direct*, tel 602-324-0100
Tucson, *Data Basics*, tel 602-721-1988
Tucson, *Internet Direct*, tel 602-274-0100

British Columbia

Victoria Free-Net, tel 604-389-6026

California

Berkeley, *Holonet*, tel 510-704-0160
Cupertino, *Portal*, tel 408-973-9111
Irvine, *Dial N' CERF USA*, See San Diego
Los Angeles/Orange County, *Kaiwan Public Access Internet*, tel 714-638-2139
Palo Alto, *Institute for Global Communications* (local conferences on environmental/peace issues), tel 415-442-0220
San Diego, *Dial N' CERF USA*, tel 800-876-2373
San Diego, *CTS Network Services*, tel 619-637-3637
San Francisco, *Pathways*, tel 415-346-4188
San Jose, *Netcom* (Maintains archives of USENET postings), tel 408-554-UNIX
Sausalito, *The Whole Earth 'Lectronic Link (The WELL)*, tel 415-332-4335; recorded message about the system's current status, tel 800-326-8354 (continental US only)

Colorado

Colorado Springs/Denver, *CNS*, tel 719-592-1240
Colorado Springs, *Old Colorado City Communications*, tel 719-632-4848
Golden, *Colorado SuperNet* (available only to Colorado residents. Local dial-in numbers available in several Colorado cities), tel 303-273-3471

District of Columbia

The Meta Network, tel 703-243-6622
CapAccess, tel 202-994-4245

Florida

Tallahassee, *Tallahassee Free-Net*, tel 904-488-5056

Illinois

Champaign, *Prairienet Free-Net*, tel 217-244-1962
Chicago, *MCSNet*, tel 312-248-UNIX
Peoria, *Peoria Free-Net*, tel 309-677-2544

Maryland

Baltimore, *Express Access*, tel 800-969-9090
Baltimore, *Clarknet*, tel 410-730-9765

Massachusetts
Brookline, *The World* (large collection of MS-DOS files, "Online Book Initiative" collection of electronic books, poetry and other text files), tel 617-739-0202
Lynn, *North Shore Access*, tel 617-593-3110
Worcester, *NovaLink*, tel 800-274-2814

Michigan
Ann Arbor, *MSEN*, tel 313-998-4562
Ann Arbor, *Michnet*, tel 313-764-9430

New Hampshire
MV Communications Inc, tel 603-429-2223

New Jersey
New Brunswick, *Digital Express*, tel 800-969-9090

New York
New York, *Panix*, tel 212-877-4854
New York, *Echo*, tel 212-255-3839
New York, *MindVox*, tel 212-989-2418
New York, *Pipeline* (has graphical interface for Windows), tel 212-267-3636
New York, *Maestro*, tel 212-240-9600

North Carolina
Charlotte, *Vnet Internet Access*, tel 704-374-0779

Ohio
Cleveland, *Cleveland Free-Net*, tel 216-368-8737
Cleveland, *Wariat*, tel 216-481-9428
Dayton, *Freelance Systems Programming*, tel 513-254-7246
Lorain, *Lorain County Free-Net*, tel 216-366-4200

Ontario
Toronto, *UUNorth*, tel 416-225-8649
Toronto, *Internex Online*, tel 416-363-8676

Oregon
Portland, *teleport*, tel 503-223-4245

Pennsylvania
Pittsburgh, *telerama*, tel 412-481-3505

Quebec
Montreal, *Communications Accessibles Montreal*, tel 514-931-0749

Rhode Island
Providence/Seekonk, *Anomaly*, tel 401-273-4669

Texas
Austin, *RealTime Communications*, tel 512-451-0046
Dallas, *Texas Metronet*, tel 214-705-2900
Houston, *The Black Box*, tel 713-480-2684

Virginia
Norfolk/Peninsula, *Wyvern Technologies*, tel 804-622-4289

Washington
Seattle, *Halcyon*, tel 206-955-1050
Seattle, *Eskimo North*, tel 206-367-7457.

Next Month

In the next issue we'll get down to the nitty-gritty of how to contact the ARRL via the Internet and explore the services we have to offer. In Parts 3 and 4, Scott Ehrlich, WY1Z, will show you how to "ftp" (transfer) files, go digging with *gopher* and more. Until then, see you in **CYBERSPACE**.

Cyberspace Reading

B. Aboba, *The Online User's Encyclopedia: Bulletin Boards and Beyond*. (Reading, MA: Addison-Wesley, 1994).

D. P. Dern, *The Internet Guide for New Users*. (New York: McGraw-Hill, 1994).

A. Engst, *Internet Starter Kit for Macintosh*. (Indianapolis: Hayden Press, 1993).

P. Gilster, *The Internet Navigator*. (New York: John Wiley, 1993).

H. Hahn, and R. Stout, *The Internet Complete Reference* (Berkeley: Osborne McGraw-Hill, 1993).

E. Krol, *The Whole Internet: User's Guide & Catalog* (2nd ed. Sebastapol, CA: O'Reilly & Associates, 1994).

T. LaQuey, *The Internet Companion Plus: A Beginner's Start-Up Kit for Global Networking* (Reading, MA: Addison-Wesley, 1993).

R. Smith and M. Gibbs, *Navigating the Internet* (Carmel, Indiana: SAMS Publishing, 1993).

J. R. Levine and C. Baroudi, *The Internet for Dummies* (San Mateo, California: IDG Books Worldwide Inc).

New Books

NATIONAL RADIO CLUB AM RADIO LOG, 14th EDITION

By Ken Chatterton

Published by National Radio Club, PO Box 164, Mannsville, NY 13661-0164; 350 pp, 8½x11 inches, looseleaf in a three-ring binder, $19.95.

Reviewed By Kirk Kleinschmidt, NT0Z
Assistant Managing Editor

As hams, it's fair to assume that we have a soft spot for good old AM broadcast radio. In fact, I'd bet that almost everyone who has a modern ham rig—with receive coverage of the AM and shortwave bands—has tuned the broadcast band a time or two in search of a favorite talk show or some far-away hometown station. Come on, you can admit it!

Like many youngsters in the '70s (and before, perhaps today, too) I got started in radio by trying to DX AM broadcasters. I'd stay up past my bedtime, tuning stations on my green Panasonic "round ball" radio, thrilled each time I'd come across a station I could identify (usually biggies such as WLS, KOA, KSL, WCCO, CKLW and occasionally, XEROK, the Mexican powerhouse on 800 kHz, a stone's throw across the border, running *big* power).

IDing the many stations, especially those on the upper end of the dial, where lots of low-power stations slug it out on the same frequencies, was difficult. A dog-eared copy of an older *White's Radio Log* (I think that's what it was called, anyway), helped, and provided QSLing addresses.

As a 12-year-old soon-to-be-ham, building a collection of broadcast-band QSL cards was fascinating. Every time I look at those cards today, I remember how exciting it was to tune a station, find it in the station listings, send away for a QSL card and run to the mailbox every day after school looking for cards.

So what's my point? NRA's *AM Radio Log* is a modern version of just such a book—only better! Its more than 350 pages (looseleaf in a plastic binder) list every AM broadcast station in North America, with frequencies, format, network affiliation, power levels, antenna patterns and day/night power levels and complete addresses for QSLing! There are 6000 listings sorted by city, state, frequency and call sign, so information on just about any station is easy to find.

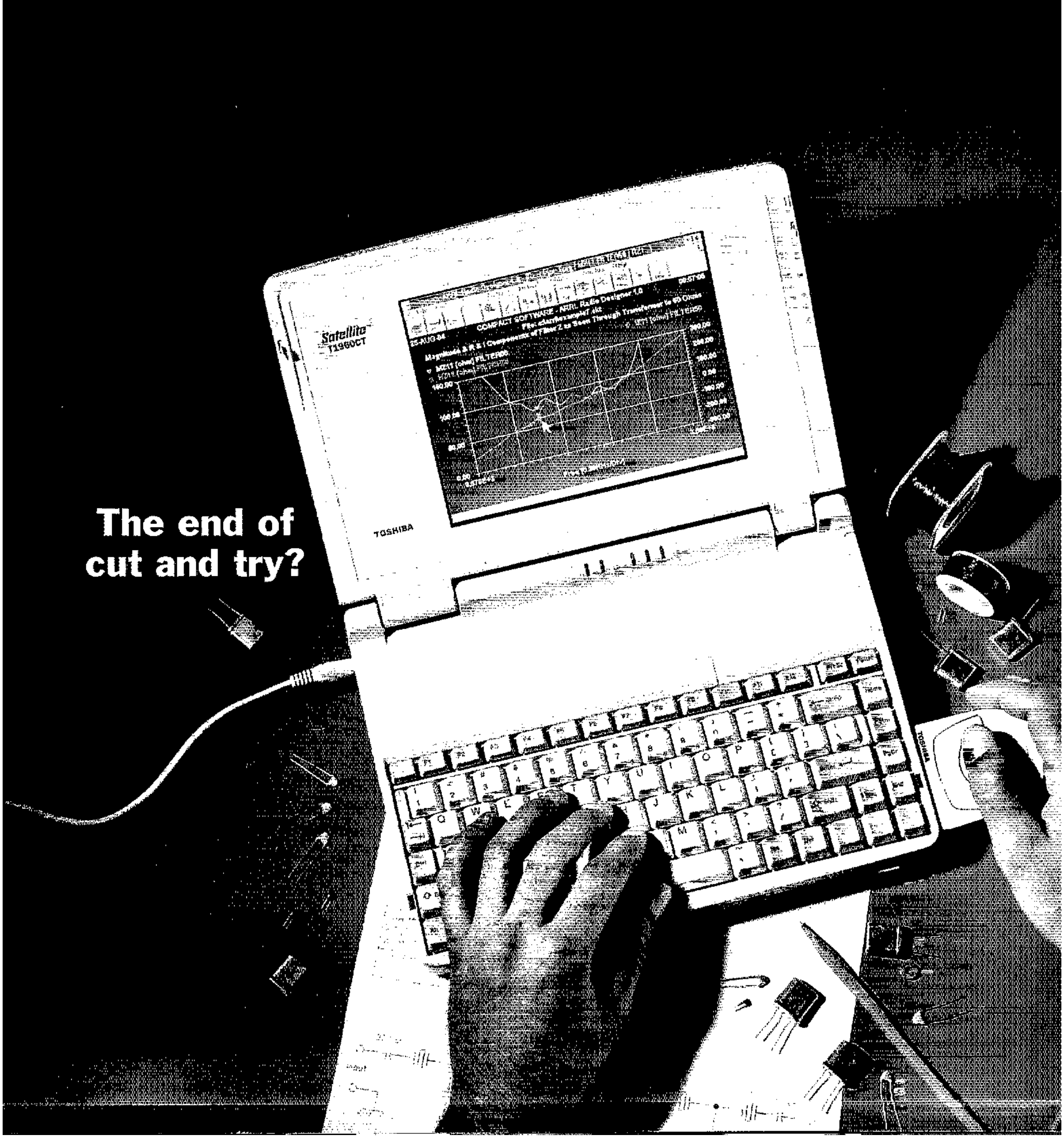
In 1994, as in the '70s, DXing the broadcast band is a lot of fun. As the sunspot cycle slowly bottoms out, activity on the HF ham bands declines, while propagation "down under" improves.

So spin your VFO down to the broadcast band and check things out. One thing's for sure, *AM Radio Log* will dramatically improve your BCB listening efforts. The 14th edition also happens to celebrate the 60th anniversary of the National Radio Club.

QST

October 1994

devoted entirely to Amateur Radio



The end of cut and try?

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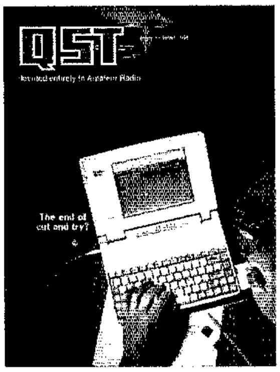


OUR COVER

Computerized circuit simulation and analysis help bring many of today's best radio designs to life. Now *ARRL Radio Designer 1.0*, a new *Windows*-based RF CAD program for radio amateurs, lets *you* put these techniques to work. The story begins on page 21.

CONTENTS

October 1994
Volume LXXVIII Number 10

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

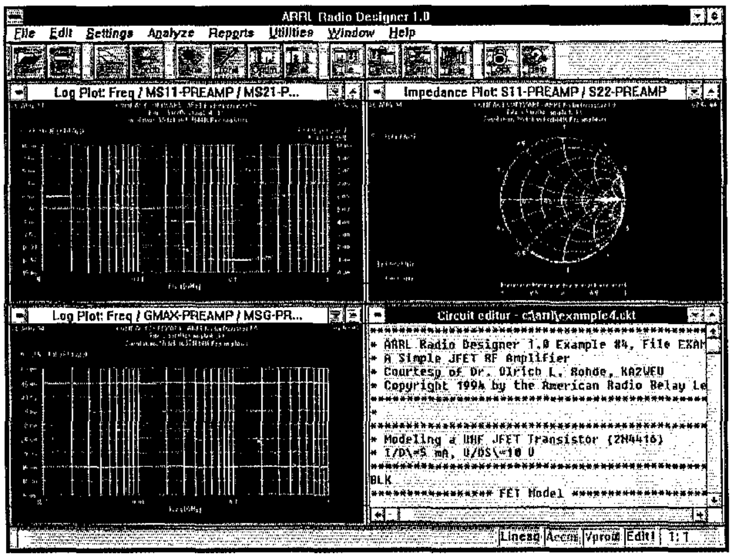
OPERATING

DEPARTMENTS

Introducing *ARRL Radio Designer*: New Software for RF Circuit Simulation and Analysis

If you've done any hobby computer-aided design at all, you've probably used CAD mainly for designing and enhancing your antenna system. Now you can put your PC to work modeling radio circuitry at the *station* end of your feed line!

By David Newkirk, WJ1Z
Senior Assistant Technical Editor



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**********************************************
* ARRL Radio Designer 1.0 Example #4, File EXAM
* A Simple JFET RF Amplifier
* Courtesy of Dr. Ulrich L. Rohde, KA2WEU
* Copyright 1994 by the American Radio Relay Le
**********************************************
*
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* Modeling a UHF JFET Transistor (2N4416)
* I/D\=5 mA, V/DS\=10 V
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BLK
***************** FET Model ******************
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A glance into just about any current Amateur Radio magazine or club bulletin confirms that computerized antenna modeling now qualifies as a standard ham activity. With program names like *NEC*, *MiniNEC*, *MN*, *ELNEC* and *ARRL MicroSmith* well-established as household words—in *ham* households, at least—we don't even blink when someone gloats over another computer-optimized Yagi or gulps when modeling unmasks a new skywire as more of a worm warmer than an ether buster.

What, then, keeps so many of us from modeling radio *circuits* with our computers—designing, simulating and analyzing the innards of the "gray boxes" we connect to the antenna systems we model with such enthusiasm? Availability, for starters. If versatile, affordable RF CAD software exists, how do we find it? Even the worthiest of the uncountable neat little (and not so little) utilities written by hams to solve or simulate or design particular radio-electronics problems or circuits rarely makes headlines.[1]

Program suitability to the task of realistic RF modeling is perhaps the biggest hurdle. General-purpose simulators like *PSPICE* and *MicroCAP* are well-established in college-level EE programs. They are offered as low-cost, general-purpose simulators, but the accuracy of their available active-device libraries, especially in the areas of noise (noise-correlation matrix calculations) and distributed parasitic reactances so important in RF modeling, significantly limits their usefulness above a 100 MHz. What's more, these programs don't "speak RF"—they're not equipped to report circuit performance in RF-standard terms like S (scattering), Y (admittance), impedance (Z) and other network parameters.

Now there's a new choice. Working in association with Compact Software of Paterson, New Jersey, ARRL is proud to unveil *ARRL Radio Designer 1.0*—realistic, affordable (price class, $150) RF CAD *Windows* software for radio amateurs![2]

What is *ARRL Radio Designer*?

ARRL Radio Designer, a derivative of *Super-Compact*, Compact Software's industry-standard linear circuit simulator, analyzes the performance of linear, small-signal active and passive dc, AF and RF circuitry, including amplifiers, filters, matching networks and power splitters and combiners. *ARRL Radio Designer* tools include:

- Analysis (prediction of circuit performance);
- Optimization (automatic adjustment of circuit performance to meet goals you specify);
- Voltage Probe (predicts the signal level at any point in a simulated circuit);
- Statistical Analysis (simulates the effect of component value variations [such as those attributable to tolerance or temperature coefficient] on circuit performance using Monte Carlo techniques);
- Time Domain Analysis (simulates circuit performance in response to a steady-state time-domain signal using impulse, step, pulsed carrier or user-defined stimuli);
- Manual matching-network synthesis via Circles, an interactive Smith Chart utility; and
- Databanks (device-manufacturer-supplied S-parameter and noise data you can incorporate in your circuit simulations).

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[1] Notes appear on page 26.

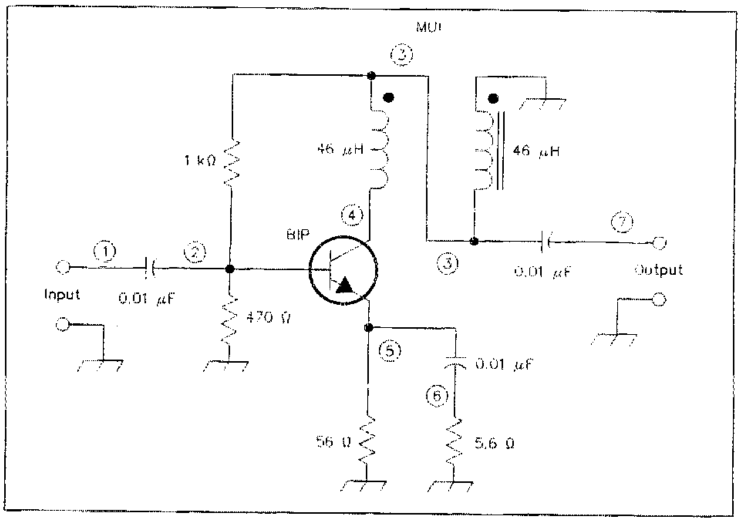
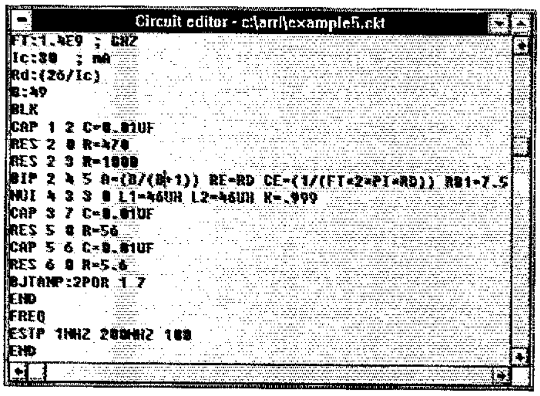


Figure 1—We'll simulate the post-mixer amplifier from the classic Progressive Communications Receiver described in November 1981 *QST* by Wes Hayward, W7ZOI, and John Lawson, K5IRK. The first step in readying this circuit for *ARRL Radio Designer* analysis merely involves marking its *nodes*—its points of component interconnection—with numbers, *0* being the default for circuit common (I circle mine to distinguish them from device pinouts, wire designators, and so on). How come no power supply connections? The "*ARRL Radio Designer* Versus Professional Simulators" sidebar tells why.



```
FT:1.4E9 ; GHZ
Ic:30  ; mA
Rd:(26/Ic)
B:49
BLK
CAP 1 2 C=0.01UF
RES 2 0 R=470
RES 2 3 R=1000
BIP 2 4 5 A=(B/(B+1)) RE=RD CE=(1/(FT*2*PI*RD)) RB1=7.5
MUI 4 3 3 0 L1=46UH L2=46UH K=.999
CAP 3 7 C=0.01UF
RES 5 0 R=56
CAP 5 6 C=0.01UF
RES 6 0 R=5.6
BJTAMP:2POR 1 7
END
FREQ
ESTP 1MHZ 200MHZ 100
END
```

Figure 2—Next, we use *ARRL Radio Designer's* Circuit Editor—just an ASCII word processor—to specify each of the amplifier's components in network list—*netlist*—form. Highlights in this particular netlist:

- The four lines after the file header contain data for use by the circuit's transistor model. They are FT (f_T, current gain-bandwidth product), Ic (I_C, collector current), Rd (r_d, diffusion resistance) and B (beta, current amplification factor).
- BLK marks the beginning of a netlist *block*.
- The BIP line specifies a bipolar transistor intended to simulate a 2N5109 running at 30 mA of collector current, using equations to derive the transistor's alpha (A) from beta (B/(B+1)) and emitter capacitance (CE) from f_T and r_d (1/(FT*2*PI*RD)). The model's emitter resistance (RE) is set equal to R_d.
- The MUI line specifies mutually coupled 46-µH inductors with a coupling coefficient of 0.999 (the number of 9s in effect sets the number of decades of frequency range);
- The *BJTAMP:2POR 1 7* line names the circuit *BJTAMP* and defines it as a two-port network with terminals at nodes 1 and 7; and
- The *FREQ*uency block tells *ARRL Radio Designer* to simulate BJTAMP's performance at 100 exponentially stepped frequencies from 1 to 200 MHz.

Our netlist completed, we click on *ARRL Radio Designer's* **Analyze** button and *go!*

ARRL Radio Designer reports the results of its simulations in graphical (rectangular and polar) and tabular form, onscreen and via any *Windows*-compatible printer, in terms of

- S, Y, Z, group delay and voltage probe parameters for *n*-port networks;
- chain (ABCD), hybrid (H), inverse hybrid (G), gain, voltage gain, and stability parameters for two-port networks;
- magnitude of reflection coefficient, phase of reflection coefficient, VSWR and return loss parameters for one-port networks;
- gain, gain matching and noise parameters; and
- complex S, Y, Z, H, G, chain (A), gain matching, noise matching and voltage probe parameters.

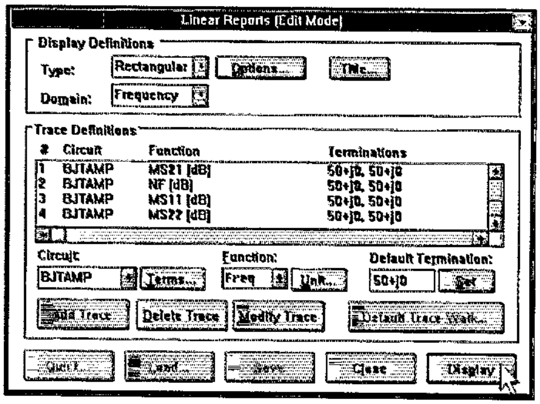
Installing and running *ARRL Radio Designer* requires, at minimum: a 386, 486, 586 or Pentium IBM PC or 100% compatible computer (math coprocessor strongly recommended); 8 Mbytes of RAM; a 3.5-inch, high-density floppy drive; a hard disk with at least 5 Mbytes of free space; *Microsoft Windows* 3.1 or higher; and a mouse or equivalent pointing device.

So much for the fact dump. You've got to see *ARRL Radio Designer* in action to appreciate it, so let's put it work!

ARRL Radio Designer in Action

We'll use the circuit shown in Figure 1—the post-mixer amplifier from Hayward and Lawson's Progressive Communications Receiver—as our first test case. Reduced to its essentials, the process involves just four steps:

1. Mark each of the circuit's *nodes*—points of interconnections between its components or *elements* (Table 1 lists



Type: Rectangular
Domain: Frequency

#	Circuit	Function	Terminations
1	BJTAMP	MS21 [dB]	50+j0, 50+j0
2	BJTAMP	NF [dB]	50+j0, 50+j0
3	BJTAMP	MS11 [dB]	50+j0, 50+j0
4	BJTAMP	MS22 [dB]	50+j0, 50+j0

Circuit: BJTAMP
Function: Freq
Default Termination: 50+j0

Figure 3—The analysis done, we pop *ARRL Radio Designer's* Linear Reports dialog to specify the simulated performance parameters we want to see (the S parameters MS_{21} [magnitude of forward gain], MS_{11} [magnitude of input reflection] and MS_{22} [magnitude of output reflection], and NF [noise figure]), and how we want them displayed.

Table 1
ARRL Radio Designer Circuit Elements

Passive Lumped Elements	
CAP	Capacitor models
DIOD	Diode models
DLY	Time delay
IND	Inductor models
MUI	Coupled inductor models
PLC	Parallel connection of inductor and capacitor
PRL	Parallel connection of resistor and inductor
PRC	Parallel connection of resistor and capacitor
PRX	Parallel combination of resistor, inductor and capacitor
RES	Resistor models
SHO	Short circuit
SLC	Series connection of inductor and capacitor
SRL	Series connection of resistor and inductor
SRC	Series connection of resistor and capacitor
SRX	Series connection of resistor, inductor and capacitor
TRF	Transformer models (ideal two- and three-winding types)
Black-Box Elements	
IMP	Two-terminal impedance
ONE	Two-terminal element specified by admittance, impedance or reflection coefficient
TWO	Three-terminal two-port specified by admittance, impedance or S parameters
Distributed Elements	
CAB	Coaxial cable models
TRL	Transmission-line models
Controlled Source and Active Elements	
CCG	Current-controlled current source
CVG	Current-controlled voltage source
BIP	Bipolar transistor model
FET	Field-effect transistor model
OPA	Operational amplifier
VCG	Voltage-controlled current source
VVG	Voltage-controlled current source

ARRL Radio Designer's entire set) with an exclusive number between 0 and 999 (Figure 1);

2. Type the circuit's elements and node numbers into a *netlist* using *ARRL Radio Designer*'s Circuit Editor (Figure 2);

3. Press *ARRL Radio Designer*'s **Analyze** button; and

4. Graph, table, save and/or print the results to your heart's content (Figures 3, 4 and 5).

A 6-dB attenuator follows this post-mixer stage in the Hayward/Lawson receiver. With it in line, they reported the amplifier's gain as "about 16 dB." Subtracting 6 dB from the MS_{21} trace in Figure 4 or Figure 5 reveals that our model's HF/VHF gain agrees quite closely with the findings of Hayward and Lawson.

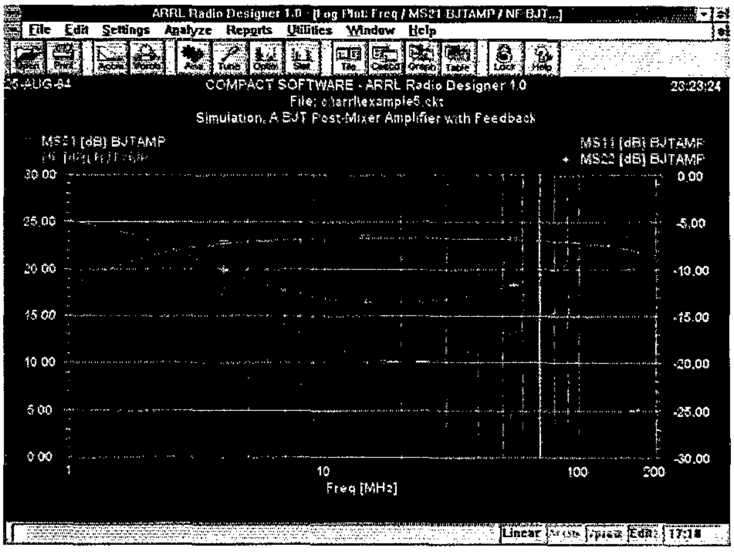


Figure 4—How *ARRL Radio Designer* delivers the goods onscreen. You can control the appearance of everything you see in an *ARRL Radio Designer* report—colors, fonts, titles, graph scaling, the works.

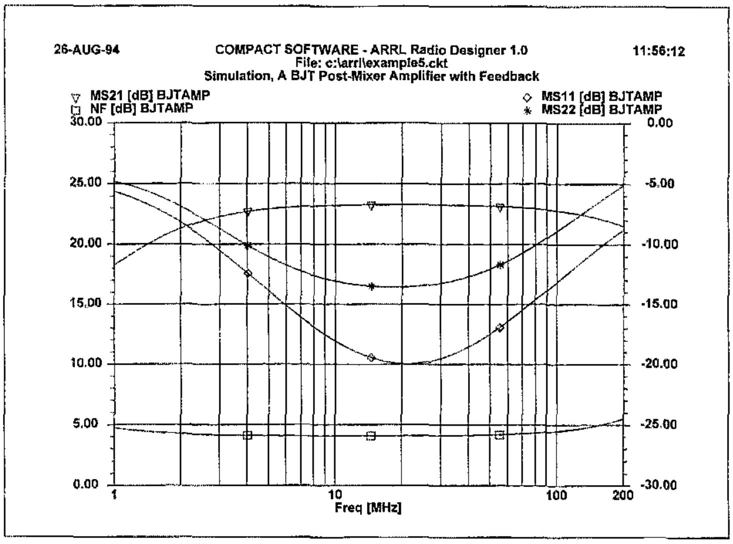
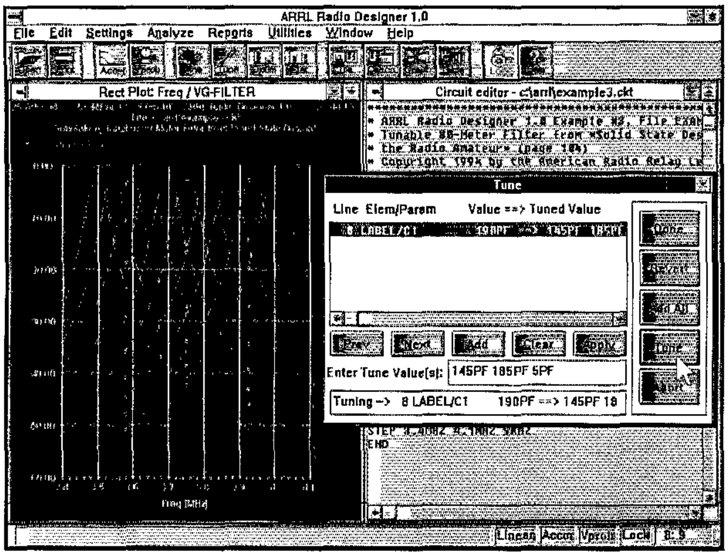


Figure 5—Need hard copy? Here's how the Figure 4 report rolls out of a laser printer in portrait orientation. *ARRL Radio Designer* gives you wide control over its printer output, including line weights, fonts and color. (We don't have a color printer, so I set all of the report traces' colors to black for this version.)

Circuit Tuning

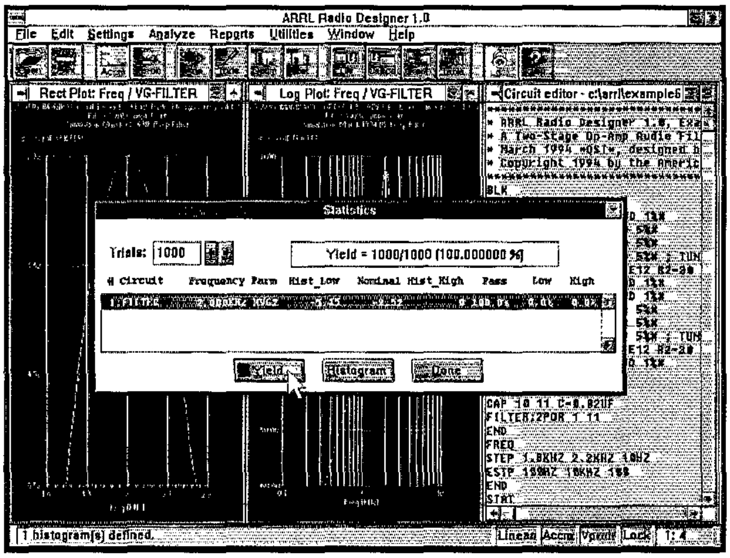
How does a double-tuned-circuit filter's response shift as you tune it with a two-section capacitor? Sure, you can *hear* the peak sweep by if the filter's in your receiver, but what does that effect *look* like? If you happen to own a spectrum analyzer and tracking generator, you can demonstrate it easily enough. Or you can let *ARRL Radio Designer*'s Tune feature show you (Figure 6).



Enter Tune Value(s): 145PF 185PF 5PF

Tuning --> 8 LABEL/C1 190PF ==> 145PF 18

Figure 6—Here's what happens when *ARRL Radio Designer*'s Tune feature steps a double-tuned-circuit filter's two-section tuning capacitor through part of its range in 5-pF increments. The filter simulated (a bottom-coupled 80-meter design from *Solid State Design for the Radio Amateur*) uses a small inductor for coupling. The coupling inductor's reactance, and therefore the coupling between the filter's resonators, increases with frequency, and you can see the effect of this as a barely perceptible insertion-loss decrease as the filter sweeps from 3.5 to 4.0 MHz. When you activate Tune, *ARRL Radio Designer* switches into Accumulate mode to let you see up to 20 simulations at once.



Trials: 1000

Yield = 1000/1000 (100.000000 %)

1 histogram(s) defined.

Figure 7—You can use *ARRL Radio Designer*'s Statistics feature to explore the effects of component tolerances on circuit behavior. In this analysis, *ARRL Radio Designer* evaluated 1000 iterations of an audio filter design while varying its frequency-critical component values within specified tolerances, pass/fail-testing each trial on the basis of loss at 2 kHz. The target loss range was 0 to −5.52 dB.

Circuit Analysis, Circuit Synthesis—What's the Difference?

ARRL Radio Designer concentrates on circuit *analysis* as opposed to circuit *synthesis*, but what does that mean? With a circuit synthesis program, you define a problem ("Build me a fifth-order Butterworth low-pass filter with 50-Ω terminations, 0.1 dB of ripple and a 3-dB corner frequency of 230 MHz") and the program responds with appropriate component values. Circuit analysis capability, if present, is usually limited to relatively simple simulations of the program's own solutions.

A circuit *analysis* program, on the other hand, predicts the performance of *your* solutions. You enter your circuit into the program in coded form (commonly, as in *ARRL Radio Designer*, this is a text file called a *netlist*, short for *network list*) and provide guidance for the program's simulation engine ("Calculate this circuit's behavior at 511 exponentially stepped frequencies from 100 kHz to 100 MHz, and graph the real and imaginary components of its input impedance with its output terminals loaded by 1.5 kΩ in parallel with 3 pF"). The appearance, accuracy and quality of the results you get depend on how completely you specify (and how completely the program *lets* you specify) your circuit's component characteristics and terminations, the accuracy of the program's mathematical component models, and the program's reporting capabilities.

ARRL Radio Designer goes far beyond plain-vanilla analysis, however. Its Circles utility can help you synthesize matching network values, Smith Chart style. You can command its Tune function to step selected components' values through sequences or ranges of values as you watch the results onscreen. You can predict the effect of component tolerances on circuit performance using Statistical Analysis. Even more excitingly, you can put *ARRL Radio Designer*'s Optimizer to work tweaking your circuit for peak performance—hand it your best cut at a low-noise 2-meter preamp, say, and walk away with a design that's a dB or two quieter. True to its name, *ARRL Radio Designer* does do "just synthesis"—it helps you design radios.—*WJ1Z*

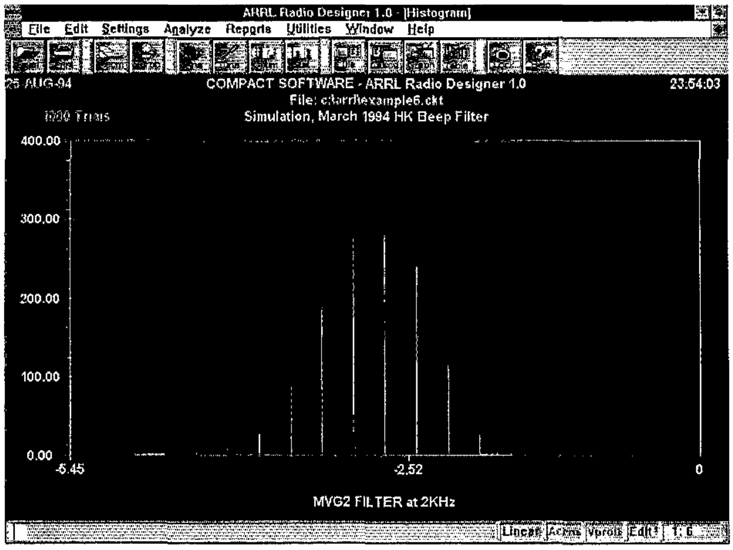


Figure 8—Clicking on Histogram charts how the filter's 2-kHz loss varies around the nominal value of –2.52 dB across a population of 1000.

Statistical Analysis, Monte Carlo Style

Real components vary in value with temperature and tolerance. Using *ARRL Radio Designer*'s statistical analysis feature, you can determine the effect of these variations on circuit performance, as I did for the op-amp audio filter described by Henry J. Perras, K1ZDI, in March 1994 Hints and Kinks. Figure 7 reports the happy news that 1000 out of 1000 K1ZDI filters built with 5%-tolerance capacitors and 1%-tolerance resistors should work acceptably well with *no post-construction tweaking whatsoever*, and Figure 8 shows how those 1000 filters' 2-kHz losses vary around the nominal value of –2.52 dB as a result of component tolerances.

Putting Optimization on the Case

It's one thing to use "cut and try" techniques to nail down one variable component value in a circuit when you've got all the others under control, but you can just about kiss a weekend's worth of experimenting goodbye if you need to vary more than one. If the optimization job you want to do requires esoteric test equipment—say, a noise figure meter—or presents a "solution surface" so complex that one-dimensional tweak-and-measure investigation is doomed from the start, you and *ARRL Radio Designer*'s Optimization engine stand to become fast friends. Figure 9 shows a simple example of such a challenge—an antenna matcher with one variable inductor and one variable capacitor. Your job: Find the L and C values that let that tuner turn a highly reactive antenna load (1.6 – *j*2256 Ω at 1.83 MHz) into 50 Ω, resistive. Your choice: Put *ARRL Radio Designer* on the case (Figure 10). Your reward: a match (Figure 11).

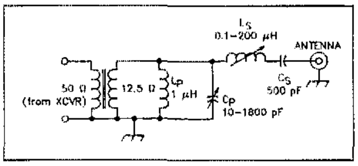
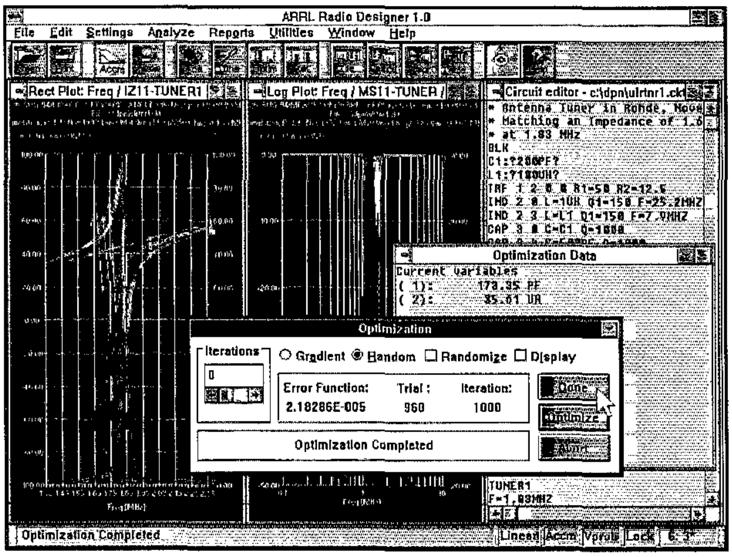


Figure 9—Ulrich L. Rohde, KA2WEU, described this antenna matching network in November 1992 *QST*'s "Recent Advances in Shortwave Receiver Design." Finding the L and C values that let this network match a particular load—assuming that that the load is within the network's matching range—is a piece of cake for *ARRL Radio Designer*'s Optimizer.

Let the Fun and Learning Begin!

That's all the space we can devote this month to *QST*'s first look at *ARRL Radio Designer*. What happens next? For starters, you can expect to see more about *ARRL Radio Designer* in future *QST*'s—whenever we can appropriately use it to confirm or improve a circuit design, or to illustrate a point, for instance. Our overriding hope, though, is that *you* will put *ARRL Radio Designer* to work, and share your experiences and findings with us and your fellow hams. The way we see it, it's only a matter of time before RF-accurate CAD becomes every bit as routine to radio amateurs as simulating an antenna with *MiniNEC* or



```
* Antenna Tuner in Rohde, Nove
* Matching an Impedance of 1.6
* at 1.83 MHz
BLK
C1:?200PF?
L1:?100UH?
TRF 1 2 0 0 R1=50 R2=12.5
IND 2 0 L=1UH Q1=150 F=25.2MHZ
IND 2 3 L=L1 Q1=150 F=7.9MHZ
CAP 3 0 C=C1 Q=1000
```

```
Current Variables
( 1):      173.35 PF
( 2):       35.61 UH
```

```
Iterations: 0
Error Function:    Trial :    Iteration:
2.18286E-005       960        1000
```

```
TUNER1
F=1.83MHZ
```

Figure 10—*ARRL Radio Designer*'s Optimization engine lets you use random and gradient techniques to zero in on complex circuit solutions you couldn't hope to achieve by cut and try. This screen dump catches *ARRL Radio Designer* in the act of adjusting the Figure 9 network to match a 1.6 – *j*2256 Ω load to 50 Ω at 1.83 MHz. (In practice, turning off the Optimizer's display feature greatly speeds a solution because the computer doesn't have to stop to calculate and redraw the graphs each time it tries new values.) The optimization goal: simultaneous achievement of tuner-input Z parameters of RZ_{11}=50 (real part of impedance, 50 Ω) and IZ_{11}=0 (imaginary—reactive—part of impedance, 0 Ω)—"50 Ω, resistive." Time to solution: 1½ minutes on a 33-MHz 486DX computer.

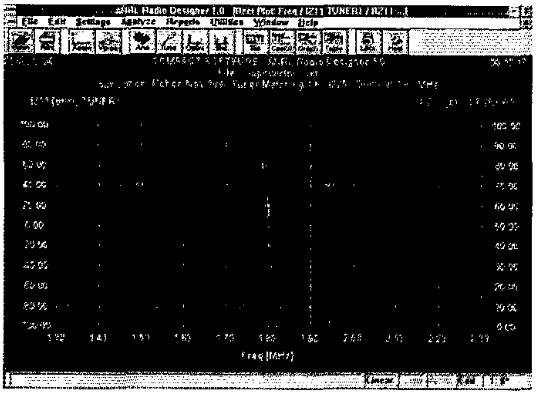


Figure 11—We have a match! (After 1000 iterations, *ARRL Radio Designer* had fought the error function down to a number on the order of 10^{-5}, so I declared the job done.) Now compare these traces' 1.83-MHz values with the optimization goal I stated in the Figure 10 caption.

cruising through a contest with *CT* or *NA*. We're pleased to offer *ARRL Radio Designer* as a doorway into that new phase of ham radio's growing computer tradition.

Notes

[1] Did you catch Dean Straw's "So What's New in *The ARRL Antenna Book*?" (last month's lead article) or check out the companion software available for *The ARRL UHF/Microwave Experimenter's Manual*?

[2] *ARRL Radio Designer* is available from Publications Sales at ARRL HQ for $150, plus $5 shipping/handing (UPS delivery). You can order by phone (203-666-1541) or use the order form in the ARRL Publications Catalog elsewhere in this issue (the order number is 4882). *ARRL Radio Designer* software and example files are shipped on two high-density 3½ inch IBM-compatible diskettes. The instruction manual includes tutorial and reference information. See the article text for computer hardware requirements.

ARRL Radio Designer Versus Professional Simulators

Even circuit-simulation software costing tens of thousands of dollars or more—as the best such programs do—can't accurately model *absolutely every* circuit function in, say, a shortwave ham transceiver. Since even the best simulators can't do everything, what subset of "less than everything" can *ARRL Radio Designer* do? To put it another way, what do professional-grade simulators got that we ain't got? The answer has three parts: schematic capture, microwave/optical capability, and nonlinear simulation.

Schematic Capture

Schematic capture, standard with some circuit simulators and optional in others, lets you draw a schematic onscreen and generate a netlist—the circuit's text equivalent in a form digestible by the simulator—from the drawing. *Super-Compact*, *ARRL Radio Designer*'s big brother, doesn't include schematic capture (Compact's *Serenade* schematic editor is available as an option); its text-based netlist interface was designed for compatibility across platforms with widely variable graphics capabilities. *ARRL Radio Designer* therefore doesn't include schematic capture. For sprawling digital circuits, schematic capture, though tedious, is almost a necessity. For RF applications and simple circuits, however, manual circuit entry is significantly faster and easier.

Microwave/Optical Capability

The lion's share of today's commercial and military radio R & D bucks flows into UHF/microwave and fiber-optics projects. At those frequencies, tuned circuits rarely consist of "lumped" inductances and capacitances—a tuned circuit consisting of 0.0001 pF in parallel with 0.0001 µH resonates at about 1592 GHz, but I dare you to actually build one!—so stripline, microstrip and other transmission-line and waveguidelike structures do the job instead. Professional-grade circuit simulators can model these structures using actual physical dimensions and the real characteristics of conductors, dielectrics and substrates. In contrast to this, *ARRL Radio Designer* models tuned circuits in terms of lumped L and C. Its practical applicability therefore declines wherever radio physics dictates that you must switch from LC circuits to stripline, microstrip, YIGs, dielectric resonators or waveguides—typically above 1 GHz.

Small- Versus Large-Signal Analysis

Oscillators stabilize their amplitudes, mixers mix, modulators modulate, amplifiers distort, rectifiers rectify, frequency multipliers multiply, and AGCed stages "AGC" because of amplitude-nonlinear device behavior. Because *ARRL Radio Designer* is a linear, small-signal simulator, it cannot simulate these large-signal effects.

To get a handle on this, check out the bipolar junction transistor amplifier in Figure 1. That drawing includes no power supply connections—*because the* ARRL Radio Designer *model doesn't need them!* You can successfully analyze the circuit's performance, and plot its gain and frequency response, without connecting its bipolar junction transistor to V_{CC}. How can this be?

This can be because the Figure 1 circuit is a linear amplifier, and *ARRL Radio Designer* is a small-signal, linear simulator. When, working at your radio bench, you power a real transistor at a particular voltage and bias it just so, you also set its supply- and bias-dependent parameters (starting with a bipolar junction transistor's alpha or a FET's transconductance) to particular values. Modeling in *ARRL Radio Designer*, you specify your devices' parameters according to the power-supply and bias levels you expect to exist in the modeled circuit, and *ARRL Radio Designer* then reports your circuit's performance with devices exhibiting those parameters.

The distinction between small-signal and large-signal analysis pretty much boils down to this: A device operates in its small-signal region when its "operating" or "bias point" (which its power-supply and bias levels determine) doesn't shift in response to its input signal. Driving the device with a signal large enough to shift its bias point results in nonlinear, dynamically variable performance that the "hardwired" device parameters of a linear simulator's netlist don't reflect.

ARRL Radio Designer therefore can't model nonlinear effects like frequency translation, intermodulation, AGC, and the results of subjecting good devices, or well-modeled designs, to absurd or greatly divergent bias, drive or power-supply levels. If, for instance, you want to see what happens when you power a 13.8-V amplifier design at 6 V, you can model its gain under the new conditions. Working carefully, you can even simulate the frequency- and phase-response shifts that may occur as a transistor's internal capacitances change in response to different terminal voltages and currents. But you won't be able to tell if your underpowered amplifier is any more or less linear, because *ARRL Radio Designer*'s active devices are crunchproof and distortion-free by definition.

Nonlinear Analysis

Professional-grade nonlinear simulators, including Compact's *Microwave SCOPE* and *Microwave Harmonica*, can do most types of nonlinear AF and RF simulation extremely well—for prices beginning at roughly the cost of a new car. *PSPICE*, largely intended for dc, digital and audio design, and available as shareware from many computer bulletin boards, can usefully simulate some nonlinear effects—if you don't need accurate results at ham-band and VHF/UHF frequencies. So where does that leave linear, small-signal *ARRL Radio Designer*?

In good company, as it turns out. Compact Software gurus tell us that perhaps 90% of all microwave and RF professional circuit simulators sold are linear. Even though we stand on the threshold of the DSP age, our radios still consist largely of linear circuits. Because linear AF and RF design techniques will remain important in professional and Amateur Radio for the foreseeable future, an affordable RF-circuit simulator stands to be of significant use to hams interested in getting current and *keeping* current with modern RF techniques. That's why we're pleased and excited to bring *ARRL Radio Designer* to you.—*WJ1Z*

Those New *QST* Propagation Charts

With LUF curves now shown in the monthly graphs, you can intelligently choose the bands and the times of day for the best probable propagation conditions to the DX areas of your choice.

By Jerry Hall, K1TD
Associate Technical Editor (Retired)
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Wethersfield, CT 06109-3309

Beginning with this issue, the propagation-prediction graphs appearing in the How's DX? column convey some different information than has appeared in the past. You should find the new information very helpful when you're choosing the bands and the times to use during a contest, for going after that DX country you've been trying to snag, for making a schedule, or for just making a general contact with a particular distant area.

As anyone who frequently uses the HF bands knows, propagation conditions change with the time of day and often from one day to the next. Propagation also changes from month to month and with the so-called 11-year solar cycle. Soon we'll see the close of Solar Cycle 22, which began in September 1986; present indications are that it will bottom out and end late in 1995 or sometime in 1996. With the decline in solar activity, band openings on 10, 12 and 15 meters will become fewer and farther between. During the solar minimum, bands at the higher end of the HF spectrum may never come alive for weeks on end. In these coming times, information from *QST*'s How's DX? charts will help answer the recurring question, "Is the band dead, or is there just no one out there transmitting?"

Propagation Predictions

Short-term changes in propagation cannot be accurately predicted far in advance, and that's where listening to WWV or WWVH will bring you up to date on current conditions. How to use that information is discussed in more detail later in this article. Long-term changes—variances with month and with the level of solar activity—are taken into account by each of many computer programs that are available. IONCAP, one of the most sophisticated programs, is used at ARRL Headquarters to prepare the How's DX? charts.[1] All IONCAP results are based on *probabilities*.

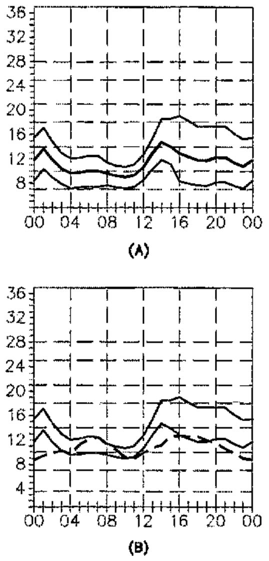

Figure 1—Curves showing propagation probabilities from Mid-USA to Central Asia during the period from mid-October to mid-November 1994. The vertical axis indicates frequency in MHz, and the horizontal axis, Coordinated Universal Time (UTC). The curves at **A** (HPF/median MUF/FOT) are those appearing in previous issues of *QST*, while those at **B** (HPF/median MUF/LUF) appear beginning with this issue. See text for details on these and other changes.

Propagation charts have appeared in *QST* since January 1977.[2] See Figure 1A, depicting the style of charts that were adopted in September 1977. This chart shows propagation estimations for the path from Mid-USA (Kansas City) to Central Asia (New Delhi, India). The uppermost curve shows the highest possible frequency or HPF. Based on probabilities, the ionosphere will support these frequencies on 10% of the days of the propagation period, or about 3 days of the mid-October to mid-November prediction period. The middle curve shows the median maximum usable frequency or median MUF. The ionosphere will likely support these frequencies on 50% of the days, about 15 days of the split-month period. (A word of caution here: Don't be misled when the word median is dropped in referring to this curve, as is often done. The median or 50% MUF is not the same as the true MUF, which is the maximum frequency that will be propagated via the ionosphere between the two end-points of a given path at a given time.)

The lower curve in Figure 1A shows the frequency of optimum transmission or FOT. Propagation at these frequencies will probably be supported by the ionosphere for 90% of the days, about 27 days of the period. By implication, you should almost be guaranteed a QSO if you attempt con-

Table 1
Path Terminal Points

General Location	*Actual Location**
West Coast	San Francisco, CA
Mid-USA	Kansas City, MO
East Coast	Washington, DC
South America	Asuncion, Paraguay
Central Asia	New Delhi, India
Southern Africa	Lusaka, Zambia
Western Europe	London, England
Eastern Europe	Kiev, Ukraine
Japan	Tokyo
Australia	Sydney
South Pacific	Pago Pago, Samoa
Alaska	Anchorage
Hawaii	Honolulu
Caribbean	San Juan, Puerto Rico

*Latitudes and longitudes used for the actual locations are those appearing in Table 4-3 of *The ARRL Operating Manual*, 4th edition.

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[1]Notes appear on page 30.

tacts at times and frequencies indicated by the FOT curve, right? This is the "optimum" frequency, so how can you go wrong? Let's follow up on this idea while referring to Figure 1A; it looks as if a great opportunity to work Central Asia from Mid-USA is on 40 meters between 0400 and 1000 UTC (10 PM to 4 AM CST). The FOT curve is relatively flat during this time period, and it hugs the broken line representing the 40-meter amateur band all the while.

Lowest Usable Frequency, the LUF

But have we overlooked something? Is 40 meters from 0400 to 1000 really the best choice of band and time? As we'll see shortly, NO! (Actually, it's about the worst choice!) Why is this? To comprehend the reason, we need to understand that the three curves of Figure 1A tell us *only* that, based on probabilities, the various frequencies will be supported by ionospheric propagation when indicated. But those curves tell us nothing of signal strengths. To get another piece of the best-propagation puzzle, we also need to consider the path losses between the transmitter and the receiver. A big contributor to losses is absorption in the ionosphere, particularly the D region. Some other losses are from dispersion of the transmitted energy in space and signal scattering at intermediate reflection and/or refraction points. When all these losses are added up, they can knock the signal from a 1500-W transmitter down to an S-1 level and even lower. So while the signal may be propagated via the ionosphere, it might be too weak to be heard at the receiving end of the path.

This means that at times, especially on longer paths, signals at the FOT or "optimum" frequency may be unusable because of path losses. The lowest usable frequency or LUF indicates the frequency below which signals will be too weak to be usable. The LUF may be calculated by taking many variable factors into account, such as all the path losses, the transmitter power, the transmitting and receiving antenna gains, the noise level at the receiver site, and even the bandwidth of the signal. To illustrate how some of these variables affect the LUF, consider that a weak signal from a 1.5-kW transmitter can get through the noise at times when the signal from a 1.5-W rig would never make it. Similarly, a CW signal will often get through the noise when an SSB signal will not.[3]

The New *QST* Charts

IONCAP and a few other computer programs support LUF calculations by taking all the path losses and many other variable factors into account. Figure 1B shows the new style for the *QST* charts. You'll see right away that the FOT curve of Figure 1A has been replaced by a heavy broken line. This is the LUF curve, shown as heavy and broken for distinction. The chart of Figure 1B is actually one of the 30 How's DX? charts appearing in this issue, but shown at a larger size. At first the curves that cross over each other in Figure 1B may appear confusing, but the explanation that follows should eliminate any confusion.

Table 2

Parameters Used for IONCAP Calculations

Transmitter power: 1500 W

Antennas at each end of the path:

Data is read from an external binary antenna file created especially for preparing these *QST* propagation charts. The data is based on dipole and Yagi antenna elevation patterns, modified for a constant gain at elevation angles above the peak response of the lowest lobe in the patterns. (Of course the peak angle changes with frequency and with antenna height.) The constant-gain characteristic at the higher angles avoids "holes" in the data that occur because of nulls in the patterns of real antennas. From the standpoint of gain and antenna height, this file emulates 8 separate monoband antennas.

The basic antennas for each amateur band are:
80, 40 and 30 m: Horizontal half-wave dipoles, 100 ft high
20 and 17 m: 3-element Yagis, free-space gain 5.5 dBd, 100 ft high
15, 12 and 10 m: 4-element Yagis, free-space gain 7 dBd, 60 ft high

Ground characteristics at each end of path: "Average" ground
Dielectric constant: 13
Conductivity: 5 millisiemens per meter
Minimum radiation angle: 1.0°
Manmade noise level in a 1-Hz bandwidth at 3 MHz at receiver site: −148 dBW (typical for rural areas)
Required reliability: 50%, half the days of the month
Required SNR: 30 dB, for CW bandwidth (10 log 100 + 10 = 30). For SSB the required SNR would be 10 log 2100 + 10 = 43 dB, 13 dB higher.

To use the new charts effectively, it is important to keep one thing in mind—the old adage that frequencies slightly below the MUF are always the best to use. That's because the signals are reflected back at shallow angles from the ionosphere, giving them a longer skip distance, and because they suffer the least absorption, yielding stronger signals. Depending on the day-to-day propagation conditions, the actual MUF may be near the HPF curve of the chart, it may be near the median-MUF curve of the chart, or it may be below the chart's median-MUF curve.

If you go lower in frequency from the actual MUF, the path losses increase, and with this the received signals will become weaker (all else being held equal). The lower you go in frequency, the weaker the signals become, until eventually you reach the LUF. This means that there is a frequency window for making contacts on a given path at a given time. That window includes all frequencies between the MUF and the LUF.

As propagation changes during the day the frequency window changes, and often even closes for a time on the longer paths. This happens when the MUF equals or goes below the LUF, indicated by a crossover of the curves. When the frequency window is closed, it will be difficult or impossible to make contacts on any frequency. In Figure 1B the median MUF goes below the LUF during two time periods, 0300 to 1100 and 1700 to 1900 UTC. So now, what about 40 meters between 0400 and 1000, as we originally selected from Figure 1A? No good at all, as we can now realize from Figure 1B! The LUF is higher than the median MUF for the entire period. It'd be better to get some sleep, and try at another time.

As the MUF depends on day-to-day propagation conditions, so does the LUF. As solar activity increases, so do path losses from absorption, and the lowest usable frequency becomes higher. In other words, on exceptionally good days when propagation conditions support the HPF, the LUF will also rise somewhat. So the frequency windows can shift up and down as propagation conditions change from one day to the next. But the MUF and the LUF do not always "track" each other. Another factor that greatly affects the LUF is the earth's geomagnetic activity, indicated by the K and A indices broadcast by WWV and WWVH. As these values increase, the LUF also increases. The *QST* charts assume the earth's geomagnetic activity is low.

QSO Windows

Refer again to Figure 1B. With the preceding information, we now see that on an average propagation day there will be two frequency windows during the forecast month for the Mid-USA to Central Asia path, from 1100 to 1600 and from 1900 to 0300 UTC. With the new charts, finding the best times and amateur bands for making schedules or for seeking DX contacts is simply a matter of selecting a big frequency window from the chart (a large separation

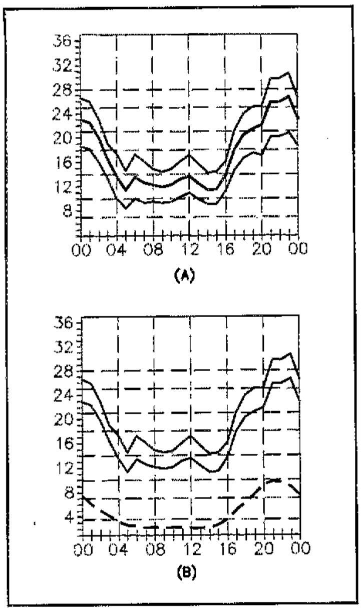

Figure 2—Curves showing propagation probabilities from Hawaii to the West Coast for mid-October to mid-November. The horizontal axis indicates Coordinated Universal Time (UTC) and the vertical axis frequency in MHz. The curves at **A** show HPF/median MUF/FOT, while **B** shows the new chart style with HPF/median MUF/LUF curves. Note in **B** that the frequency window is open around the clock. See text for determining the FOT from the new chart style.

between the median-MUF and the LUF curves) and choosing the band nearest the median MUF—a QSO window, if you will. We see in Figure 1B that on average there will be a brief 20-meter band opening around 1400 UTC. (On exceptionally good days this may develop into a 17-meter band opening from 1400 to 1800.) We also see that the chances for a 30-meter band opening are good from 2200 to 0200 UTC. The best propagation during the 24-hour period is likely to be during the brief 20-meter QSO window at 1400, as the LUF is below 12 MHz.

If you're looking for a longer QSO window, then try 30 meters from 2300 to 2400; the median MUF is not far above 10 MHz and the predicted LUF is below 9 MHz. If you are making a schedule, it'd be wise to have alternative bands to allow for short-term propagation changes. Twenty meters between 2300 and 2400 would be a good second choice, as the HPF is above the 20-meter frequencies.

Among all the charts in the How's DX? column you'll see several paths where the frequency window is open all day long. The Hawaii to West Coast path is one example, shown in Figure 2B. Choosing the best band for use at a particular time of day is simple; just take the one nearest the median-MUF curve, keeping in mind that day-to-day changes may have some effect.

FOT Values from the New Charts

Just because the FOT curves will no longer be appearing in *QST*, you should not assume the FOT information has no value. Indeed, the FOT will almost always provide consistent "armchair copy" on short paths, and on any path when the FOT is significantly above the LUF. The FOT information is very helpful in point-to-point communication where 90% or higher reliability is required. Say that you live on the West Coast (San Francisco) and want to maintain continuous reliable contact with Hawaii (Honolulu) for an extended period of time. Figure 2 shows the probabilities for this path. Figure 2B indicates the frequency window is open all day long with the LUF significantly below the median MUF, so amateur bands near the FOT would be a wise choice.

Even though the FOT data is not included in the new chart, you can still obtain FOT values to a good approximation by taking 80% of the median MUF.[4] For example, note the median MUF in Figure 2B at 0600 UTC. The value there is 13.0 MHz. Taking 80%, FOT = 13.0 × 0.8 = 10.4 MHz. For comparison of this result, see the FOT value plotted in Figure 2A. At 0600, IONCAP predicts the FOT will be 10.0 MHz.

Let's look at another time, 1600 UTC. The median MUF is 14.0 MHz. The approximation is FOT = 14.0 × 0.8 = 11.2 MHz, whereas the value from IONCAP is about 11.5 MHz. Although there may be some slight difference between this approximation and more precise calculations, the appropriate amateur band should be easy to discern from the approximations. For this path, Hawaii to West Coast, the FOT results indicate the 30-meter band should provide communications for at least 90% or 27 days of the prediction period from about 0400 to 1600 UTC.

Specific Path Terminal Points

The 30 charts in the How's DX? column cover as many paths as possible within a reasonable amount of page space for all the readers of *QST*. For a particular geographic area, the number of those to use is reduced. If you're in the western, central or eastern part of the US you'll find charts to South America, Central Asia, Southern Africa, Western Europe, Eastern Europe, Japan and Australia. There are two charts to South Pacific from the US, Mid-USA and East Coast, and one chart from the West Coast to the Caribbean. In addition, there are two charts for Alaska showing probabilities for the East Coast and Western Europe, and three charts for Hawaii to the West Coast, East Coast, and Western Europe. A chart for the East Coast to West Coast completes the set.

In earlier issues of *QST* a chart was included for the West Coast to the South Pacific. Beginning with this issue, that chart has been replaced by the popular path, Hawaii to the West Coast. If you are interested in data for the the West Coast to the South Pacific, you can linearly interpolate the sets of curves for Hawaii to the West Coast and the West Coast to Australia. The results will be quite accurate, as the bearings are essentially identical and Pago Pago (the South Pacific terminal point) is very close to being halfway between Hawaii and Australia.

Another change in the charts is the terminal point in Australia. Now the charts use Sydney, rather than the former Melbourne. This change better covers New Zealand, and also facilitates the interpolation for West Coast to South Pacific, mentioned in the previous paragraph. Another minor change is a name change only for the path from the Caribbean to the West Coast, formerly shown as Puerto Rico to the West Coast.

From these statements it becomes obvious that the labels for the charts give generalized geographic areas. The actual locations for all terminal points are listed in Table 1. If you are located some distance from an actual terminal point, you can make some correction based on latitude. The general rule is that the higher the latitude, the lower will be the median MUF.

IONCAP Data

The results from all IONCAP calculations pertain to a calendar month that is specified when the program is run. Because the *QST* prediction period straddles two calendar months, some extra steps are taken at ARRL Headquarters to provide information for the *QST* prediction period. Two sets of calculations are run—one for the first calendar month and one for the second. The means of the results (averages, in common parlance) from the two months are then used to prepare the charts.

The technical data that goes into IONCAP for the calculations are summarized in Table 2. The legal power limit (1500 W) for the calculations is used, not because everyone will be running this power, but because the LUF can be adjusted upward for lower levels. Similarly for the antenna gains; the LUF can be adjusted upward if you have antennas with lower gains than those used for the calculations, or adjusted downward if you have super antennas with greater gains.

Short-Term Propagation Changes

As mentioned previously, the actual MUF may be near the HPF curve of the chart, it may be near the median-MUF curve of the chart, or it may be below the chart's median-MUF curve. To get a report of current propagation conditions, listen to the WWV propagation broadcasts at 18 minutes past the hour or WWVH at 45 minutes past.[5] Daily observed 2800-MHz solar flux values are given in the broadcasts. Compare this number with that in the caption for the How's DX? curves. If the flux number for the day is higher than that used to prepare the charts, the frequency windows will generally shift to higher frequencies. The higher frequency bands will usually open earlier and close later than predicted. Conversely, if the daily flux value is lower than that used for the charts, the frequency windows will be lower; the higher frequencies will open later and close earlier than indicated. Higher frequency bands showing only brief openings in the charts may not open at all.

Also pay attention to the A and K indices, indicators of the Earth's geomagnetic activity as measured at Boulder, Colorado. The A index is calculated from the previous day's K index values,[6] so it tells you mainly how yesterday was. As the A index rises, so does the LUF. The K index is updated every 3 hours in the WWV/WWVH broadcasts. The absolute values are meaningful, with values of 3 or 4 and above indicating unsettled geomagnetic conditions, but the trend of the 3-hour periods also conveys useful information. Rising values mean the LUF will be going higher still, while falling values mean the LUF will be dropping.

Another way to learn about current propagation conditions is to spend a few moments tuning the bands and comparing what you hear against appropriate charts in *QST*. Observations for paths to not-too-distant locations from your area can be applied to the really long-haul paths. Listen for areas where there is usually a lot of amateur activity.

If a chart says you should have a good QSO window to an area less than roughly 5000 miles away and signals are booming in from that area, this means conditions at the time are close to those predicted in the charts. On the other hand, if you hear only a few weak signals from there, chances are that the LUF is higher than predicted. If you hear no signals at all from there, then either the LUF is significantly higher or else the MUF is lower than predicted. A way to help determine which is to listen on the next-higher frequency band. If you hear any signals from the same area on this band, then the LUF is higher; if you again hear nothing from the area, the MUF could be lower, but also the LUF could be considerably higher because of geomagnetic activity. This all assumes the *QST* chart indicates a good QSO window. If you hear very little or no activity on any bands at times when the charts indicate good QSO windows, then the LUF is considerably higher than predicted.

In Summary

With the LUF curves present in the monthly How's DX? charts, it's easy to see the very best times and bands for making schedules or for seeking DX contacts—just pick the time when the widest gap or frequency window exists between the MUF and the LUF curves (with the LUF curve on the bottom!), and choose the band nearest the MUF.

It's also easy to see what paths will be difficult, no matter what the time of day or what band is used. Knowing when none of the amateur bands may be usable becomes important on the longer paths.

Is the band dead, or is there simply no activity? With *QST*'s new monthly propagation charts and a knowledge of current propagation conditions, now you can determine the answer to that question.

Notes

[1] IONCAP was written by the Institute for Telecommunication Sciences and its predecessors in the US Department of Commerce. For more information see J. Hall, "Propagation Predictions and Personal Computers," Technical Correspondence, *QST*, Dec 1990, pp 58-59.

[2] See D. Sumner, "Chart Your Way to Better DX," *QST*, Jan 1977, pp 58-60. (HPF curves were not included during the first 8 months of chart appearance in *QST*.)

[3] For additional information on the LUF and some practical examples, see J. Hall, "Propagation Predictions for HF—A New Look," *QST*, Feb 1992, pp 48-50.

[4] The rule for earlier manual methods of predicting propagation was to take 85% of the predicted MUF to obtain the FOT, but taking 80% of the median MUF from the *QST* charts produces results closer to those of IONCAP.

[5] For additional information on interpreting the data from the broadcasts, see R. Healy, "Propagation Broadcasts and Forecasts Demystified," *QST*, Nov 1991, pp 20-24.

[6] G. Jacobs and T. J. Cohen, *The Shortwave Propagation Handbook*, 2nd ed. (Hicksville, NY: *CQ* Publishing, Inc, 1982), p 110.

New Products

GETTING STARTED ON VHF

CQ Communications, 76 N Broadway, Hicksville, NY 11801-9962, tel 800-853-9797; VHS video, 52 minutes, 5 seconds; $19.95 plus $4 s/h.

Reviewed By Steve Ford, WB8IMY
ARRL Assistant Technical Editor

If there was ever a videotape ideally suited to new codeless Technicians, *Getting Started on VHF* may be the top candidate. As soon as the tape begins, you recognize that this show was created with the Technician audience in mind. It should come as no surprise that the majority of its 52 minutes is devoted to FM. Surveys show that most new hams (and codeless Technicians in particular) get their feet wet on FM these days.

Getting Started on VHF covers FM operating in complete detail. The video explains everything from transceiver programming to repeaters. Experienced FM operators will be pleased to know that *Getting Started on VHF* takes great pains to prevent new hams from falling into the dreaded "repeater speak" pattern. The writers specifically caution against using terms like "handle," "personal," "destinated" and so on. If only the FCC could make this video required viewing for all new amateurs!

As you watch the video, you see the dimensions of FM operating beyond repeater chit-chat. Contesting and public service are shown as attractive, exciting activities. The video also mentions an activity that's rarely covered in the ham press: FM DXing. (It isn't as common as VHF DX on SSB and CW, but it does happen.)

As with its other tapes, CQ Communications employs superb production techniques. This is a slick video! Animation is used to good effect, and the narration is sincere and enthusiastic. Much of the credit for effective narration goes to Jeff Ronner, WB2AEQ. His camera presence is smooth and highly professional. You'd expect this, considering his background as an announcer for WROW radio in Albany, New York. Now that he owns his own production company, Jeff's voice is heard on radio and TV commercials throughout the Northeast.

Getting Started on VHF spends several minutes exploring other popular aspects of VHF and UHF operating. It devotes a substantial amount of time to packet, for example. (They even mention my book, *Your Packet Companion*. Bless them!) I was gratified to see a discussion of the so-called weak-signal modes such as moonbounce, satellites and terrestrial SSB and CW.

The NCDXF/IARU International Beacon Network—*Part 1*

This worldwide beacon network, in operation for almost 15 years, is now being expanded to a multiband system.

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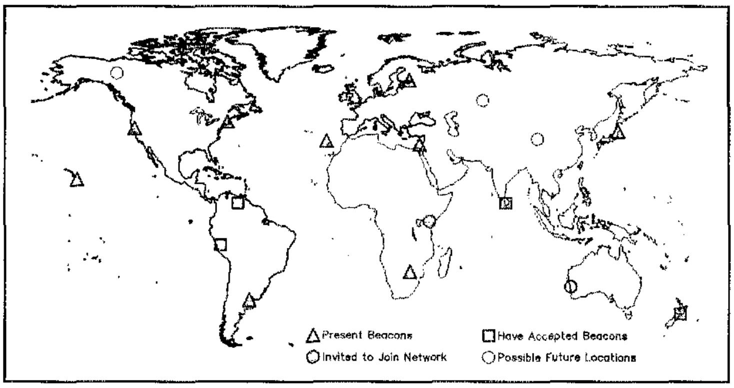
The present 14.1-MHz beacon network is one of those neat achievements that comes about when the vision, technology, and cooperative energies of many people combine to create something unique and useful. The worldwide network consists of nine frequency-sharing CW beacons that have been in operation continuously for almost 15 years. The network is there for every listener to use as a do-it-yourself propagation tool. Each beacon transmits a one-minute message every 10 minutes, 24 hours a day.

Over the next year or so, the network will be modernized and made even more useful. New beacons will be added to the nine that are now active, and all of them will be able to operate on up to five bands. The transmissions will be shortened to 10 seconds, so that the listener can monitor 18 beacons in three minutes. This two-part article explains how the current network evolved and how you can expect to use the modernized network.

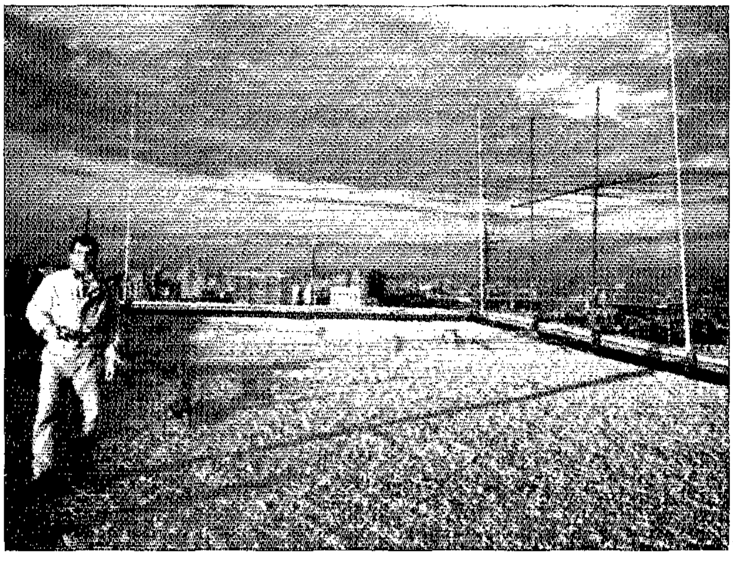
Phase I, the First Beacon

The 14.1-MHz beacon network was the creation of the Northern California DX Foundation (NCDXF).[1] The first beacon was designed and built from scratch by Jim Ouimet, K6OJO, in 1979, based on a suggestion by O. G. "Mike" Villard, Jr, W6QYT. You may remember Mike as the person who introduced SSB on the amateur bands in the late 1940s. This "Phase I" beacon was duly licensed by the FCC for operation on 14.1 MHz with the call sign WB6ZNL/B (with the /B indicating a beacon transmitter). The transmitter was heavy and it came to be known among the workers as the "guy-wire anchor," but it did its job as planned, transmitting for 75 seconds every 15 minutes for over a year. It featured Mike's idea of decreasing the power in 10-dB steps by transmitting a long dash at each of four power levels: 100, 10, 1 and 0.1 W.

The power stepping was a completely new idea for Amateur Radio beacon technology. It is much more useful to know what power level can be heard than to merely know the signal strength of the received signal. The knowledge that a certain power level can be



The NCDXF/IARU International Beacon Network, showing existing and future beacon locations.


JOHN G TROSTER W6ISQ

Dave Rosen, K2GM, chief operator at the 4U1UN/B beacon, on the roof of the United Nations building. The 14.1-MHz beacon antenna is a dipole at the far end of the roof.

Table 1
Present 14.1-MHz Beacon Network Transmitting Sequence

Time	*Station*	*Location*
00:00	4U1UN/B	United Nations, New York
00:01	W6WX/B	Stanford University, California
00:02	KH6O/B	Kaneohe, Hawaii
00:03	JA2IGY	JARL, Tokyo, Japan
00:04	4X6TU	Tel Aviv University, Israel
00:05	OH2B	Helsinki Technical University, Finland
00:06	CT3B	AARM, Madeira Island
00:07	ZS6DN/B	Transvaal, South Africa
00:08	LU4AA	RCA, Buenos Aires, Argentina

The same sequence repeats every 10 minutes. W6WX/B also transmits for 10 seconds every two minutes at 21.150 MHz followed immediately on 28.200 MHz; the call sign and one dash at each of four power levels are transmitted.

14.100 MHz Guarded Frequency

Almost from the beginning of the 14.1-MHz beacon network, the IARU drew up a band plan that suggested a 1-kHz "guarded" beacon frequency at 14.1 MHz. This guarded frequency was reaffirmed at the Region 2 IARU meeting in Curacao in 1992 and also has been adopted by IARU Regions 1 and 3. Recently, however, digital stations have swamped this frequency. It would be quite helpful to their fellow amateurs, and a courteous thing to do, for the operators of those digital stations to move either up and down in frequency and to avoid transmitting on exactly 14.1 MHz.

Table 2
Transmissions by Each Beacon in the Present Network

Power Level (W)	*CW Message*
100	QST de (call sign)
100	. ______________ (9-second dash)
10	. . ______________ (9-second dash)
1	. . . ______________ (9-second dash)
0.1	. . . . ______________ (9-second dash)
100	$\overline{\text{SK}}$ (call sign)

Total transmission time: 57 seconds
Speed: 22 wpm

heard compares the signal to the background noise. Sometimes an S3 signal is perfectly usable; sometimes it is totally useless. The power levels give the listener a better feeling for the quality of the propagation. If you heard only one power level yesterday but hear all four today, then the band is in better condition today. Anyway, it's more fun to tell a friend that you heard the South African beacon at 0.1 W than to report that a 100-W signal was S5.

Phase II, Worldwide Expansion

The horrible prospect of reproducing eight or more of these beacon behemoths for a worldwide network stirred the creative juices of Dave Leeson, W6QHS. Dave designed a beacon that consists of a small controller box to work in conjunction with the Kenwood TS-120S transceiver. The controller used the Intel 8748 computer-on-a-chip to adjust the power output of the transceiver directly via the transceiver's ALC input. This chip combines an eight-bit microprocessor with a 1024-byte EPROM. Jack Curtis, K6KU, wrote the assembly language program for the microprocessor firmware. Jack had used a similar chip for his popular Curtis Keyers.

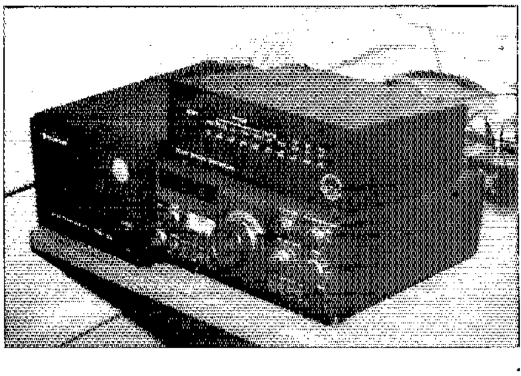
The late Cam Pierce, K6RU, engineered this prototype beacon into production with the help of Merle Parten, K6DC, and between 1982 and 1985 nine beacons were built and distributed worldwide.[2,3] During this expansion period, the call WB6ZNL/B was changed to W6WX/B.

The time slots during which the various Phase II beacons transmit on 14.1 MHz is given in Table 1, and the format for each transmission is shown in Table 2. Listeners who are not able to copy code at 22 wpm can figure out which beacons they are listening to according to the time that they hear each beacon. Table 1 provides the necessary information.

International Amateur Radio Union

In 1984, Alberto Shaio, HK3DEU, then Secretary of Region 2 of the International Amateur Radio Union (IARU), proposed that the IARU beacon program follow the general NCDXF frequency-sharing plan used by the 14.1-MHz network. Later, that proposal became the basis for NCDXF and IARU cooperation, and in recent years the two groups began planning together to expand the network and to develop a prototype multiband beacon. NCDXF provided the engineering, and IARU offered the international associations to help obtain locations for an expanded network and to disseminate beacon information worldwide to all 140 IARU member societies. Much of the funding will also be provided through the IARU.

After many years of operation it is apparent that the Phase II beacon network should be expanded to cover more of the world. The format of the transmissions should be changed to reduce the time it takes to listen for all beacons. And the beacon network should be expanded to cover more bands.


DAVID V. LASSEN, ZS6DN

The 14.1-MHz beacon at ZS6DN/B, with the controller sitting on top of the transceiver.

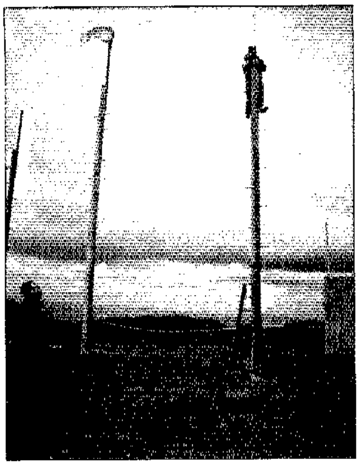
Expansion

The Phase II beacon network does not provide adequate worldwide coverage. With IARU assistance, we contacted five national societies and asked them to accept beacons and join the network. The societies that accepted are Radio Club Peruano, Radio Club Venezolano, New Zealand Association of Radio Transmitters, and Radio Society of Sri Lanka. The Wireless Institute of Australia and the Radio Society of Kenya have been invited, but they have not replied at the time this is being written. We hope to find one or more locations in Central Asia and to add other beacons later.

Transmission Format

Many people find the 10-minute cycle of the beacons too slow. As more beacons are added, the cycle would get even longer if we maintain the one-minute format for each beacon transmission.

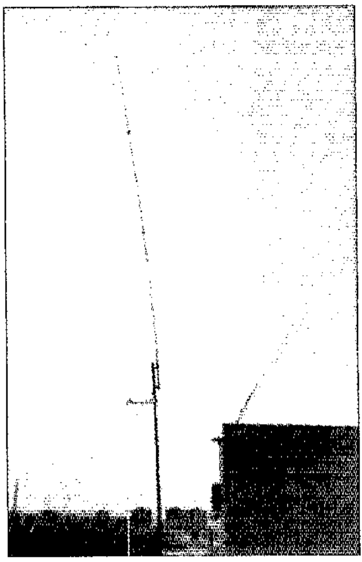
We experimented with shortening the transmissions. We assumed that each beacon would send its call followed by four equal-length long dashes, one at each power level. We recorded simulations of a network of beacons with 15 seconds per beacon, then with 10 seconds, then with 7.5 seconds. We played these recordings for the NCDXF Board and they recommended the 10-second format. The same recording format was


JOHN G TROSTER W6ISQ

Lance Ginner, K6GSJ, secures the vertical antenna for the W6WX/B beacon, as Bob Fabry, N6EK, watches, atop Mt Umunum, overlooking San Jose, California.

Kenwood TS-140S Transceiver Donation

Bob Ferrero, W6RJ, of Ham Radio Outlet, and former president of the NCDXF, contacted Kenwood Communications Corporation about a possible discount on the Kenwood TS-140S transceivers that are the heart of the beacon system. Paul Middleton of Kenwood responded with the donation of 16 TS-140S transceivers to the network. We are deeply grateful to Kenwood for their generous endorsement of the beacon project. This magnificent gift was presented by Kenwood in memory of Jim Rafferty, N6RJ, the former Vice President of Ham Radio Outlet. A small plaque will be attached to each transceiver with this inscription: "This Kenwood TS-140S transceiver was donated to the NCDXF/IARU International Beacon Project by Kenwood Communications Corporation and Ham Radio Outlet in memory of Jim Rafferty, N6RJ."



The antenna for the 4X6TU/B beacon, mounted on the physics and astronomy building of Tel Aviv University.

played for the delegates of IARU Region 2 at a meeting in Curacao in September 1992. The delegates concurred that 10 seconds sounded about right. This timing allows 18 beacons to transmit in sequence around the world in three minutes.

Multiple Bands

Expanding the beacon network to cover additional bands will provide valuable additional propagation information. The Phase III network will transmit on the 14, 18, 21, 24 and 28-MHz bands. (The 10-MHz band is not included because the band is still shared with other services, but it could be added later.) In addition to the listener being able to detect band openings on an individual band, he will also be able to quickly check all five bands to see which band has the best propagation to a particular part of the world.

To select frequencies for five-band operation, Bob Knowles, ZL1BAD, International Coordinator of the IARU Monitoring Service was asked to study the bands from 14 to 28 MHz. His report, based on the work of the IARU worldwide volunteer monitoring system, was instrumental in developing the tentative primary frequencies: 14.100, 18.110, 21.150, 24.930, and 28.200 MHz. Alternate frequencies at the high and low ends of each band also are being considered, but at the time of ZL1BAD's report there was interference from commercial stations outside the high end of some bands that could be copied inside the amateur bands. We are continuing to study the situation and will be able to re-program these frequencies if necessary by issuing a firmware upgrade.

Phase III, the W6WX/B Prototype

A new control unit has been built that can key a Kenwood TS-140S on five bands. This unit is now in operation at W6WX/B, but is restricted to operation on 14.100, 21.150, and 28.200 MHz because those are the frequencies for which it is licensed by the Federal Communications Commission. The technical aspects of this new beacon design will be described in Part 2 of this article.

New Format

W6WX/B transmits its regular one-minute message in its turn on the 14.1-MHz network at 0001Z, and every 10 minutes thereafter, just as it has for 15 years. When it has completed its 14.1-MHz one-minute message, it switches to 21.150 MHz and transmits for 10 seconds in the new format, "W6WX/B, dah-dah-dah-dah." It then immediately switches to 28.200 MHz and sends the new-format message. The 21 and 28-MHz message is repeated every two minutes. The other seven beacons will continue to transmit their one-minute message on 14.1 MHz only until they are replaced by multi-band equipment.

When the new beacons are in place, each will be able to transmit its call sign and four dashes on each of five bands. Ultimately, we hope all beacons will be licensed by their governments for five-band operation.

Financing

The three IARU Regional Executive Committees were solicited for funds to build and distribute the beacons in their respective Regions. Region 2 (North and South America) responded with a pledge to fund one beacon, at an estimated cost of $2500 per beacon, plus $1000 for continuing support. Region 1 (Europe, CIS and Africa) did the same. The ARRL Foundation generously contributed $5000. We hope that other major national amateur organizations will also become sponsors. John Downing, K6YRU, of the Downing Foundation, which has funded NCDXF with several grants, provided funds for prototype Phase III beacon construction.

Conclusion

The present 14.1-MHz beacon network is for everyone, whether you are a DXer looking for general band-opening information, or a contester looking for spot-opening information, or perhaps a high school or college student working on your science project, or a laboratory researcher, or SWL, or just a rag chewer who would like to find out what's new. Get the 14.1-MHz habit now. Flip in your CW filter and listen along.

DXpeditions even find the beacons useful. Bill Schmieder, KK6EK, in his recently published book *3Y0PI, Peter I Island Antarctica* remarked that conditions at one point were very poor. So just to make sure conditions were as bad as they sounded, he "listened on 14.1 for the beacons, but...heard nothing."[4]

In part 2 of this article, we will discuss the implementation of the beacons for Phase III, including the use of a GPS satellite receiver to provide timing.

Notes

[1] J. G. Troster, W6ISQ; O.G. Villard Jr, W6QYT; J. K. Ouimet, K6OPO; C. G. Pierce, K6RU; "The WB6ZNL Beacon," *QST*, Jan 1980, p 57.

[2] J. G. Troster, W6ISQ, and C. G. Pierce, K6RU, "World-Wide Beacon Net: The Possibilities Abound," *QST*, Jun 1983, p 27.

[3] J. S. Stover, W5AE, "20-Meter Beacons Revisited," *QST*, Dec 1988, p 60.

[4] R. Schmieder, KK6EK, *3Y0PI, Peter Island Antarctica* (Walnut Creek CA: Cordell Expeditions, 1994). QST

New Books

MORE ABOUT CUBICAL QUADS

By George W. McCarthy, W6SUN

Published by Worldradio Books, PO Box 189490, Sacramento, CA 95818; tel 916-457-3655. First Edition, 1994. Paperback, 60 pp, 8½ ×11 inches, B&W photographs and diagrams. Retail price $10, $2 s/h.

Reviewed By James D. Cain, K1TN
QST *Senior Editor*

Owning and operating a cubical quad is a lot like being married. If only I had realized this 25 years ago.

Acquiring a fancy, sophisticated HF cubical quad takes a lot of effort, and maintaining it requires almost constant time and energy. Ham visitors will lust (at least in their hearts) for your quad. And, as the author makes clear, even after every effort to be good to it, your quad may turn on you on a whim (or a breeze).

George McCarthy has obviously conducted a long love affair with the cubical quad—with many of them, in fact. "The Quad is a big, beautiful box kite," he opines. He remembers with affection each of them, for their quirks, and for what he learned from them. And he learned plenty.

This modest (60-page) book is what I'd call an "inspirational" work; if you've had quads in the past, it reminds you about them, and if you haven't, it may make you want to try one. Quads are that way.

George McCarthy, who's 73 (nice age), has been a ham since 1939 and says he's played with quads for 25 years. That places him in 1969 when, it seems to me, quads were more popular. Or maybe it's that I had one in the mid-'60s (actually, I had two).

George reveals, in the friendly style that makes this book fun to read, that he's not an engineer, but has educated himself about quads by chatting up others over the years. "Ask two quad owners and you'll get three opinions." He's at his best in his introductory chapters on deciding whether to try a quad, and then in discussions of the myths and legends surrounding the cubical quad design.

He does a credible job (so sez our resident antenna guru) of dispelling some of the myths. Two examples:

The quad is kind of a big sloppy antenna...it will work well if you just get some squares up in the air. "Absolutely not true," says George.

Many of us can vouch for this: sling a rope between a couple of trees, hang loops for reflector, driven element and director, and find the resulting "3-element quad" is no better than a reference dipole.

Quads really do need to be tuned, and George tells you how. On the other hand, although he does talk about using single full-wave loops and fixed quad loop arrays, there's not much info here (it's available elsewhere, though).

George agrees with the belief that quads are less susceptible than Yagis to rain static (he doesn't mention snow static, but then, he lives in the desert). He says that raindrops are electrically charged, and doesn't it make sense that they're more likely to hit a one-inch-diameter hunk of aluminum (as in a Yagi) than to hit a 1/16-inch-diameter wire? It does make sense.

The exploded myth I like best: Quads can be made to survive as well as any Yagi, to which OM McCarthy responds, "And that yellow stuff in the plastic tub tastes like real butter."

Other chapters do a pretty thorough job of discussing choosing a design, tuning, matching, and so on. A list of current suppliers of quad parts will allow the budding quadophile to compare items and prices.

I have no idea what the worldwide ratio of Yagi-to-quad antennas might be, but I do know quads are for the lunatic fringe. You'll get this feeling from most of what George says. One thing I find troubling is that every quad pictured in his book is on a crankup tower. This is virtually a necessity, because quads are nearly impossible to assemble atop a guyed tower, ditto for repair, and their elements must be accessible for the all-critical adjustment and tuning.

This is fine if you're willing to spend the extra money (a lot extra) for a crankup tower, which I think are much more common on George's Coast than elsewhere.

This would be a fine introductory book for a newcomer who wants to assemble a quad for himself; doing so can be a lot of fun and a learning experience, and George McCarthy imparts that throughout his book. Unfortunately, such newcomers may be a dying breed. I hope I'm wrong.

Strays

QRO MARITIME MOBILE

◊ One of the major events in the Pacific during WW II was the landing of US forces to free the Philippines. One part of this effort that involved a number of hams was the installation and operation of one of the most powerful shipborne shortwave broadcasting stations ever built.

News correspondents had great difficulty getting their dispatches out of the Southwest Pacific Theater of Operations, often having to fly their stories to Australia to reach the nearest radio facilities. Finally, network officials in New York and war correspondents in the Pacific appealed to Gen Douglas MacArthur to provide them with a high-power, broadcast-quality, shipborne broadcasting facility that could transmit directly to the US. Gen MacArthur recruited a small group of Signal Corps officers and noncommissioned officers, most of whom were hams, to accomplish this unique task.

They were given the transport ship USS *Apache* and two 10-kW transmitters (one for the standard AM broadcast band and one for shortwave broadcast) to install on the ship, and control-room equipment, two 50-kW, 240-V three-phase power plants, and accessory equipment. The entire facility was assembled in Sydney, Australia, between July 1 and September 20, 1944, with final equipment installation and adjustments made as the *Apache* sailed to join the convoy of ships for the invasion of the Philippines.

After the invasion was launched, at 10 AM on October 20, 1944, the standing order for radio silence was lifted and soon thereafter, the *Apache* transmitted the first Philippine invasion bulletin. At 3 PM the same day, the ship's transmitters carried Gen MacArthur's famous "I have returned" speech via a VHF link from the *Apache's* remote truck on Leyte.

In addition to its role in handling news broadcasts back to San Francisco for radio feeds to the networks, the *Apache* made "Voice of Freedom" broadcasts on 7.795 MHz to friendly guerrilla and civilian forces in the Philippines, telling them of the invasion's progress and urging them to take up their own arms. Reports reached the ship of American and Allied prisoners of war receiving the broadcasts on concealed receivers. The *Apache* even got an indirect SWL report when Tokyo Rose said in one of her propaganda broadcasts, "Look out, *Apache*; we know where you are!"

The *Apache* story was told in the article, "Those Singing Masts," by 1st Lt Tony Borgia, W6EOU (now K6DR), in September 1945 *QST*. Several of the *Apache's* radio crew are known to be living: 1st Lt Al Pierce, ex-W9CHO, the operations officer, retired as manager of television operations for CBS in Chicago and now lives in Sarasota, Florida. Tony Borgia, the ship's transmitter officer, was a career Army officer until his retirement and now lives in Carmichael, California. 1st Lt Sanford Terry, ex-W3AGH/W4AGH, an engineering officer, later worked in TV broadcasting, finally retiring as vice president for engineering at WWBT-TV in Richmond, Virginia; he still lives in Richmond. The whereabouts of the other *Apache* radio crew members is unknown.—*Sanford Terry, ex-W3AGH/W4AGH*

QST congratulates...

◊ Charles "Chip" Townsend, WA4DCN, supervising engineer, Department of Management Services, State of Florida, who was honored as outstanding employee in DMS in a ceremony at the Governor's Mansion. He took his own equipment from Tallahassee, Florida, to Pensacola, FL and eliminated stray broadcast signals from the Florida Labor Department's telephone lines. The local telephone company and the maintenance company had not been able to solve the problem. —*Thanks, John Hills, KC4N, Assistant Director, ARRL Southeastern Division*

QST Product Reviews:
A Look Behind The Scenes

Learn how the ARRL Laboratory evaluates new products—and what all those numbers mean to you!

By Mike Gruber, WA1SVF
ARRL Laboratory Engineer

Today's *QST* Product Reviews are better and more comprehensive than ever. They're a tremendous service to any ham interested in new or used equipment. We're always working to improve the quality of our Product Reviews. Recent trends include more products, comparison reviews and increased Lab testing.

Let's take a look at some of the measurements you will see published in a typical product review for an HF transceiver, and discuss how you can use them for evaluating an equipment purchase. Not surprisingly, our tests are broken down into two categories: transmitter tests and receiver tests. We'll consider receivers first.

Receiver Tests

As many newcomers to ham radio have discovered, comparing the performance of one receiver to another is often difficult. The *features* of one receiver may outweigh another, even though its overall *performance* might not be as good. Here is where a little knowledge and some Product Review data can be invaluable.

Some of the more important receiver measurements you'll find in *QST* Product Reviews are sensitivity, blocking dynamic range and two-tone IMD (intermodulation distortion) dynamic range.

Receiver Sensitivity

Sensitivity measures a receiver's ability to hear weak signals. You'll find several methods used to express it. The mode under consideration often determines the best choice. Do not be confused by this—the basic concept is really the same for each. Put simply, receiver sensitivity is the level of input signal required to produce a given signal output. The only variables are the units used to measure the input signal, and the defined receiver output.

Sensitivity is typically expressed in µV, dBm and occasionally in dBµV. You'll see dBm used to describe the power level in dB relative to one milliwatt; dBµV describes the voltage level in dB compared to 1 microvolt.

Logarithmic power units like the dBm are not only very convenient to use when amplifying or attenuating signals, but they're also useful when making comparisons between radios.

One of the most common sensitivity measurements you'll find for CW and SSB receivers is the minimum discernible signal (MDS). MDS is the input level to the receiver that produces an output signal equal to the internally generated receiver noise. Hence, MDS is sometimes referred to as the receiver's *noise floor*. The typical MDS of a modern radio is −135 to −140 dBm with 500 Hz bandwidth.

Receiver sensitivity is also often expressed as 10 dB signal-plus-noise to noise (10 dB [S+N]/N) or 10 dB signal to noise ratio (10 dB S/N). The procedure and measurement are identical to MDS, except that the input signal is increased until the receiver output increases by 10 dB for 10 dB (S+N)/N and 9.5 dB for 10 dB S/N.

AM Sensitivity: AM receiver sensitivity is usually expressed as 10 dB (S+N)/N or 10 dB S/N. The basic procedure is similar to that used to measure MDS, except the signal generator is set for a 30% modulated, 1 kHz test signal. (The modulation in this case is keyed on and off and the signal level is adjusted for the desired increase in the audio output.)

FM Sensitivity: SINAD is the most common sensitivity measurement normally associated with FM receivers. It's an acronym for SIgnal plus Noise And Distortion and is a measure of signal quality:

$$\mathrm{SINAD} = \frac{\text{signal} + \text{noise} + \text{distortion}}{\text{noise} + \text{distortion}}$$

where the ratio is expressed in dB.

Let's take a closer look at SINAD if we consider distortion to be a part of the receiver noise. (Distortion, like noise, is something unwanted added to a desired signal by the receiving system.) If we also assume that the desired signal is much stronger than the noise, SINAD now closely approximates the signal to noise ratio:

$$\mathrm{SINAD} = \frac{\text{signal} + \text{noise} + \text{distortion}}{\text{noise} + \text{distortion}} \approx \frac{\text{signal}}{\text{noise}}$$

where the ratio is expressed in dB.

The most common SINAD spec for amateur receivers is 12 dB SINAD. This corresponds to 25% total harmonic distortion (THD) and a 4:1 S/N:

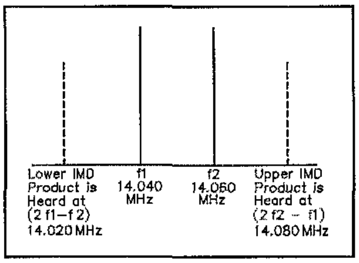


Figure 1—IMD shows up as false signals generated within your receiver. In the example above, a receiver may pick up IMD if it is tuned to 14.020 or 14.080 MHz. These IMD products are produced by the mixing of two strong signals in your radio at 14.040 and 14.060 MHz.

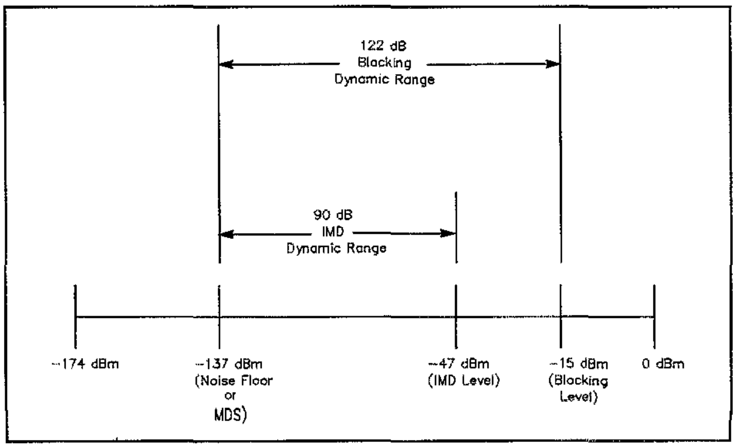


Figure 2—The relationship between MDS and dynamic range.

$12\ \text{dB SINAD} = 20 \log (1/[25\%]) = 20 \log 4$

The basic test setup for measuring SINAD is similar to the MDS hook-up. (The only difference is that the audio voltmeter is replaced by a distortion meter.) To measure 12 dB SINAD, the input signal level is adjusted for 25% THD as indicated by the distortion meter. The standard input signal typically used for narrowband FM SINAD measurements (such as used for amateur communication) is 3 kHz peak deviation, modulated at 1000 Hz.

Dynamic Range

Dynamic range is the difference between the weakest signal a receiver can hear, and the strongest signal a receiver can simultaneously accommodate without noticeable degradation in performance. Two types are considered in *QST* product reviews: *blocking dynamic range* (blocking DR) and *third-order intermodulation distortion dynamic range* (IMD DR).

Blocking dynamic range describes a receiver's ability to maintain sensitivity (or not to become desensitized) in the presence of a strong undesired signal on a different frequency. IMD dynamic range, on the other hand, is an indication of a receiver's ability to not generate false signals as a result of the two strong signals on different frequencies outside the receiver's passband. Both types of dynamic range are normally expressed in dB relative to the noise floor.

Evaluating the Dynamic Range Data

Dynamic range can be a major consideration, especially if you're in an urban or RF congested area (like a major city). When the IMD DR is exceeded, false signals begin to appear along with the desired signal (Figure 1). When the blocking DR is exceeded, the receiver begins losing its ability to amplify weak signals. The larger the dynamic range number, the better. Modern radios typically have an IMD dynamic range that is greater than 85 dB, and a blocking dynamic range greater than 120 dB (see Figure 2).

Dynamic range, especially blocking dynamic range, is sometimes reported in our test results as being "noise limited." This means we couldn't make the measurement. Phase noise is masking and interfering with the desensitization or IMD product we're trying to measure. In the case of blocking dynamic range, the normal 1-dB decrease in the desired signal could not be measured due to a 1-dB *increase* in the phase noise. This means the blocking dynamic range is greater than the noise-limited measurement.

A low noise-limited measurement is a clue that the receiver might be phase noisy. A particularly high noise-limited dynamic range indicates both good dynamic and range and low phase noise. In this case, the noise simply won out over the desensitization. If the noise-limited point is not specified, no conclusion can be drawn from the measurement.

While it's true that you can never have too much dynamic range, it's also not necessary to have more than you'll need. For example, with all other parameters equal, a receiver having 85 dB of dynamic range will not perform better than one that has 92 dB—*if only 60 dB is required!* Lots of dynamic range may also come at the expense of reduced sensitivity.

The amount of dynamic range you'll need depends on the presence of strong signals at your receiver's input. Consider all sources, especially nearby transmitters. Your antenna can also affect your dynamic range requirements. Its gain and directivity can enhance a desired signal while simultaneously rejecting the undesired. (As a typical example, consider a Yagi pointing toward a desired station and away from an offending signal source.) Small transmitting loops can reject unwanted signals with their narrow bandwidths and high Q. An antenna tuner can also be helpful in reducing unwanted signals.

Third-Order Intercept Point

Another parameter used to quantify receiver performance in *QST* product reviews is the third-order intercept point (IP3). This is the extrapolated point at which the desired response and the third-order IMD response intersect. Greater IP3 indicates better receiver performance.

Second-Order Intercept Point

We've recently added a new test to the battery: second-order IMD distortion dynamic range. These products, like third-order products, can also be generated within a receiver. In today's busy electromagnetic environment they can become offensive under certain conditions. Consider the case of two strong shortwave broadcast stations, on two different bands, that sum to the frequency of the weak DX you're trying to copy. If their intermod product is strong enough, you may not be able to copy the station through the interference.

Image and IF Rejection

Images are the unwanted results of the mixing process that takes place within a superhetrodyne receiver. They appear as second, false signals that are displaced on your radio dial by twice the local oscillator frequency.

A station operating at a receiver's intermediate frequency (IF) can wreak havoc if its signal is strong enough and the receiver's intermediate rejection is insufficient. The signal will appear across a radio's entire receiving range! Be sure to consider IF rejection when evaluating a transceiver—especially if you have local RF sources at the intermediate frequency.

We test both image and IF rejection in the Lab. We first measure the level that causes the unwanted signal to be at the MDS. Then, we reference this level to the MDS in order to express the measurement in dB.

Audio Power Output

A receiver's audio output capabilities are usually expressed as power (in watts) developed across a given load (usually 4 or 8 Ω) with a specified *total harmonic distortion* (THD). How much audio power do you need? That depends on a variety of factors, including speaker efficiency, ambient noise and personal taste. If you're a mobile operator on a noisy highway and your speaker is wedged against the underside of your car's console, you'll want more audio than someone in a quiet shack with headphones. On the other hand, 1 W of audio can be ear shattering in a small room with an average speaker.

Transmitter Tests

Some of the transmitter measurements reported in *QST* Product Reviews are power output, spectral purity, two-tone IMD, unwanted sideband and carrier suppressions,

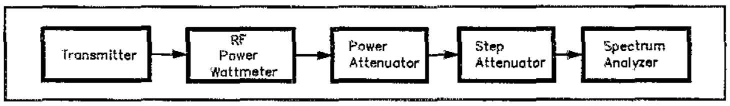


Figure 3—Testing for spectral purity and power output.

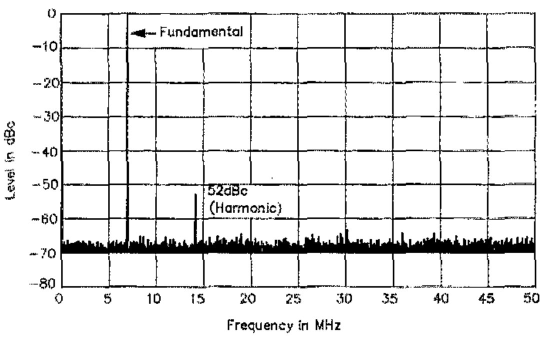


Figure 4—A typical spectral-purity plot.

and keying and turnaround tests.

Power Output

Figure 3 shows the test set-up for power output. It's measured using a laboratory grade wattmeter across each band for each available band and mode. The power attenuator simply serves as a dummy load for this test.

Most amateur transceivers are in the 100 W class. Some of them struggle to get there; others exceed their specification comfortably. These differences are not significant in most cases. The guy on the other end will not be able to hear a 5 or 10-W deviation from a transmitter's 100-W specification.

By the way, don't forget to consider a transmitter's duty cycle rating. Some modes, such as RTTY, are far more demanding than CW or SSB. If you intend to operate RTTY, you may have to reduce power or make short transmissions. At the other end of the spectrum, QRP operators must consider a transmitter's *minimum* power capability. It may not go as low as you'd like.

Spectral Purity

All transmitters emit signals outside their intended frequency range. These signals are called *spurious emissions*—a term that includes all signals that are not the fundamental and its desired modulation. They can include harmonics, parasitics, intermodulation products, noise and frequency conversion products.

We measure spectral purity in the Lab using the hook-up shown in Figure 3. The power attenuator again serves as a dummy load. In addition, it provides an output signal that is attenuated by 30 dB. The step attenuators then further reduce the signal as required for compatibility with the spectrum analyzer and its attenuator.

The spectrum analyzer displays signal amplitude with respect to frequency. Figure 4 shows a typical spectral purity plot obtained with an analyzer. Each vertical division is 10 dB; each horizontal division is 5 MHz. The signal fundamental is at 7.150 (or approximately 1.2 divisions from the left), and the attenuators and analyzer are adjusted to set the fundamental at 0 dB (or the top horizontal line). All other signals can now be referenced to the fundamental and are visible as lower-level pips on the display. The second harmonic at 14.300 MHz, for example, is at −52 dBc (dB referred to the carrier).

Part 97 (the FCC regulations that govern the Amateur Radio service) requires that all transmitters meet standards for the purity of their signals. Spectral purity must be in compliance with these regulations, not just to avoid a QSL from the FCC, but to ensure the transmitter doesn't interfere with other hams and other radio services. The FCC requirement for an HF transmitter operating at 100 W is that all spurs be at least −40 dBc (97.307(d)). The more negative this number, the better. Since our hypothetical transmitter's worst case is −52 dBc, it's spectrally legal. (Other requirements apply for different frequencies and power levels.) All transceivers and transmitters advertised in *QST* must meet the minimum FCC requirements for their frequency and power class.

Two-Tone IMD

Just like your receiver, your transmitter produces IMD distortion products. To measure IMD we use an audio generator. The generator produces two tones, one at 700 Hz and the other at 1,900 Hz. We put the transmitter in the desired SSB mode and modulate it with the two-tone generator. The output is measured by the special PEP-reading wattmeter and observed on the spectrum analyzer.

Figure 5 shows a typical IMD plot. Each vertical division is again 10 dB; each horizontal division is 2 kHz. You can see the two desired signals in the center of the display. All other pips spreading out to the left and right are the odd-order distortion products. In the example shown, the worst-case third-order product is 35 dB below PEP, the worst-case fifth-order product is 39 dB down and the worst-case seventh-order product is down 45 dB.

We test the transmitter in both upper and lower sideband; *QST* Product Reviews always show the worst case. Poor IMD performance can result in unnecessary interference to others on the band and illegal out-of-band products. The greater this number, the better. Typical modern transmitter should have IMD performance better than 25 dB below PEP for the worst case low-order product. High-order products must be much less.

Unwanted Sideband and Carrier Suppression

Single-tone audio-input signals can be used with the same setup as the IMD test to measure unwanted sideband and carrier suppression. To make this measurement, we set the single tone to the 0-dB reference line. The unwanted signal components can then be read directly from the display in dBc.

Unwanted sideband and carrier suppression is important to minimize bandwidth and nearby frequency interference. A modern transmitter should have at least 35 dB of suppression for these parameters. Better than 50 dBc is typical

Keying and Turnaround Tests

Our product review test battery includes three keying and turnaround tests:

- CW keying waveform test
- PTT to SSB/FM RF output test
- Transmit/receive turnaround test time

Each requires an oscilloscope and a keying test generator. Dual-trace scopes are best in most cases, and provide easy-to-read time-delay measurements between keying input and RF (or audio) output signals.

CW Keying Waveform

We measure the CW keying waveform and time delay using a keying test generator (see Figure 6). It is used to repeatedly key the transmitter at a controlled rate—typically 20 ms on and 20 ms off.

PTT to SSB/FM RF Output Test

For SSB and FM voice modes, a PTT-to-RF-output test is similar to CW keying tests. It measures rise and fall times, as well as the on/off delay times. An audio generator is required in the SSB mode and the transmitter is keyed with the generator connected to the mike input. This can be an important test for some digital modes.

Transmit Receive Turnaround Time Test

Turn-around time is the time it takes for a transceiver to switch from the 50% fall time

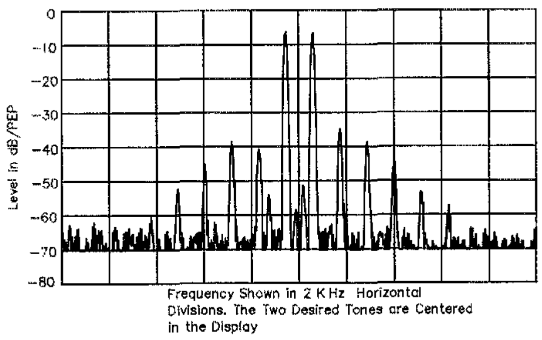


Figure 5—Transmitter IMD testing usually produces a plot like this.

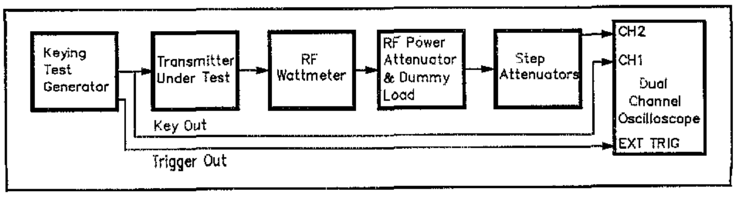


Figure 6—Block diagram of the CW keying waveform test.

Stepping Through the Process

Product Reviews are a team effort requiring a wide range of talent and dedicated individuals. The three primary players typically include the Product Review Editor, the Test Engineer and the Reviewer.

The Product Review Editor oversees the entire review and monitors its progress. The editor, with input from other Headquarters staff and members, first selects the equipment for review. His decision is usually based on anticipated member interest in the item, its uniqueness and its contribution to new technology. The editor then locates a suitable reviewer or reviewers, typically someone with product-related technical expertise. ARRL policy requires reviewers to be either members of the Headquarters staff (except for the Advertising Department) or Technical Advisors (TAs).

The ARRL Purchasing Department obtains all items for Product Reviews. They make every attempt to ensure that the integrity of all equipment is not compromised during its purchase. If a manufacturer, for example, could influence the actual specimen destined for a review, he or she might attempt to substitute a "gold plated" or specially tailored version of the product. To prevent this, the Purchasing Department obtains products off-the-shelf whenever possible and on a competitive-bid basis. They may solicit any number of dealers, and the sale is always made with the understanding that the manufacturer or distributor will not be notified of the purchase.

Whenever a procurement must be made through a manufacturer, or if a review item is particularly sensitive, Purchasing will obtain that item through an unidentified third party. As an added precaution, after making a purchase directly from the manufacturer, the review team compares that product's characteristics to a randomly selected sample.

Product Evaluation

Once an item arrives at Headquarters, the Product Review Editor submits it to the ARRL Lab for testing and evaluation. Depending on the product's complexity, this can take anywhere from several hours to a week. A typical HF transceiver, for example, is subjected to a battery of 21 to 25 tests, depending on its features. More tests are also possible if the product has any special features, functions or specifications. Products of this complexity typically require several days on the test bench.

Once the engineer completes all testing, it's the reviewer's turn. The reviewer is a volunteer who uses and operates the product as if it were his or her own, usually at home (or in his car, if appropriate). The review period usually lasts a month or more. Although doing a review can be fun, it's not easy. During this time, the reviewer must become thoroughly familiar with the product, its manual, its strengths, its weaknesses, and any problems that may have been encountered.

The reviewer's last step is to write a manuscript and forward it to the editor. The editor combines the edited text with any graphics and the test-result table. The table presents the measured performance characteristics along with the manufacturer's specifications and the FCC requirements. The data appears in a standardized format showing both the measured results and manufacturer's specifications.—*WA1SVF*

of a keying pulse to the 50% rise of audio output. Turn-around time is an important consideration with some digital modes. AMTOR, for example, requires a turn-around time of 35 ms or less for long-distance work.

Composite/Phase Noise Test

The composite-noise test measures the phase and amplitude noise, as well as any close-in spurious signals generated by a transmitter. We can assume, however, that nearly all of the noise observed during this test is phase noise. It's the primary noise component in any well-designed transmitter.

Frequency synthesizing circuitry is notorious for generating phase noise. Phase noise often manifests itself as broadband hiss caused by a phase-noisy oscillator chain in the transceiver. You'll hear it when you're tuned to a frequency adjacent to a strong signal.

Composite/phase noise is measured with a spectrum analyzer in the ARRL Lab—but not directly. Doing so would exceed the analyzer's dynamic range capabilities. (There's a tremendous relative difference between the strong desired signal and the low-level noise.) Added circuitry is required to remove the carrier while leaving the noise components unaffected.

The transmitter's CW output signal is converted down to a frequency near 0 Hz, using a signal generator as the local oscillator and an external mixer. This downconverted component of the original signal, as well as any unwanted heterodyne components that are generated by the downconversion process, are then filtered out, leaving only the downconverted signals that result from the noise and spurious signals that were present in the transmitted signal. The noise and spurious signals are then amplified (by a low-noise audio amplifier) and displayed on a spectrum analyzer.

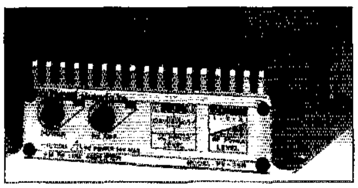
Conclusion

This issue contains the 226th *QST* Product Review. The reviews have grown tremendously in popularity since they first appeared, and the future promises even more growth. New products and technologies will also spawn new challenges. (I can't even imagine what procedures and instrumentation will be required for the 500th review.) We look forward to meeting these challenges and maintaining the standard of excellence you expect in *QST* Product Reviews. QST

New Products

VHF POWER AMPLIFIER

◊ If your hand-held VHF transceiver needs a boost, the all-mode PV-35R can help you maintain simplex contacts farther or hit a distant repeater. If you regularly use your hand-held rig in a vehicle, this unit will give your portable radio that "mobile rig" muscle. It's a 2-meter linear power amplifier with a built-in GaAs FET preamplifier. You can feed it 1 to 5 W input and get 30 W output (maximum claimed output power 35 W) and it draws 5 A at 13.8 Vdc. The preamplifier is rated at 10 dB gain with a noise figure of less than 1.5 dB. The front panel sports separate switches and LEDs for Power and RX (receive preamplifier) On/Off, three RF output-power LEDs (10, 20 or 30 W) and a 10.5 to 13.8 V supply voltage indicator; on the back are standard SO-239 (UHF) coaxial input and output connectors. The compact aluminum case with heat-sink fins measures just 4×4×2 inches. Retail price is $149. The PV-85R is an 85-W version for $249. An optional FM/SSB/AM/CW switch is $10. Naval Electronics Inc, 5417 Jetview Cir, Tampa, FL 33634; tel 813-885-6091, fax 813-885-3789. QST



KD7IK's "Quad Lite"

Do you think a 20-meter quad is out of your reach? Not this one!

By Jack Böck, K7ZR
7317 South Jewett Rd
Clinton, WA 98236

That—*thing*—in KD7IK's back yard is neither a giant spiderweb nor monster macramé. The 50-foot-high geometric pattern—composed of rope, wire, wood and wheel—is Randy Brink's full-size, rotatable, 2-element, 20-m quad. It's an antenna with very little visible means of support. But it works!

Randy is a veteran DXer in the 300+ country category. Twenty years ago, he fell from a tree while installing an antenna. That accident left Randy confined to a wheelchair. Since then, he's usually had to rely on outside help with antenna projects—*except for this one.*

Cause and Effect

Last year, Randy's tree-mounted, 3-element tribander was wiped out by high winds. Sky-high costs for professional tree-climbing services precluded repairing the antenna and led Randy to search for an affordable alternative. That alternative is the antenna shown in Figure 1 and the accompanying photographs.

Construction

The quad's suspended from a 3/8-inch diameter polypropylene rope strung between an 80-foot-tall tree in Randy's back yard and a 40-footer in the front yard. Despite his handicap, Randy erected the quad by himself. His antenna-raising equipment consisted of a Wrist-Rocket slingshot, a fishing reel, some monofilament fishing line and a handful of sinkers. Bracing the rod and reel against his wheelchair cushions, he lobbed line and sinkers over the trees and support rope. Then he wheeled around to the other side to tie on, pull up, or haul over as required. In keeping with house rules, Randy kept one wheel on the ground at all times.

A 1/4-inch-diameter poly line running through a pulley (hanging loosely a foot or so below the appropriate spot on the support rope) raises and lowers the entire array. This line is fastened to the experimentally determined balance point of the top boom, which is made from an 8-foot-long piece of 2×2-inch lumber. Two 16-foot horizontal spreaders are **U** bolted to the ends of the boom. These top spreaders are made of 10-foot lengths of 1/2-inch-diameter electrician's metallic tubing (EMT), with 6-foot EMT extensions attached to the 10-foot sections.[1] This is accomplished by running lengths of 1/2-inch-diameter wooden dowel inside the tubing sections. The bottom spreaders are constructed similarly, but are made of 1/2-inch-diameter PVC pipe.

The antenna elements can be made of any stranded wire #18 or larger. The loop circumference is determined by the formulas:

$$\text{Driven Element (in feet)} = 1005 \div F_{MHz}$$
$$\text{Reflector (in feet)} = 1035 \div F_{MHz}$$

The driven element is fed at the top center with RG-58 or RG-59 coax routed to the center of the top boom, then dropping down (alongside the hoisting line) to the bottom of the array. The driven element is split; the coax center conductor goes to one side, the shield to the other. The reflector element is a closed loop and, because its circumference is greater, the reflector's bottom spreader hangs somewhat lower than that of the director's. Element wires are simply taped to the spreaders with no attempt to connect to them electrically.

Lengths of 1/4-inch line drop from each end of the top boom, connect to the center of the PVC bottom spreaders, then continue down to another boom (an 8-foot piece of

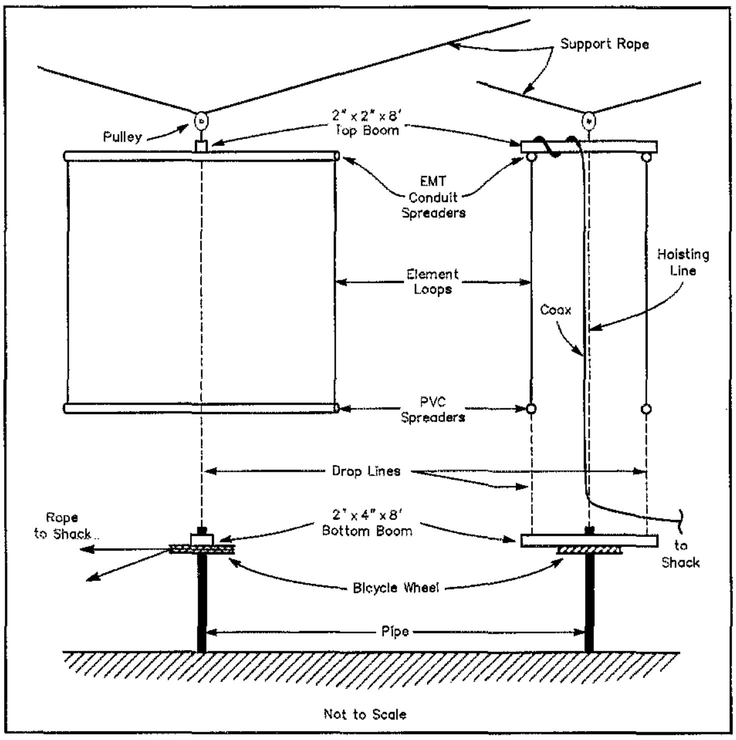


Figure 1—Details of KD7IK's quad. The support rope is strung between two trees: one in the front yard, another in the back yard. A pulley and length of rope secure the 2x2 top boom that separates the two 16-foot EMT spreaders. Element loops run between the top spreaders and the lower PVC spreaders. The latter are secured to the lower boom by drop lines. The hoisting line and coaxial feed line run between the two elements to the pulley and top boom. The bicycle wheel beneath the bottom boom allows the antenna to be rotated manually—by pulling on a rope around the wheel rim.

---

[1] EMT is inexpensive and readily obtainable, but it tends to rust, so it should be weatherproofed for longevity. Of course, aluminum tubing can be substituted for the EMT at higher cost.—*Ed.*

Figure 2—Randy, KD7IK, alongside the bottom of the quad. The water pipe supports the bicycle wheel rim and 2×4 lower boom. Rope around the wheel rim leads to the shack wall (in the background) allowing Randy to rotate the array by simply pulling on the ropes (the "armstrong method").

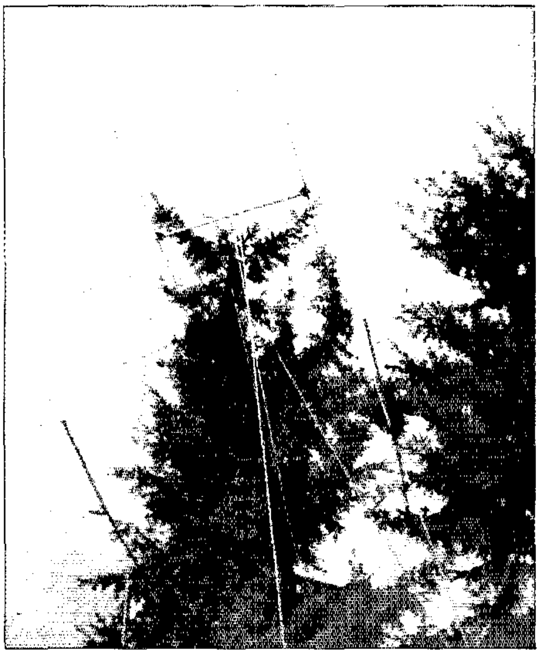


Figure 3—Here's the quad raised to its duty position ready to haul in the DX.

2×4-inch lumber) a few feet above the ground. Randy drove a length of 1½-inch-diameter water pipe into the ground directly under the pulley on the support rope. He fastened the bottom 2×4-inch boom across a bicycle wheel and slipped it over the pipe. The drop lines fasten to the ends of this 2×4. A ground-wire clamp attached to the pipe serves double duty as a wheel rest and a height adjustment for the lower 2×4. Once set, the drop lines are cinched up tightly.

Handling the Quad

When the bottom boom is rotated, the top boom follows neatly, swiveling on the short line extending from the pulley. Randy's shack is alongside the wheel at the bottom of the antenna. A line around the wheel, with both ends run through holes in the shack wall, permits Randy to rotate the antenna from inside the house using the "armstrong method."

It takes only seconds to raise or lower the array, making it a snap to dress up boom balance and line tightness. Because almost everything is tied together, few tools are required. The total cost, using all new parts, is less than $100. At the cost of a little more wire, additional director elements (set in the diamond configuration) could be draped over the support rope about 8 feet in front of the driven element. The added elements wouldn't be rotatable, but could add gain in the direction the support rope runs.

Summary

This quad operates as well as a rigid metal model. It sways a bit in gusty winds, although with little effect on signals. If a severe storm is anticipated, the whole array can quickly be dropped to the ground. Repairing any damage would be a snap.

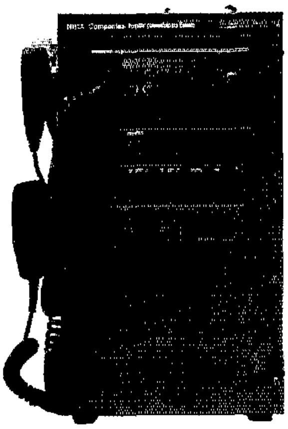
Randy maintains regular schedules with another handicapped ham, Victor Shinov, RW0CV, of Sovgavan, Siberia. Randy lives in Everett, Washington, a sister city of Sovgavan. For the past few years, the two hams have coordinated a number of cultural and student exchanges as well as medical assistance across the Pacific. Although Randy's tribander went kaput in January, this quad keeps these international goodwill efforts right on track.

Jack Bock was first licensed 1940 as W9KCY. During WW II, he was a radio operator for the AACS. After the war, with the assistance of the GI Bill, Jack attended the University of Michigan, obtaining his MA in Journalism and Political Science. Between 1960 and 1966, Jack worked as a technical representative for Philco and Burroughs in Japan, Okinawa and Thailand, then went on to Civil Service work as an antenna engineer in 1966, and also worked for the US Navy Comm Sta Japan from 1966 to 1972. After returning to the US in 1973, Jack retired to Whidbey Island to work DX in 1975. Jack edited the West Washington DX Club Newsletter from 1981 to 1987, and in 1989, went with a club group to the (then) USSR as guest of the Zilan DX Club (Kazan). Jack says "I'm now fully retired, sweating out the bottom of the sunspot cycle and listening to my arteries harden."

New Products

RADIO CONSOLE

◊ Instead of stacking mobile radios all over the shack or scattered around your vehicle with a rat's nest of cables, combine them in one handy, practical unit and be prepared for any operation or emergency communication activity. The Portable Communications Console (PCC) includes an internal 25-Ah lead-acid battery, 12-A external power supply and battery level indicator. As many as three of the newer radios from ICOM, Kenwood, Yaesu and so on fit into this handsome unit with its modular faceplate design. PCC models start at about $799, depending on configuration and options. Doug Wynn, KB6YZD, NIDA Companies, 2712 Foothill Blvd, Suite A, La Crescenta, CA 91214-3516; tel 818-957-1248, fax 818-957-0719.



A CW "Stamp" Identifier

You can tuck this CW IDer almost anywhere. It's about the size of a large postage stamp.

By Pat Bunn, N4LTA
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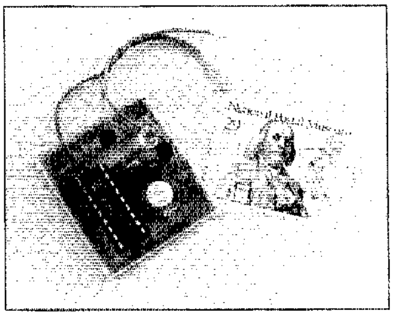
Spartanburg, SC 29302



To say that microprocessors have changed the way we do things is a bit of an understatement. It seems like every few months, a new generation of chips—with more speed, more math capability and more memory—is introduced. These devices now put the power of old mainframe computers within everyone's reach.

There's another version of the microprocessor that is less evident to the average person. This chip is hidden in the guts of your new automobile, your new transceiver and in many new electronic devices. These devices are designed to handle tasks such as timing, data gathering, and bit-oriented input and output. They are fast, powerful and becoming less and less expensive. Some of the most recent devices come in very small packages.

Background

In the past, most microcontrollers were produced with a masked memory that was programmed as the chip was made. Setting up the program mask was quite expensive, and for the microcontroller to be economical, a large number of them had to be sold. Prototyping chips with erasable programmable read-only memory (EPROM) were available, but these were relatively expensive for one-time use. During the past few years, a number of microcontrollers have been introduced with one-time programmable read-only memory. These micros allow you to program your instructions into the chip *once*. After the device is programmed, you can't erase the instructions, but the chips are inexpensive compared to EPROMs.

Microchip Technology has developed a line of reduced instruction set computer (RISC) chips that are quite compact and powerful: the PIC series. PICs range from a simple bit-oriented 16C54 to an advanced 17C42 with electrically erasable, programmable, read-only memory (EEPROM), direct LED drive and high speed. Many versions of the PIC series have built-in analog-to-digital conversion (ADC) and high-speed counters/timers.

In the PIC line-up, the PIC 16C54-RC/P processor is the simplest available. It costs about $5 in small quantities and doesn't require a crystal for its clock—it uses a simple RC circuit. Clock stability isn't critical and the clock speed can be varied on the fly. The 16C54 is also available with EPROM for approximately $16.

CW IDer

On many past occasions, I've needed a small CW generator: for a fox-hunt beacon, balloon launch ID, repeater ID, MEDFER and LOWFER beacon ID, ATV ID and VHF propagation beacon. Almost all of these uses require low power consumption, small size and low cost. I've constructed many IDers using the Intel 8748 series microcontroller with good results. The problem with the 8748 series chips is that they're power hungry and come in a 40-pin DIP package that takes up a lot of PC-board real estate.

The PIC 16C54 chip looked like a good candidate for an ID controller. It's packaged in an 18-pin DIP, uses only 3 mA at 6 V and can be purchased for about $4 in quantities of 10. The PIC 16C54 doesn't require a fragile crystal, allowing it to be used in rough service environments. A complete IBM PC-compatible programming package is available for $180.[1] This package allows you to program several other processors in the PIC series, including devices with ADCs and EEPROM. Although $180 may seem like a large cash outlay for programming equipment, similar equipment for some of the older microcontroller families costs several times more and is out of the reach of most experimenters. A source of programmed and tested PIC 16C54s is available[2] if you don't want to purchase programming equipment, and ready-made PC boards can be had, too.[3] The IDer described in this article is a simple implementation of the PIC 16C54-RC/P microcontroller. Only 2 of the 12 available I/O lines are used. Both are configured as output lines. One line, RA0 drives an LED (DS1) to visually monitor the keyed signal. The other line, RA1, drives Q1, a 2N3904 NPN switching transistor configured with an open collector output. The collector of Q1 is pulled low as the IDer keys. This configuration will drive most solid-state equipment. With vacuum-tube equipment, a transistor with a higher collector voltage rating, or a relay, may be required.

The microcontroller's clock signal is generated using a simple RC circuit comprised of R3, R4 and C1. R4 is a 100 kΩ trimmer potentiometer that allows the PIC's clock speed to be varied. Many microcontrollers require a very stable clock, but the PIC16C54 does not. In this design, the processor's clock speed is changed to vary the software timing and hence, the CW ID's speed. The adjustment isn't linear, but CW speed can be adjusted from several words per minute to over 30 wpm.

The PIC16C54RC/P is designed for 6-V operation. U2, a 78L06 voltage regulator, supplies the required voltage at just over 3 mA plus the current required by DS1 and Q1's base. Total current draw is approximately 13 mA. Eliminating DS1 will cut current consumption by one-third. If you have a 6-V source, the voltage regulator can be eliminated. With the regulator, the input voltage can range from 8 to 20 V.

The IDer's software is simple. It's of the "cheap and dirty" nature, but it works well and can be adapted easily for different messages. The example sends **DE N4LTA**, waits several seconds and repeats. It is composed of four subroutines that are called by the main program:

DIT sends a dot
DAH sends a dash
LETSPC waits one letter space time
WDSPC waits one word space time

To write a program to send Morse code, you simply call the proper subroutines to generate the required letters. A code sample is shown in Table 1. The timing is determined by the three-instruction routine, DITIME. This routine delays or waits for one dot time period. Two registers are loaded and then decremented to generate the time delay.

---

[1] Notes appear on page42.

Table 1
Code Sample for the CW IDer

The following example sends *ARRL*.

```
Call Dit            Call Dah
Call Dah            Call Dit
Call Letspc ;A      Call Letspc ;R
Call Dit            Call Dit
Call Dah            Call Dah
Call Dit            Call Dit
Call Letspc ;R      Call Dit
Call Dit            Call Wdspc ;L
```

When a dot is sent, The program turns on the two output pins (see Figure 1), calls DITTIME to wait one dot period, then turns off the two output pins. To send a dash, the same process occurs except that DITTIME is called three times. The entire program to send **DE N4LTA** uses only 67 bytes of program memory. The PIC 16C54 has 512 available bytes of PROM, so there's room for a fairly large message.

If you plan to do a lot of work with PIC controllers, purchase the EPROM version of the chip. That allows you to change your software until you get things exactly the way you want them. Then, you can program a cheaper ROM version and use your EPROM chip for another development project.

Building the IDer

Assembling the IDer is easily done with an etched PC board (see Notes 2 and 3). You can also build the IDer using wire-wrap or "ugly" construction methods, too. It's wise to use a socket for the PIC 16C54 just in case you want to change the program some day.

When installing the parts on the PC board, be sure you properly orient Q1, U1, U2 and DS1. Program the PIC 16C54 and install it. Make sure you get the PIC oriented properly. Check carefully for solder bridges and cold joints.

Connect a power source to the V_{cc} input. Any well-filtered dc source of at least 8 V and no more than 20 V will work fine. Current requirements are less than 15 mA. To test the IDer, turn on the power source. DS1 should begin flashing the CW message immediately on power up. If not, first check your circuit board for construction errors, then check your software. If you've changed the software, examine your changes carefully.

Summary

It's very easy to configure another I/O pin as an input port. This port could be used to start a message, rather than loop continuously. After receiving an input signal, the microprocessor would begin the message transmission. At the conclusion of the CW message, the microprocessor would wait for another signal from the input before sending the message again. This type of operation is suited for repeater ID use.

Other software can be developed easily. The PIC instruction set is simple, yet powerful. The PIC series of microcontrollers is well suited for other Amateur Radio uses such as keyers, display drivers, synthesizer controllers and a multitude of other possible uses. Let's see what you can do!

Notes

[1] One source of PIC programmers is Digi-Key Corp, 701 Brooks Ave S, PO Box 677, Thief River Falls, MN 56701-0677, tel: 800-344-4539, 218-681-6674; fax: 218-681-3880. (See page 78 in the Digi-Key catalog.) The Parallax unit (part number PIC-PGM-ND) is $179; Microchip's programmer (part number DV163003-ND) is $195.

[2] A kit consisting of the PC board and all parts—except the PIC 16C54—is available from 624 Kits for $12 postpaid. A custom-programmed PIC 16C54 (with up to 20 Morse characters) is available for $9 with the kit, or $12 postpaid if ordered separately.

[3] A template package containing the PC-board pattern and a part overlay is available from the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please include a business-size SASE and identify your request for the BUNN CW STAMP IDer TEMPLATE PACKAGE. If you're interested in a commented code sample, request the BUNN CODE SAMPLE and affix two First Class stamps to your business-size SASE.

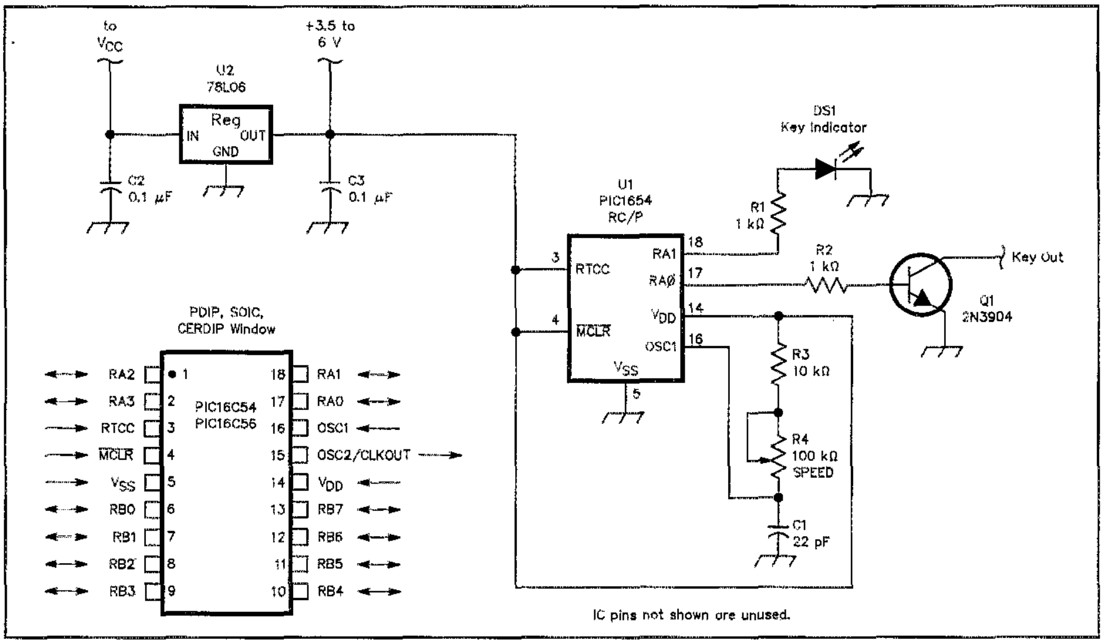


Fig 1—Schematic diagram of the PIC CW IDer. Equivalent parts can be substituted. Unless otherwise specified, resistors are 1/4-W, 5%-tolerance carbon-composition or film units. Capacitors are 50-V disc ceramic types.

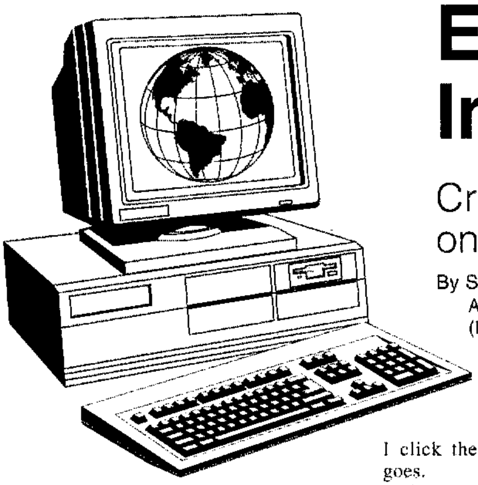
R4—100-kΩ trimmer potentiometer (DigiKey #36C15-ND).

C1—22-pF, NP0 ceramic-disc capacitor.

DS1—LED, T1-3/4.

U1—PIC 16C54 RC/P (DigiKey #PIC16C54-RC/P-ND).

Misc: 18-pin DIP socket, PC board (see Notes 2 and 3).



Exploring the Internet—*Part 2*

Cruising through ARRL Headquarters on the information superhighway.

By Steve Ford, WB8IMY
Assistant Managing Editor
(Internet: sford@arrl.org)

ARRL Headquarters is no stranger to *cyberspace*—the computer-based universe that some call the "information superhighway." Our little corner of cyberspace begins within the building itself. Every office computer is linked by our local area network, or *LAN*. We use Novell *Netware* version 3.12 in a *client-server* system that currently supports 114 connections. A "server" is a centralized computer that holds programs and data that the staff can access from computers at their desks (the "clients"). Our LAN is connected to the Internet for electronic mail (e-mail) and other functions.

With all the publicity about Internet (it made the cover of *Time* magazine!), hams are exploring cyberspace in increasing numbers. The result has been an explosion in our Headquarters Internet activity. Thanks to Internet, we receive an ever-increasing amount of e-mail correspondence. In addition, the ARRL e-mail *Infoserver*, which I'll explain in a moment, logs more than 30,000 requests each year.

E-mail: Fast and Convenient

E-mail has become a favorite method of communicating with ARRL Headquarters. The reason is easy to understand. Not only is e-mail extraordinarily fast, it's convenient for everyone concerned.

For example, e-mail messages arrive directly at my office from anywhere on the globe. I might be using my computer to write an article, but the background connection to the ARRL LAN is always ready for incoming mail. I hear a soft beep whenever a message reaches my computer. With a click of a mouse button, I leave the article momentarily and display the message on my monitor.

If I can answer the question or comment right away, I will. I simply tap out a reply and click the **send** button. If it looks like someone else could give a better answer, I click the **forward** button and away it goes.

Because of the nature of my job, I'm in front of my computer most of the time. It's relatively easy for me to respond quickly to whatever mail arrives. This isn't true for everyone at Headquarters. Even so, staffers usually check for e-mail several times a day. Reading through the batches of e-mail that accumulate overnight is a morning ritual for most of us.

We try to answer messages in a timely fashion, but you may not receive a reply the same day. Like other correspondence, e-mail messages are answered according to their urgency. If several days pass without a response, the person may be on vacation, or traveling on League business.

In the sidebar "Headquarters Internet Addresses" you'll find e-mail addresses for a number of individuals on the Headquarters staff. If you don't have a direct Internet connection, don't worry. Almost all on-line services (CompuServe, Delphi, America On-line, Prodigy, GEnie and so on) will allow you to send and receive Internet e-mail. For example, you can send a message to me from CompuServe by addressing it as follows:

internet:sford@arrl.org

Each service uses a different method, so read your user manual or ask the customer service department. Some services charge a fee for messages going to or from the Internet. The charges are usually nominal, but check with the service to make sure. If you send a message to us from an on-line service via Internet, our reply will travel via Internet back to you.

We also log onto the on-line service at least once each day to check for mail. If you leave a message for us, it will be downloaded and routed to the appropriate person. Here are the ARRL addresses on various services:

CompuServe:	70007,3373
Prodigy:	PTYS02A
GEnie:	ARRL
America On-Line:	HQARRL
MCI Mail:	2155052

If your service has an Internet connection, however, it's always best to use it for communicating with Headquarters directly. Your message will reach its destination much faster.

Using the ARRL Infoserver

The information server, or Infoserver, is an automated system that gives you access to many information files relating to various facets of Amateur Radio. You can retrieve any file by simply sending an Internet e-mail message. Each file you request is then mailed to you automatically.

To request something from the Infoserver, send a message to:

info@arrl.org

The server ignores the subject line, so you can leave it blank. The important information—the Infoserver commands—go in the body of your message. They are:

reply <address>

This command is optional. You only need to use it if the FROM: line in your message does not contain a valid Internet address. In most cases, it will. If you use this command, it must appear on the first line of your message.

help

Sends a handy Infoserver "help" file.

index

Retrieves a list of available files.

send <filename>

Use this command to retrieve the file you want. You can ask for more than one file in a single message by placing **send** commands on separate lines.

quit

This command *must* appear on the last line of every message sent to the Infoserver.

Let's say you want to grab the latest W1AW transmission schedule and a list of kit suppliers. The body of your message would look like this:

```
send W1AW.SKD
```

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Table 1

A Sampling of Files Available from the ARRL Infoserve

Filename	*Size*	*Description*
PROSPECT.TXT	2k	How to get your Amateur Radio license
EXAMS.TXT	52k	Current exam schedule info—updated bi-weekly
EXAMINFO.TXT	9k	Examinations, what to bring, requirements
ARRLCAT.TXT	39k	Catalog of ARRL Publications—commercial content
JOIN.TXT	2k	How become an ARRL member
SERVICES.TXT	5k	A condensed list of ARRL membership services
TOUR.TXT	28k	An electronic tour of ARRL Headquarters
DIR.HQ	5k	Visiting ARRL Headquarters—directions and tour information
HFBANDS	7k	Breakdown of users of HF spectrum
Q-SIGS	1k	A list of Amateur Radio Q-signals
W1AW.SKD	2k	W1AW schedule of transmissions and operation
PRODREV1.TXT	12k	Which rig is best? Part 1—*QST* Lab Notes
PRODREV2.TXT	22k	Which rig is best? Part 2—*QST* Lab Notes
ILIST.TXT	6k	*QST* Bibliographies List
RFIGEN.TXT	37k	How to solve an EMI/RFI problem—*QST* Lab Notes
RFISOURC.TXT	13k	Where to buy filters, EMI-proof telephones, etc
ADDRESS.TXT	16k	Ham/electronic company addresses
KITS.TXT	6k	A list of companies that sell kits
BBS.TXT	12k	A list of ham-radio landline bulletin boards

send KITS.TXT
quit

The Infoserver will receive your message within minutes. Almost immediately the requested files will be on their way back to you. If you're using an on-line service, find out if they impose restrictions on the maximum size of incoming e-mail messages. Some Infoserver files are large. Table 1 is just a sample of the files you'll find on the Infoserver.

It's a good idea to request the index and help files in your first message to the Infoserver. This will set you up to get the most efficient use out of the system. If you have any questions or comments, send an e-mail message to Mike Tracy at mtracy@arrl.org. Don't write comments in your messages to the Infoserver. They'll either be deleted or they will cause the server to reject your request.

Question and Answers

Q: Do ARRL Headquarters staff read the USENET Amateur Radio newsgroups?

A: Some do, when they get a chance. Most of us at Headquarters are quite busy, so it's difficult to spare the time to read through the newsgroups. If you have a question or comment, send it to us directly. Don't post your comments to a newsgroup in the hope that we'll see it.

Q: Can I connect to the ARRL BBS via Internet?

A: No. The BBS doesn't have an Internet connection at this time. To connect to the BBS, you have to call 203-666-0578. The BBS accepts up to four connections simultaneously. If the BBS doesn't respond (the phone just keeps ringing), chances are it's not off-line. If all four lines are in use, the fifth call isn't answered. Try again later.

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Use your Internet e-mail capability to order books and other materials from ARRL Headquarters. Simply turn to the *Publications Catalog* in this issue and write down the items you want to order along with their order numbers. Compose an e-mail message with the following information:

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- Your name as it appears on the card
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- A daytime telephone number
- The order numbers of the items you want
- The titles or descriptions of the items and the quantities

Send your e-mail order to:
ltardette@arrl.org

Shipping and handling charges will be added to your order total and billed to your credit card. Your order will be on the way as soon as possible!

Q: I want to check on the status of my DXCC application. What address should I use?

A: Send your inquiry to dxcc@arrl.org.

Q: Can I send my contest logs to the ARRL via electronic mail?

A: Yes. Send them to contest@arrl.org. The file must be in ASCII text and must conform to the ARRL contest file format. A description of the format is contained in a file you can request from the Infoserver (CONTEST.DAT). You can also use this address to submit Special Event and Contest Corral information for publication in *QST*.

Q: I want to find the addresses of ham clubs in my area. How do I do this via e-mail?

A: Simply send your city and state to ead@arrl.org. In most cases a list will be sent to you via US mail, so include your complete home address.

Q: Can I get an FCC Form 610 sent to my home?

A: No problem. E-mail your postal address to gswanson@arrl.org.

Next Month

Scott Ehrlich, WY1Z, takes the helm for parts 3 and 4 of our series. Next month he'll guide you through the wonders of *telnet*, *archie* (no, not the comic books!) and *ftp*.

Update

Compuserve users now access the Internet USENET newsgroups. When you're connected to CompuServe, GO INTERNET for more information.

New Products

CT KEYBOARD LABELS

◊ Custom Key Overlays for *CT* is a set of *self-adhesive keyboard labels for the popular* contest-logging software (versions 8 and 9) by Ken Wolff, K1EA. The QSOs are coming in fast, but you don't remember all the commands in K1EA's *CT* computer software? When the contest gets hectic—or the operators just get a bit groggy—it's easier to press the proper key if all the special ones are labeled and color coded. These overlays are precut to fit the keytops on standard IBM-compatible computer keyboards. They show all the commands available on all the keys programmed by *CT*, including all shifted (Shift, Alt, Ctrl) key combinations. Retail price is $14.95 plus s/h. Kenneth Wells, V73C, High-Rate Products, PO Box 1255, APO AP 96555-0008; tel 805-238-7994. Available exclusively from Oklahoma Comm Center, 13424 Railway Dr, Oklahoma City, OK 73114; tel 800-765-4267

FEEDBACK

◊ A New Products announcement on page 70 of August 1994 *QST* listed an incorrect telephone number for Maldol USA in Seattle, Washington. The correct number is 206-525-9158. We apologize for any inconvenience.

◊ A New Products announcement on page 39 of September *QST* listed an incorrect price for the Schurr mahogany base *straight* key. The correct price is $249.95. We apologize for any inconvenience.

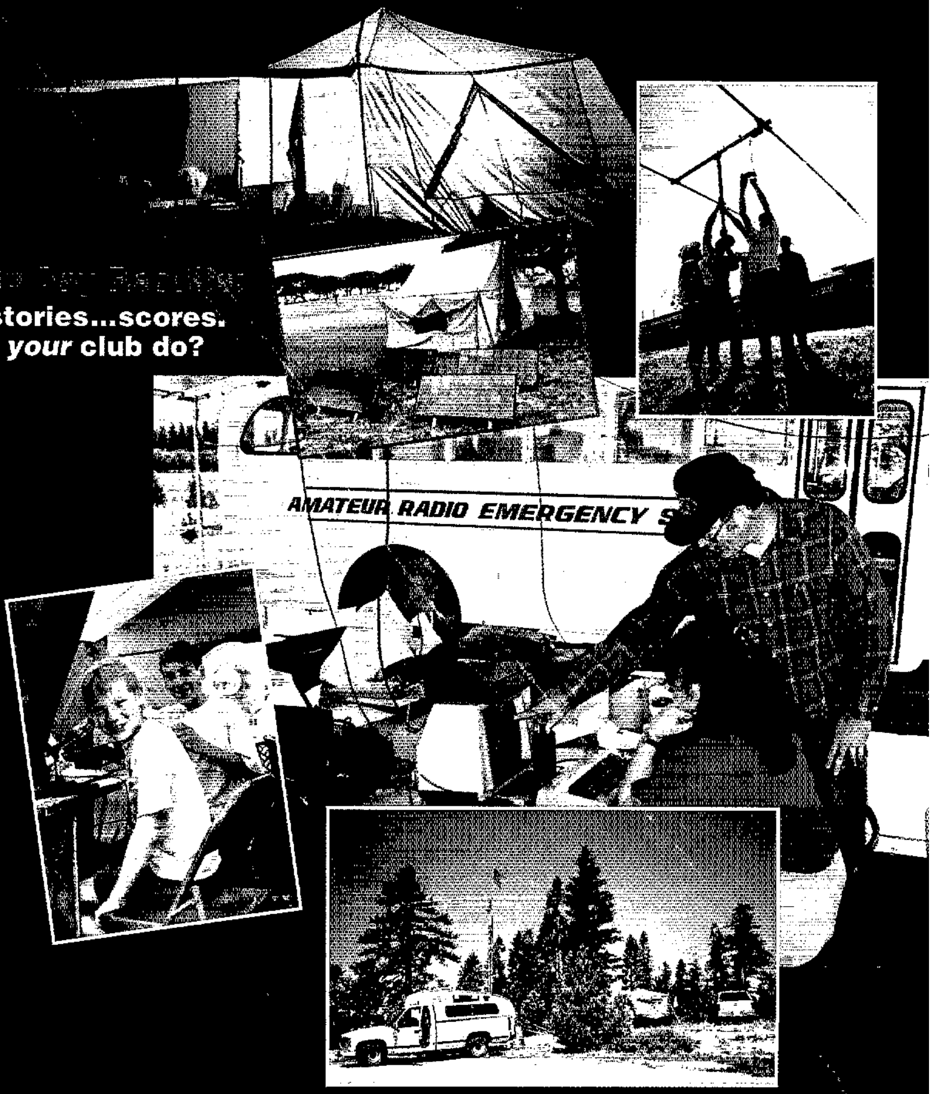
QST

November 1994

devoted entirely to Amateur Radio

1994 Field Day Results

Photos...stories...scores.
How did *your* club do?



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OUR COVER

Field Day '94:

Set against the late-night glow of USECA's tent: KC6HOY and the Siskiyou County ARES bus (center); KH6WO's solar panels (upper left); the W6JN crew raising an antenna (upper right); WG6N and company mountaintopping (lower right); and (lower left) middle-schoolers Brad Parker, KB1AKA (rear), and Patrick Conroy (awaiting his call sign) at N1II, learning how to beat us *all* a few years from now!

CONTENTS

November 1994
Volume LXXVIII Number 11

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

An ATV Station for 915 MHz
Part 1—A Miniature 1.5-W ATV Transmitter

Getting on ATV can be simple and inexpensive —especially if you're willing to have some fun building your own gear!

By Rudolf Graf, KA2CWL
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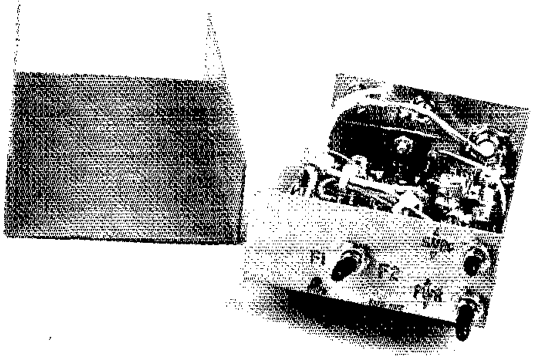


BRAD THOMAS, KC1EX

The 902 to 928-MHz amateur band, herein referred to as 915 MHz, can be an interesting introduction to microwave work. Traditionally, the microwave spectrum is considered to start at 1000 MHz; 915 MHz is at the doorstep of this interesting region. Antennas for 915 MHz are small and, although somewhat larger than those commonly seen at 1296 or 2304 MHz, parabolic antennas, cavities, and even waveguides are practical here. Midway between 440 and 1296 MHz, 915 MHz is an excellent stepping stone for amateurs interested in 1296-MHz work, but who would like to get a bit more UHF experience before attempting operation on 1296.

There's a good supply of obtainable used test equipment with an upper frequency range of 1 GHz. The tremendous growth of the cellular telephone and the wide acceptance of 800 MHz for two-way radio has made available a number of low-cost RF transistors suitable for use at 915 MHz. For additional reading, see two other articles by the authors on related ATV equipment for 440 MHz.[1,2]

General Description

This article describes the construction of

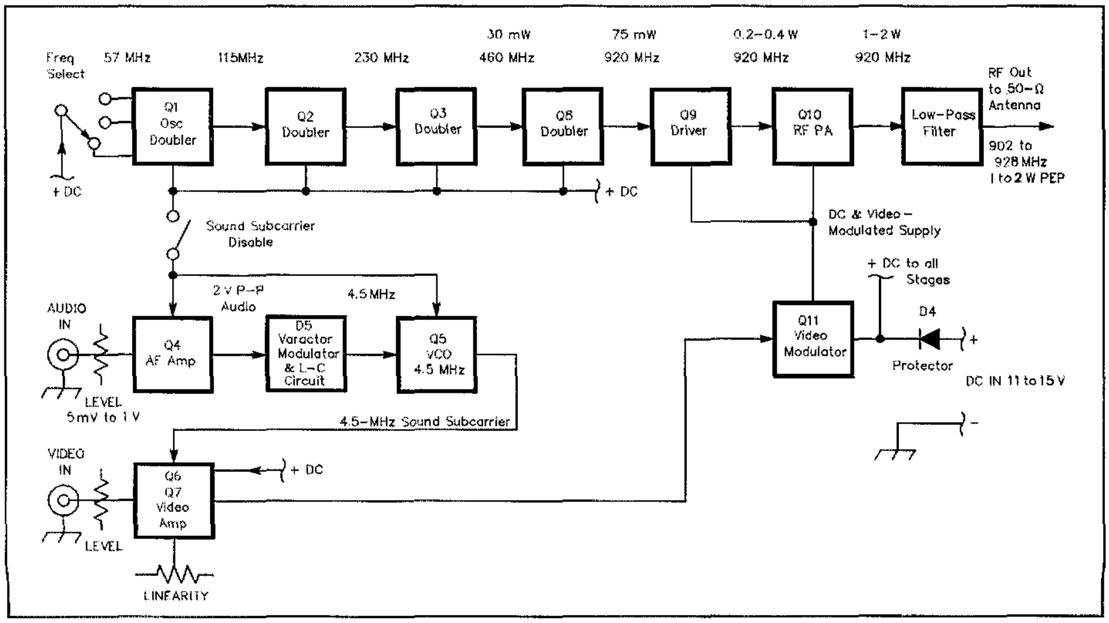
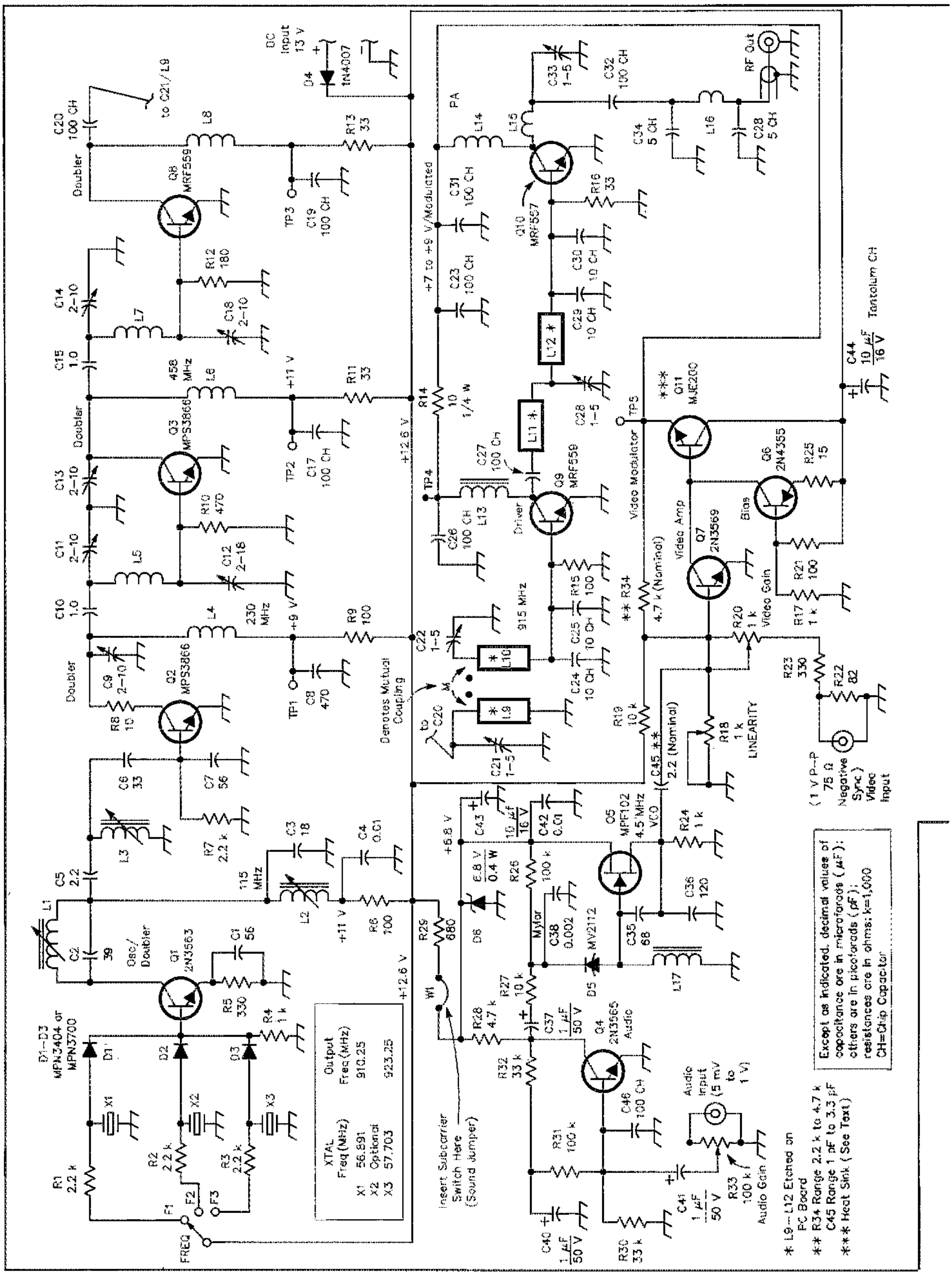


Figure 1—Block diagram of the 915-MHz ATV transmitter.

[1] Notes appear on page 29.



XTAL	Freq (MHz)	Output Freq (MHz)
X1	56.891	910.25
X2	Optional	
X3	57.703	923.25

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

* L9—L12 Etched on PC Board
** R34 Range 2.2 k to 4.7 k
C45 Range 1 pF to 3.3 pF
*** Heat Sink (See Text)

Figure 2—Schematic of the 915-MHz ATV transmitter. Unless otherwise specified, all resistors are 1/8-W, 5%-tolerance; equivalent parts can be substituted. There are some minor deviations from standard *QST* part identification to ensure compatibility with the available kits.

C1, C7—56 pF NP0
C2—39 pF NP0
C3—18 pF NP0
C4, C42—0.01 disc
C5, C45—2.2 pF NP0
C6—33 pF NP0
C8—470-pF, 20%-tolerance disc ceramic
C9, C11, C13, C14, C18—2 to 10 pF trimmer
C10, C15—1 pF NP0
C16—Not used
C12, C37—2 to 18-pF trimmer
C17, C19, C20, C23, C26, C27, C31, C32, C46—100-pF chip
C21, C22, C28, C33—1 to 5-pF trimmer
C24, C25, C29, C30—10-pF chip
C28, C34—4.7 or 5-pF chip
C35—68-pF NP0 or silver mica
C36—20-pF NP0 or silver mica
C38—0.0022 µF, 50 V Mylar
C39, C40, C41—1 µF, 35-V or 50-V electrolytic
C43—10-µF, 16-V electrolytic
C44—10-µF, 16-V tantalum chip
D5—MV2112 variable-capacitance diode
D6—1N754, 6.8-V, 0.4-W Zener diode
L1-L17, inclusive—See Table 1
R14—10 Ω, 1/4 W
R18, R20—1-kΩ potentiometer
R33—100-kΩ potentiometer
S1—3-position slide switch
X1—56.890625 MHz (910.25 MHz)
X2—As required (not supplied with kit; see Note 1)
X3—57.703125 MHz (923.25 MHz)

Misc: PC board, enclosure, SPDT (on/off switch), LED (power on indicator), TO-220 mica insulator, heat sink (G-10 PC board 5/8 inch high × 1 3/4 inches wide; see text), 2 phono jacks, ferrites for coils, BNC female chassis-mount connector, dc power connector, #22 enameled wire, #32 enameled wire, LED, miscellaneous hardware.

a small (2.5×4×1-inch), 1.5-W transmitter for 915-MHz ATV use.[3] The transmitter is stable, crystal controlled, offers switch selection of three channels, a sound subcarrier generator, operates with dc supplies ranging from 11 to 15 V and has excellent "cable-quality" video. This project is easy to duplicate and can be tuned up with commonly available equipment. Although not recommended as a first-time project, amateurs who've homebrewed some RF gear should have no trouble duplicating this transmitter.

Because this is an AM transmitter, it can be used for AM voice work as well, simply by using the high-level video modulator for audio or data signals. (Don't knock good, old-fashioned AM! It's the easiest form of modulation to detect, requires little test equipment, and is better than FM for very weak signals, although not as good as SSB.) AM gear is easy to homebrew—good news for the ham with limited test equipment. A simple downconverter and a general-coverage receiver can be used à la 2-meter AM in the pre-FM and repeater days. Forget signal-to-noise ratio, efficiency, etc. At 915 MHz, static is nearly nonexistent and good AM sounds great and works well. Ask the airlines—they still use it for communication.

This transmitter produces a complete NTSC video signal and a 4.5-MHz FM sound subcarrier, modulated on a UHF carrier in the 915-MHz amateur band (33 cm). This UHF signal is above the normal tuning range of a TV receiver, but can be received on any TV receiver outfitted with an RF downconverter capable of tuning 902 to 928 MHz. A dedicated downconverter with output on an unused VHF channel (channel 3 or 4) can be used in a receiving setup for these frequencies. This also provides an optimized RF front end and best reception range. The transmitter's small size is compatible with many small PC-board cameras and is small enough for R/C applications in which a video link is needed, surveillance, amateur TV, and for use in a video H-T using one of the pocket TV receivers as a display.

With minor adjustments to video drive, power supplies delivering 11 to 14.4 V (including lead-acid, NiCd, or alkaline battery packs) can be used. (Use of a power supply delivering less than 11 or more than 15 V is not recommended.) Typical power output is 1 to 2 W PEP over the given voltage range. We installed crystals for 910.25 and 923.25 MHz with space for one other crystal at some future date. (Operation requires a crystal whose frequency is determined by dividing the desired output frequency by 16.) You'll need third-overtone, series-resonant crystals with frequencies between 56.375 and 58.000 MHz.

The transmitter requires a standard NTSC or PAL video signal. The audio can also be retuned for a 5.5-MHz sound subcarrier used by the PAL system. The video input is standard 1-V P-P, 75-Ω, negative sync.

Circuit Description

Refer to Figures 1 and 2. Q1 is a crystal-controlled oscillator using any one of three crystals chosen by three PIN diodes in a dc-controlled selector circuit. Only one bias resistor receives voltage from frequency selector switch S1. The crystal's associated PIN diode is forward biased, inserting it in series with Q1's base and ground. The two unbiased diodes appear as open circuits, so the other two crystals are effectively out of the circuit. If you need fewer than three channels, simply omit the associated components.

Q1's output network couples second-harmonic energy at 112.5 to 116 MHz to Q2, which doubles the signal frequency to 225 to 232 MHz. Q2's output network rejects unwanted frequencies and matches the input impedance of the next stage, Q3, another doubler. Q3 doubles its input signal to an output of 451 to 464 MHz.

At the output of Q3, about 50 mW of RF is present. This is fed to Q8, the last doubler stage. Q8 produces 75 to 100 mW of RF in the 902 to 928-MHz range. Q8 feeds a double-tuned matching network consisting of several components. L9 and L10 are small enough to be etched on the PC board. This network is tuned to 902 to 928 MHz, the final output frequency. This network rejects all other multiplier products. Careful layout and construction are important here.

RF fed to driver Q9 is amplified to about 400 mW. RF choke L13 feeds video and dc to Q9. The base impedance of final amplifier Q10 is only a few ohms, hence the seemingly large values of capacitance. L11 and L12 are part of the PC board. This network helps filter out residual spurious unwanted signals. Q10 amplifies the signal to about 1.2 to 2 W, depending on the supply voltage. Matching network L15 and C33 match the impedance of the collector of Q10 to the load impedance (50 Ω). RF choke L14 feeds video and dc to Q10. C34, L16 and C28 form a low-pass filter to ensure a spectrally clean RF output to the antenna. Q10 handles considerable power and it is heat-sinked through its leads directly to the copper foil ground plane of the PC board to help dissipate heat and to provide a low-inductance emitter return.

To video modulate the RF carrier, the supply voltage for Q9 and Q10 is taken from the emitter of modulator Q11. Q6, Q7, and Q11 are connected as a feedback amplifier with a nominal gain of about 5 to 10 (determined by R34 and the setting of R20). Video-signal input to the transmitter is amplified in this feedback amplifier circuit and inverted so that sync and black levels are positive-going. The video signal at Q11's output is superimposed on the supply voltage to Q9 and Q10, causing amplitude modulation of the RF output of Q10. R18 sets the carrier level of the RF output for symmetrical modulation. R20 is set for maximum modulation level without distortion or clipping. D4 protects the circuitry against reverse polarity. Note that modulator Q11 is effectively in series with the power supply to the RF power amplifier and dissipates considerable power. Therefore, Q11 is heat-sinked. The video amplifier is dc coupled and attention should be paid to video-input dc levels and interfacing considerations.

Subcarrier Audio

A 4.5-MHz sound subcarrier generator is needed for audio. Incoming audio is fed through **AUDIO GAIN** control R33 to Q4, which has a gain of about 43 dB. Audio from Q4 is fed to varactor diode D5, which controls the frequency of VCO Q5, resulting in frequency modulation of the 4.5-MHz sound subcarrier frequency. Subcarrier from Q5 is fed into the video modulator via C45, where it is mixed with the video. The value of C45 sets the subcarrier level. A jumper in series with the power supply to regulator diode D6 and R29 permits disabling Q4 and Q5 (if sound is not required) via an external switch. This switch is optional and can be omitted if subcarrier disabling is not needed.

If, for instance, a separate communication channel on 2-meter or 440-MHz FM is used, and no 4.5-MHz subcarrier is needed or wanted, simply omit Q4 and Q5 and their associated components as they play no part in video transmission.

The transmitter draws around 6 to 8 W at 13.8 V and typically outputs 1.5 to 2 W peak power on sync tips. This means that around 5 W may be dissipated. Because this transmitter is physically small, some attention should be given to thermal considerations. The suggested heat sink (a piece of G-10 PC board, 7/8 inch high and 1 3/4 inches wide) is adequate for intermittent use (3 minutes on, 3 minutes off) and normal tune-up and testing. The transmitter will run cooler if additional heat-sinking is provided. The G-10 shield can be replaced with a metal (copper or brass) plate, which can be fastened to a chassis or a radiator fin. The dc supply should be clean and stable as any noise or ripple in the supply voltage will modulate the RF carrier and/or shift the carrier level, causing sync clipping or interference in the received video. Less than 100 mV of ripple and noise is required, less being desirable. This should not present any problems, as most quality supplies easily meet these specifications.

Construction

This is fairly straightforward, but a few pointers are in order. All trimmer capacitors and grounded resistor leads are soldered *on both sides of the board.* This is essential for good RF grounding. Certain parts—such as the chip capacitors and Q8, Q9 and Q10—are mounted on the *bottom* of the board. *All* parts are mounted tight and close to board: *no* exceptions. Good UHF construction techniques are a must. Note all ground jumpers in the layout diagram and see that they are installed (see Note 3). (Plated-through holes are not used on the PC board as that precludes easy home fabrication.) Mount Q11, metallic side down, to the heat sink. Install a mica insulator between Q11 and the heat sink. Install the heat sink and Q11 vertically on the PC board. Solder as much of the seam as possible on both sides of the heat sink. An inside view of the transmitter is shown in Figure 3.

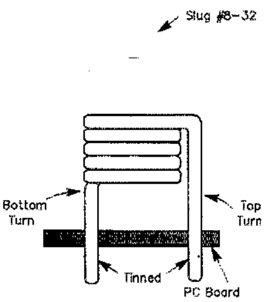
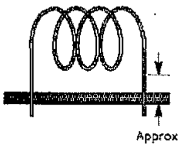
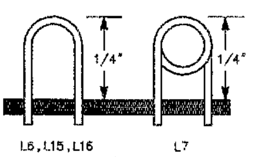
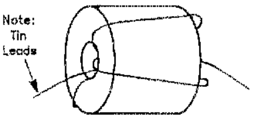
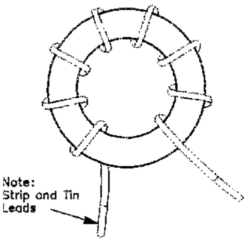
Be sure that no coil turns short together on L4, L5, L7, L8, and L14. See Table 1 for coil-winding information. It may be wise to make L6 and L15 slightly larger than specified to allow room for final tuning. L17 has an extra turn that may need to be removed; this is determined during final test and adjustment.

Tune-Up

The successful testing and tuning-up of this transmitter is dependent, to a large extent, on your understanding of the circuit's operation and having access to some basic test equipment. If you're going to be experimenting with amateur TV operation, you'll need this equipment anyway for everyday use. The minimum required equipment includes:

- A VOM or DVM. A nonelectronic, old-fashioned VOM is preferred simply because

Table 1
Coil-Winding Information

Osc Circuit L1, L2, L3  Coil Must Fit Flush With PC Board After Installation	L1—6½ turns L2 and L3—3½ turns L1-L3, inclusive, consist of #22 enameled wire wound on a #8-32 screw used as a coil form. After winding the coils and forming and tinning the leads, remove the screw and insert a #8-32 ferrite slug inside the coil.
1st Doubler Circuit L4, L5 915-MHz RF Chokes L8, L14  	L4—4 turns L5—5 turns L8 and L14—3 turns L4, L5, L8 and L14 are made of #22 bare, tinned wire wound on a #8-32 screw coil form. After winding and forming the leads, remove the screw and space the coil turns evenly so as to fit the existing PC-board holes. Mount each coil 1/16 inch above the PC board.
2nd Doubler Circuit L6, L2 PA Output Inductor L15  Mount on PC Board as Shown	L6, L15 and L16—½ turn L17—1½ turns L6, L15, L16 and L17 are made of #22 bare, tinned wire wound on a #8-32 screw coil form. After winding and forming the leads, remove the screw and mount the coils as shown.
RF Choke L13  3 Turns #32 Enameled Threaded Through Ferrite Bead	L13 consists of 3 turns of #32 enameled wire threaded through a ferrite bead.
4.5-MHz Toroid L17  13 Turns #22 Enameled on 3/8" Ferrite Toroid	L17 consists of 13 turns of #22 enameled wire space wound on a 3/8-inch-diameter ferrite toroid. (Use 11 turns of #22 enameled wire for PAL operation at 5.5 MHz.) Note: L9 through L12, inclusive, are etched on the PC board.

they are easy to read and interpret while making peaking adjustments.

- A 50-Ω dummy antenna good up to 1000 MHz, preferably one with a power meter. Most ham, CB and hobby dummy antennas sold for HF and 2 meters use are useless at 915 MHz. A relative reading is OK as long as the termination is a good 50 Ω at 915 MHz. If you don't have one, you can make a very good dummy antenna for 450 MHz and higher from a 100-foot length of RG-58 coaxial cable. (See the sidebar "A Simple Dummy Antenna for ATV Transmitter Testing.") Terminate the far end in a detector of some kind that shows any response at 915 MHz. A relative reading is sufficient. See the sidebar drawing for a suitable detector circuit.
- A good dc power supply with 0 to 15-V variable output. Regulation should be 1% or better from no load to full load. Its current capability should be greater than 1 A, preferably with current limiting. Hum and noise amplitudes should be less than 0.1 V at any load. Most good-quality experimenter power supplies meet these requirements.
- A source of standard-level video (1V P-P, 75 Ω, negative sync) and an audio source. A B&W or color camera, camcorder, or VCR will serve. In a pinch, you can use an audio source to check the video modulator.
- A TV receiver capable of tuning 902 to 928 MHz. Usually a TV receiver tuned to channel 3 or 4, and fed with a tunable ATV downconverter is employed. A B&W or color set is OK. (You'll need this for two-way communication, anyway.) Commercial TV receivers don't tune to frequencies this high. Some older, continuously tuned UHF tuners reach 890 MHz (channel 83) and it may be possible to modify one to tune up to part or all of this range, but their performance will be poor. In a pinch, you can use a scanner covering this frequency range. An ICOM R-7000 is ideal for use as a receiver if you have or can borrow one.
- A frequency counter or other means of measuring 4.5 MHz and, if possible, up to 1000 MHz. A general-coverage shortwave receiver and/or wideband scanner can also be used.
- Cables and test leads.
- In addition, the following equipment is helpful, but *not* necessary:
- An oscilloscope with a bandwidth greater than 20 MHz.
- A spectrum analyzer or wideband receiver (ICOM R-7000, etc.)
- A dip meter covering up to 250 MHz or more.

Don't try to tune this transmitter solely for "best picture!" You'll get nowhere—fast! First, the RF circuits must be checked and tuned for best performance and maximum output into a 50-Ω load. Then, certain modulator adjustments are made. Only after this is done is video connected to the transmitter. At this point, only minor adjustments will be found necessary. Video quality should be excellent, so don't settle for less.

A Simple Dummy Antenna for ATV Transmitter Testing

Coaxial cable acts as an effective attenuator, especially at 915 MHz, because of its high losses. A dummy antenna (see Figure A) can be made from a length of coaxial cable that feeds a diode detector a sample of transmitter output. The coax also terminates the transmitter in a 50-Ω load that is better than anything you probably can make or buy cheaply. The 10-pF capacitor and RF choke remove RF components and the 1-kΩ resistor provides a load for the detector. Use a hot-carrier diode because the 1N914, 1N34 and 1N270 used at lower frequencies are ineffective at UHF. The 47 or 51-Ω resistor value isn't critical, but the resistor should be a carbon-composition unit—not film—or you can use a chip resistor for lowest inductance.

The detector output amplitude depends on the input power and cable loss, but will be 0.1 to 1 V with a 1 to 2-W output transmitter, assuming a cable loss of about 15 dB. To observe the detected video, use an oscilloscope with 10-MHz bandwidth and a low-capacitance probe, or connect a scope to the detector output via a short cable (6 inches or less) to avoid the loss of high-frequency video components. (For dc-metering purposes, such precautions are not necessary.)

Caution: Do not use less than 50 feet of cable and do not use this circuit below 400 MHz: The cable attenuation may not be the 10 dB or more needed for good attenuation and to provide a decent 50-Ω termination for the transmitter. Above 1300 MHz, measurement errors may result, too.—*William Sheets, K2MQJ*

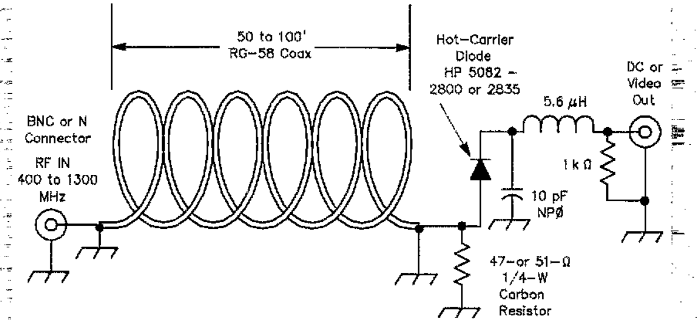


Figure A

PC-Board Checkout and Transmitter Tune-Up

First, visually check the PC board for errors. Then, connect a dc supply to the circuit with the negative lead to the ground foil and the positive lead to the looped end of D4. The supply should be set to 0 V. If your supply has no ammeter, place one in series with the positive lead to D4 (you can use a VOM). Now, *slowly* raise the power supply's output voltage to 13.8 V, watching the ammeter. Less than 100 mA should flow. If over 150 mA is being drawn before reaching 13.8 V, there may be a short or other problem. Watch for any smoking components. If all is OK, then check for following voltages with negative lead of the VOM connected to ground. Keep the supply set at 13.8 V.

Collector Q1	> +12.5 V
Collector Q2	> +12.5 V
Collector Q3	> +12.5 V
Junction R29 and D6 (jumper installed between R29 and R28/D6)	+6.8 V
Collector Q4	+3 to +4 V
Junction of D5/C38	+6 to +6.8 V

Next make the following ohmmeter checks:

Junction of C28/L16 to ground	Infinity
Q9 base to ground	100 Ω
Q10 base to ground	33 Ω
Q9 collector to TP5	10 Ω
Collector Q10 to TP5	<0.1 Ω
Junction L15/C33 to TP3	< 0.1 Ω
Junction C27/L11 to ground	33 Ω

If these checks are satisfactory, preset the **VIDEO GAIN** (R20) and **LINEARITY** (R18) controls to center. Connect the positive VOM lead to the emitter of Q11. Rotate R18. The voltage should vary smoothly between at least 2 to 12 V. If it doesn't, try moving R20 a little in either direction. (R20 varies this voltage, but to a lesser degree than R18.) Leave R18 set for about 7 or 8 V at Q10's emitter.

Preset sound trimmer C37 to half mesh (plates half meshed). Connect a frequency counter to the emitter of Q11. You should be able to obtain a stable reading somewhere around 4.5 MHz. Disregard any erratic or unstable reading. If you cannot get a stable reading, try loading the emitter of Q11 with a 330-Ω resistor connected to ground. If needed, adjust R18. Set C37 for a reading

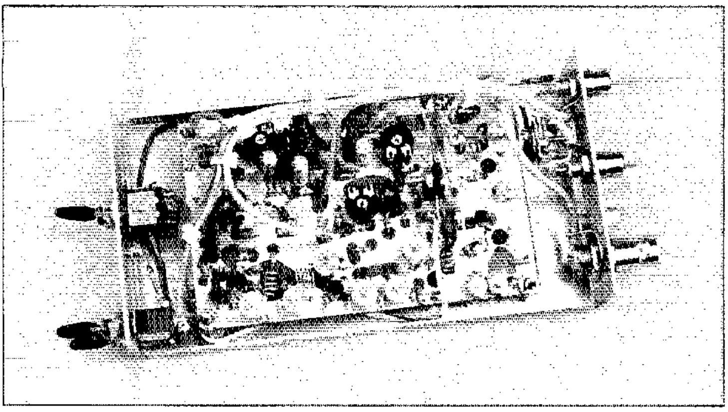


Figure 3— An inside view of the 915-MHz transmitter. With this photo and the title-page picture, you can get a good idea of the location of the hardware mounting. In this view, the crystal-selection switch is at the top; the power-on and subcarrier-off LEDs are beneath the switch and on each side of it. The audio subcarrier selection switch is mounted above the power on/off switch. On the rear panel, from top to bottom, are the audio and video-input phono jacks and IF-output BNC connector. The dc-input jack is mounted between and below the audio and video-input jacks. Near the right end of the PC board is the final amplifier heat sink and shield, made from a piece of PC board and mounted vertically.

between 4.495 and 4.505 MHz. If you're unable to reach this frequency, add turns to L17 if the frequency is too high, or remove them from L17 if the frequency is too low. An oscilloscope can confirm the presence of the subcarrier signal at the emitter of Q11.

If a frequency counter is not available, you can use a short-wave receiver tuned to 4.5 MHz—its antenna lead near Q11—to listen for this carrier. You can adjust this later. The main objective is to see if the circuit is functional. When done, disconnect the 13.8-V supply and, if you used it, the 330-Ω resistor from the emitter of Q11.

Review the circuit description section, then rig up a test lead so that you can enable each crystal in the oscillator circuit by applying 13.8 V dc to R1, R2, or R3, one at a time. This selects X1, X2, or X3 respectively. If you are using fewer crystals, you need only to connect to those positions. Now preset the coil and capacitor adjustments as follows:

L1, L2, L3	Slugs half inserted
C9	30% meshed
C11	50% meshed
C12	50% meshed
C13	30% meshed
C14	30% meshed
C18	20% meshed
C21	40% meshed
C22	50% meshed
C28	25% meshed
C33	25% meshed

Note: These figures are typical and some variation is to be expected. You should, however, come fairly close after tune-up.

This step aligns the oscillator stage. If only one crystal frequency is used, this step is not needed. Connect the positive lead of your VOM to TP1. Apply 13.8 V dc to board. Enable X1. Now, slowly back out the slug in L1. *Use only a nonmetallic tool.* Watch for a sudden small drop in the voltage at TP1. It will start out at full supply voltage and drop 0.3 V or more. Back out the slugs in L2 and L3 to maximize the voltage drop. It's possible to obtain a drop of 1.5 V or more. Enable X2 and repeat this procedure; do the same with X3. Go back and forth until you can get as much drop as possible with each crystal without readjustments. Some compromise may be necessary.

Connect a VOM to TP2. Enable the highest-frequency crystal and adjust C9, C11 and C12 for the lowest obtainable voltage at TP2. Enable the other crystals in turn and repeat the adjustments for lowest voltage. Then, readjust L2 and L3 to get the voltage at TP2 as low as possible with all crystals in turn. Again, some compromises may be needed. The object of this exercise is to get Q3 to draw 30 to 60 mA with each crystal used. This indicates adequate RF drive to the next doubler (Q8) at all crystal frequencies. Assuming use of a 13.8-V supply, the voltage between TP2 and ground will be between 11.1 and 12.1 V. Note that the trimmer capacitor settings should not be radically different from the presets if your coils have been made correctly and the chip capacitors are properly installed. If you have difficulty in obtaining the stated results, any radical variation in the trimmer settings may indicate problems or incorrect tuning. This should be regarded as suspicious in case other problems are encountered. The tuning networks have wide range and if coils are made incorrectly, it is possible to *triple* instead of *double* frequency in the Q2 and Q3 stages. Locate and correct any problems before proceeding.

Next, tune doubler Q8 for maximum output at 902 to 928 MHz. You should connect a dummy antenna to the RF output as sometimes the PA produces some output during the Q8-stage tune-up. That's because the presets are often quite close to their final settings since L9-L12 are printed on the circuit board. Place the positive meter lead on the supply line and the negative meter lead on TP3. Tune C13 and C14 for maximum voltage reading. This could be 3 to 5 V, as Q8 typically draws 135 mA. Leave C18 at preset position for now. If results seem anemic, you can repeak C9, C10, and C12 leaving the meter connected to TP3. (We recommend repeaking previous stages as there is generally some interaction between adjustments.)

Now, we'll tune-up the PA and make the video checks. Remember: 1.5 W of RF at 900 MHz is a fair amount of RF energy. It can upset a lot of nearby electronics. Therefore, always use a nonradiating load for tune-up and testing. Don't place any parts of your body close to a radiating antenna that is used with this transmitter, especially a gain antenna. This means maintaining a distance of 1 to 10 feet between yourself and the antenna, depending on antenna gain. Remember, too, that a number of otherwise fine video devices are very poor at rejecting strong RF fields. They may be made inoperative or even damaged by them. So, keep radiating antennas at least 10 to 20 feet from any devices used with this transmitter unless proper RF shielding measures can be taken if needed. Allow a distance of at least 5 feet between the transmitter and the antenna to avoid feedback. Otherwise, you may be fooled into tracing transmitter "problems" that do not exist. In short, use common sense.

Connect the transmitter RF output cable to a dummy antenna and some kind of RF indicator known to be good at 915 MHz. A relative indicator is okay, but you'll not be able to measure actual power output. Preset R18 fully clockwise so that nearly full supply voltage appears at the emitter of Q11. Check the resistance to ground at the emitter of Q11. If the reading is 10 Ω or less, a short exists in the RF power-amplifier stages (Q8 and Q9). Eliminate the short before applying power. If the reading is a few hundred ohms or more, all is okay. (Reverse the ohmmeter leads and take the higher reading as the correct one.) Next, apply +13.8 V to the transmitter at D4. Leave the crystals disabled. Measure the voltage at Q11's emitter and verify that R18 can vary this voltage down to less than 3 V. Next, connect the positive meter lead to the 13.8-V dc lead, attaching the negative lead to TP4. Reset R18 for full voltage. Connect a lead from 13.8 V to R1, R2, or R3 to enable the highest frequency crystal. While closely watching the meter, tune C21 and look for a small change in the reading. Adjust C22 and C18 for an increase in the reading. Then, go back and readjust C21, then C13, C14, C18 and C22 for maximum reading. You are adjusting the tuning for maximum collector current through Q9. By now you should see some RF output from Q10.

If your power supply is equipped with an ammeter, you'll see the total current drain increase to 400 mA or more. Next, adjust C28 and C33 for maximum RF output from the transmitter. Retune C21 and C22 for maximum RF output, then adjust C28 and C33. Finally, repeak all the trimmer capacitors for maximum RF output.

Check with other crystal frequencies and readjust as needed. For wide frequency spreads between crystals (greater than 15 MHz), some small compromises may have to be made. If all is well, you should be able to obtain 1.2 W or more output with a supply voltage of 13.8 V. The supply current should be between 500 to 600 mA. Although Q8, Q9, Q10 and Q11 run rather warm, nothing should get too hot to comfortably touch for a few seconds or so. While observing RF output, reduce the supply voltage to 9 V. The transmitter should still produce some RF output at a reduced level. If not, adjust the supply voltage to the point at which RF output is noted and retune L1, L2, L3, and C9, C10, and C12 for best output. Then, recheck operation at 13.8 V dc. It may not be possible to achieve optimum tuning at both voltages without resetting the trimmer capacitors, but the trimmer settings should be close to each other.

If operation at lower supply voltage is important, retune the entire transmitter at 11 V after confirming operation at 13.8 V. Full power output cannot be expected at lower supply voltages, but 0.5 W or so output should be possible with an 11-V supply. The transmitter performs best with power-supply voltages between 12 and 15 V. Do *not* exceed 15.5 V!

Expect to spend a little time optimizing everything and don't be afraid to experiment! You can always realign from scratch if you mess things up too badly. Actually, none of the adjustments is critical and once you get the feel of it, the alignment is quickly done. Run the transmitter under load for about 2 to 3 minutes. The heat sink will get quite warm, but not too hot to touch for a few seconds. This is normal. If this bothers you, use a larger heat sink or mount another on the existing one. For continuous duty, increased heat-sinking is recommended if cooler operation is desired.

Now we'll set up the transmitter for video and audio. Remove dc power and connect a source of video to the video-input terminals (junction of R22/R23 and ground). Make sure you have a 1-V P-P, negative sync, NTSC or PAL video input signal. Connect audio to the audio-input terminals; line-level audio is okay. Preset R18 and R20 at their midpoints and R33 at 1/4 open (9 o'clock). Tune a TV receiver to the transmitter frequency (902 to 928 MHz). Enable the highest-frequency crystal and turn on the video and audio sources. Apply 13.8 V dc to the transmitter. You should get a picture on the TV set. Adjust R18 for a stable picture, then adjust R20 for best contrast. Readjust R18 and R20 for best video and maximum output power without either black or white clipping on received video. If results seem poor, make sure you are not overloading the TV receiver or experiencing RF feedback to your video source. Adjust R33 for best audio. Recheck C37 and reset if needed to correct audio distortion.

All Done!

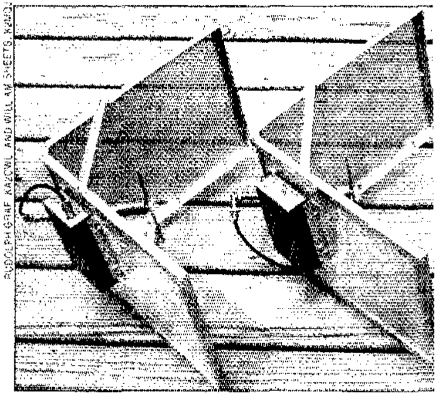
This completes testing. The transmitter draws about 500 to 650 mA at 13.8 V, depending on the modulating signal. Don't exceed the power-supply limit of 15.5 V, or expect optimum results at less than 11 V. Remember that power-supply garbage can modulate video and a drop in supply voltage can cause sync clipping. If low voltage is expected, set R18 at this voltage to avoid this problem. You can now mount the transmitter in a case and add jacks, controls and other items. Remember to allow adequate heat sinking and access to adjustments that may be needed at a later date.

Summary

Use the transmitter with a good antenna. At 915 MHz, antennas are small and a good (>10 dB) gain antenna should be used. The three most important factors determining range of transmission are: (1) antennas; (2) antennas and (3) antennas. We ran tests between the home and a mobile receiving setup using 10-dB-gain corner reflectors at the transmitter and receiver (see Figure 4). These tests were run in hilly country in northeastern New York state. Excellent initial results were achieved over a path of 7 to 15 miles, with the transmitter at K2MQJ's home and a 12-V TV receiver and downconverter in the mobile setup. If there were no high hills in the path, a picture could be received at many locations. The receive antenna fed a downconverter with output on channel 3. The antenna height at the receiver was that of the author's Chevrolet Blazer, roughly 6 feet above ground. At times, the antenna was inside the vehicle, facing out the window. We even detected a P3 picture at one spot 17 miles from the transmitter. The transmitter was mounted on the corner reflector antenna, 25 feet off the ground. The results were quite good.

In all fairness, the transmitter location was on a hill, but the (frankly) lousy receive setup probably negated the good transmitter location. A good Yagi antenna setup at reasonable height (30 feet or more) with greater than 17 dB gain should easily triple this distance if used at both ends. For test purposes, portable use or demonstrations, smaller antennas can be used. You can even use 3-inch quarter-wave whips, but don't expect more than about a mile of useable range line of sight with such small antennas. From experience, in a given location, range correlates pretty much with antenna size. Remember that feed-line losses are high at this frequency, so if you can, mount the transmitter close to the antenna.

By the way, this setup, using two corner reflectors (such as those shown in Figure 4), makes an excellent demonstration of amateur television. If you have never played with ATV, it's quite impressive to see the person with whom you're having the QSO. A path of several miles can be covered with this setup, more with two good Yagis or parabolic reflectors. This is adequate for many uses. There are a number of ATV repeaters around the USA; if you are near one, they'll extend your range. (See *The ARRL Repeater Directory* for information on ATV repeater locations.) Only a camera (color or B&W), a TV set with downconverter, and a 13-V supply are needed. Our security camera had a built-in microphone, but you can use an external mike or omit the mike if no audio is needed.



RUDOLPH GRAF, KA2CWL AND WILLIAM SHEETS, K2MQJ

Figure 4—The 915-MHz transmitter and a companion downconverter mounted on homemade corner reflector antennas.

A companion 915-MHz downconverter for use with this transmitter will be featured in an upcoming issue of *QST*.

Notes

[1] R. Graf and W. Sheets, "Amateur TV Transmitter," *Radio Electronics*, Jun 1989, pp 45-50 and Jul 1989, pp 45-50. (This article describes a 2-W ATV transmitter for 440 MHz and is the transmitter from which the 915-MHz unit described here evolved.)

[2] R. Graf and W. Sheets, "Amateur TV Station," *Electronics Now*, Jul 1994, pp 33-40 ff, and Aug 1994, pp 39-46 ff. (This article describes a 5-W, 440-MHz ATV transceiver and a related 0.75-W, miniature ATV transmitter.)

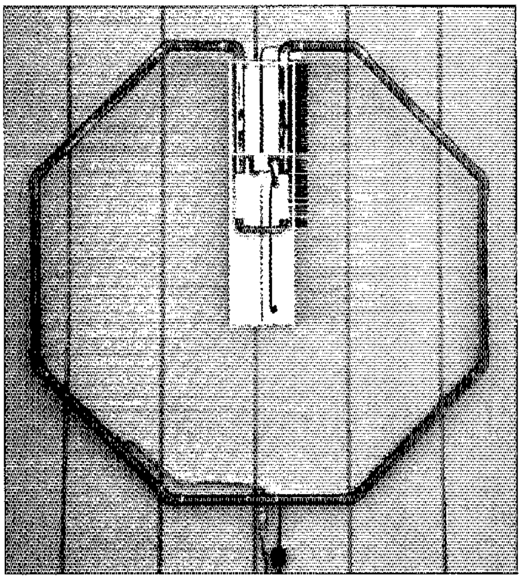
[3] A complete kit of parts for this transmitter, consisting of the PC board and all parts that mount on it is available from North Country Radio, PO Box 53, Dept. R, Wykagyl Station, New Rochelle, NY 10804-0053. The kit includes two crystals for the 910.25 and 923.25 MHz ATV frequencies and detailed construction information. Parts such as connectors, switches, jacks, wire, hardware and the enclosure are *not* included in the kit. Price: $139 US plus $4.50 shipping and handling. (New York residents please add sales tax.) Money orders and personal checks accepted. (Please allow time for personal checks to clear.) Charge cards are not accepted. Catalogs describing 25 other kits, cameras and accessories are available for $1 and an SASE bearing 75 cents First Class postage. (A free catalog is shipped with each order.) For technical assistance with North County Radio kits only, call 518-854-9280 between 9 AM and 5 PM Eastern.

A PC-board template package is available from the ARRL. Request the GRAF AND *SHEETS 915-MHZ ATV TRANSMITTER* TEMPLATE from the Technical Department Secretary. Please include a #10 SASE with 52 cents of First Class postage. QST

A Home-Brew Loop Tuning Capacitor

If you're itching to build a loop antenna and haven't found a suitable tuning capacitor that handles high voltage, your search has ended!

By Bill Jones, KD7S
83 Redwood Ave
Sanger, CA 93657
(Photos by the author)



The KD7S loop-tuning capacitor used with a 3-foot-diameter loop antenna. The antenna tunes from 10 through 20 meters and is rated for 100 W input.

Judging from a number of recent on-the-air contacts, interest in small-diameter HF loop antennas is on the rise. Enterprising hams are discovering that a 3-foot-diameter circle of copper tubing can be used to work the world, even when mounted a mere foot or two off the ground.

Ted Hart, W5QJR, showed us how to design and construct highly efficient, small loop antennas.[1] His simple equations make it easy to lay out a loop to suit virtually any need. Hart's work is required reading for anyone interested in such an antenna.

The Difficult Part

A common stumbling block in building a small transmitting loop is finding a suitable tuning capacitor. Most junk-box air-dielectric variable capacitors are unsuitable for use with loop antennas because they are lossy. A loop capacitor's stator and rotor plates must be securely fixed—soldered or welded, not merely clamped—to their respective supports to keep resistance to a minimum. Mechanical wiper contacts are undesirable for the same reason. Hart overcame this problem by using a split-stator capacitor. Each end of the loop is connected to one of the two stators, and the rotor acts as a coupler between the sections.

Feeding a small loop antenna with 100 W produces a kilovolt-level potential across the capacitor. To withstand such high voltages, wide capacitor-plate spacing is needed. Considering the effects of humidity and environmental pollutants, ¼-inch spacing between the plates of an air-dielectric variable capacitor is desirable.

A Different Approach

The home-brew tuning capacitor shown in Figure 1 and the accompanying photographs is a low-cost alternative to commercial units. You can easily construct this capacitor in a day or two using simple hand tools. As a bonus, the need for an extremely low-speed tuning motor is offset by using a simplified drive mechanism.

Mechanically, the capacitor consists of two lengths of ½-inch-ID copper pipe that telescope in and out of two pieces of ¾-inch-ID pipe. The larger-diameter pipes are analogous to the stators of a conventional split-stator capacitor; the smaller-diameter pipes are the rotors.[2]

In my capacitor, the center-to-center spacing between the two larger pipes is 3¾ inches. The smaller pipes have a 1/16-inch layer of Teflon plastic sheet wrapped around the outside to form a high-voltage insulator. The dielectric constant of Teflon also serves to increase the overall capacitance. This entire assembly is mounted on a ¼-inch-thick piece of ABS plastic. An ⅛-inch plastic cover completes the enclosure.

A 180-rpm gear-head motor turns a length of #8-32 threaded brass rod soldered to the crossbar that holds together the smaller pipes. Depending on the rotational direction of the motor, the smaller pipes are either pulled into, or pushed out of, the larger pipes. The larger pipes are soldered directly to the loop. This construction results in an extremely low-loss capacitor capable of withstanding very high voltages.

You can construct the capacitor to provide virtually any desired capacitance range. The formula for calculating the capacitance[3] is

$$C = \frac{0.224KA}{d}(n-1) \qquad \text{(Eq 1)}$$

where:

C = capacitance in picofarads (pF)
n = number of plates
K = dielectric constant,
A = area of one plate (in square inches)
d = spacing between plates

To calculate the surface area of the smaller tubes, multiply pi by the outside tube diameter times the length. A ½-inch-ID type M copper pipe has an OD of 0.625 inches. Substituting the values in the formula, we have $A = 3.14159 \times 0.625 \times 1 = 1.9635\ \text{in}^2$ for a 1-inch length of pipe.

As mentioned earlier, the Teflon insulation increases the capacitance as a result of its dielectric constant (2.1). The spacing, d, between the inner pipe and outer pipe is 1/16 inch, or 0.0625 inch. We calculate the capacitance of a 1-inch length of tubing using the following values: n = 2; K = 2.1; A = 1.9635, and d = 0.0625

$$C = \frac{0.224 \times 2.1 \times 1.9635}{0.0625}(2-1) \qquad \text{(Eq 2)}$$

Equation 2 yields a value of 14.778 pF. This value is for a *single* capacitor section. Because the sections are connected in series, the value is halved, to 7.389 pF per inch. Measurements show approximately 5 pF of stray capacitance for a 10-inch-long capacitor. This stray capacitance must be added to the overall value. Therefore, with a 1-inch section of capacitor, the actual value is about 12 pF. As the pipes telescope, the capaci-

[1]Notes appear on page 32.

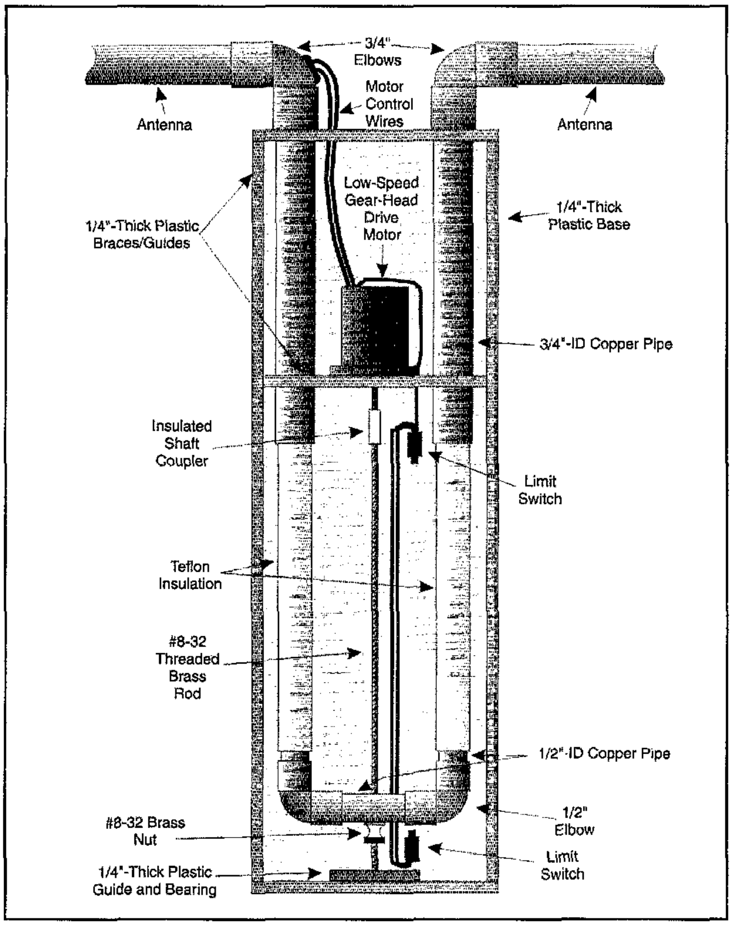


Figure 1—Mechanical drawing of the KD7S loop-tuning capacitor.

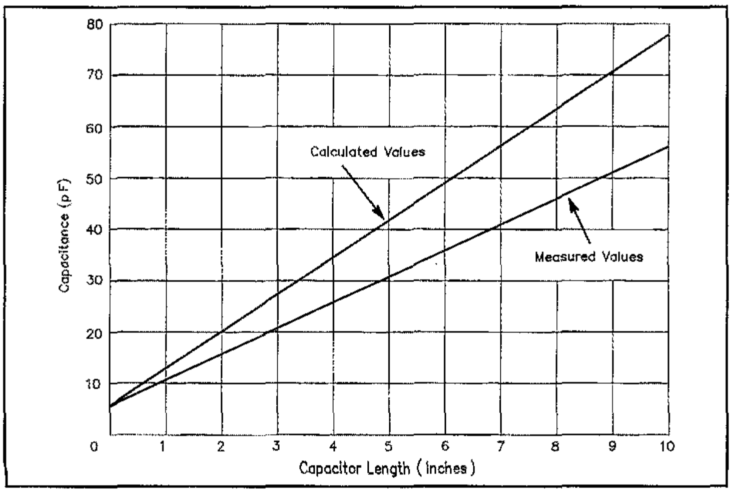


Figure 2—Comparison of the calculated versus measured values of capacitance using Equation 1. Stray capacitance must also be taken into account (see text).

tance increases linearly. For a 2-inch capacitor, the capacitance would be just under 20 pF, and so on.

Capacitance Variation

After I built the first capacitor, I discovered the measured capacitance to be somewhat less than the calculated value. These differences are shown by the graph in Figure 2. The actual value of capacitance was about 5.25 pF per inch instead of 7.389 pF per inch. I suspect the difference is due to the dielectric constant of the Teflon used for the insulation (see Note 2).

Voltage-Handling Capabilities

If you're concerned about the voltage-handling capabilities of a capacitor with a spacing of only $^1/_{16}$ inch between the inner and outer plates, don't be. According to the *Handbook*,[4] the puncture voltage of Teflon insulation is 1 to 2 kV per mil (0.001 inch). This capacitor's voltage rating is *well* within limits for a typical 100-W transmitter.

Tuning Mechanism

Capacitor tuning is accomplished by turning the threaded brass rod with a low-speed gear-head motor. A short piece of plastic tubing connects the rod to the motor. The tubing acts as an insulator and a flexible coupling to smooth out minor shaft-alignment errors. The other end of the rod is threaded into a brass nut soldered to the crossbar holding the $^1/_2$-inch pipes together. I used a 12-V motor rated at 180 rpm, but it has sufficient torque to work with as little as 4 V applied.

Instead of a sophisticated variable duty-cycle speed control circuit, I used an LM317 adjustable voltage regulator to vary the motor-control voltage from 4 to 12 V. Tuning speeds range from 11 seconds per inch at 12 V to 40 seconds per inch at 4 V. The higher speed is necessary to jump from band to band in a reasonable amount of time. The lower speed makes it easy to fine-tune the capacitor to any desired frequency within a particular band.

Construction Pointers

I've intentionally omitted step-by-step construction details from this article because of differences in drive motors, tubing sizes and available materials. However, by studying the photos and drawings, you should have no problem building your own capacitor. Table 1 contains a list of the parts I used in constructing my capacitors.

When building the capacitor, keep in mind that the smaller tubes must telescope in and out of the larger tubes with silky smoothness. Any binding will cause erratic tuning. For the same reason, the #8-32 brass threaded rod must be straight and properly aligned with the brass nut. *Take your time with this part of the project.*

If you can't locate a source of Teflon sheet for the insulation, use clear polyethylene. Your capacitor will, however, have a *slightly* higher overall capacitance because of the difference in the polyethylene's dielectric constant compared to that of Teflon. You may

Table 1
KD7S Loop-Tuning Capacitor Parts List for a 50-pF Capacitor

Qty	*Description*
2	10-inch length of $^3/_4$-inch-ID type M copper pipe
2	10-inch length of $^1/_2$-inch-ID type M copper pipe
1	3-inch length of $^1/_2$-inch-ID type M copper pipe
2	$^1/_2$-inch, 90° copper elbows
2	$^3/_4$-inch, 90° copper elbows
2	10×22-inch pieces of 0.005-inch-thick Teflon or polyethylene sheet plastic
1	12-inch length of #8-32 threaded brass rod
1	#8-32 brass shoulder nut
2	22×5$^1/_2$×$^1/_4$-inch ABS plastic sheet (top and bottom covers)
3	1×5$^1/_2$-×$^1/_4$-inch ABS plastic sheet (end pieces and center brace/guide
2	1×22-×$^1/_4$-inch ABS plastic sheet (side rails)
1	50 to 200-rpm gear-head dc motor
1	DPDT center-off toggle switch (up/down control)
2	SPDT microswitches (limit switches)
50 ft	3-conductor antenna rotator control cable
1	Enclosure (switch housing)

Misc: 4 to 12-V, 400-mA variable-output power supply, motor-mounting hardware, rubber grommets, hook-up wire, adhesive.

also find polyethylene is a little more difficult to work with than Teflon. The high-voltage-handling capabilities are about the same for either plastic.

Perhaps the easiest way to form the insulation is to pre-cut a length of plastic sheet the proper size. Place a lengthwise strip of double-sided tape on the tube to secure one end of the plastic. Begin wrapping the plastic around the tube while keeping it is as tight as possible. *Don't allow any wrinkles or ridges to form.* Secure the other end of the plastic with another piece of tape. Once both tubes are covered, ensure they are just short of being a snug fit inside the larger tubes. Confirm that the insulation completely overlaps the open end of the small tubes. If not, the capacitor is certain to arc internally with more than a few watts applied to it. To protect the insulation, you can brush on a light coat of epoxy resin. Make sure the added thickness of epoxy doesn't interfere with the movement of the tubing.

Route the drive motor wiring *inside* the antenna pipes to minimize the amount of metal within the field of the antenna. Bring the wires out of the bottom next to the coaxial connector. As shown in Figure 3, a 3-wire system allows the use of limit switches to restrict the movement of the "rotors."

Be sure to solder together all metal parts of the capacitor. Use a small propane torch, a good quality flux and 50/50 solid solder. *Never use acid-core solder on an electrical project!* Clean all parts to be joined with steel wool prior to coating them with flux. Heat the joints to the point where the solder flows on its own. Very little solder is needed to make a strong, low-resistance joint. Be extremely safety conscious while soldering. No one wants to play host to the fire department or paramedic team when you could be working DX on 20 meters instead.

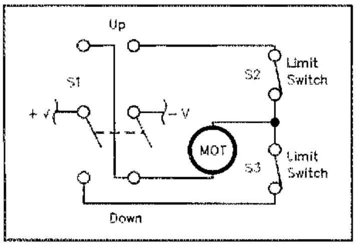


Figure 3—Schematic of the capacitor's tuning control circuit. S2 and S3 are microswitches used to limit the travel of the telescoping tubes (see Figure 1).

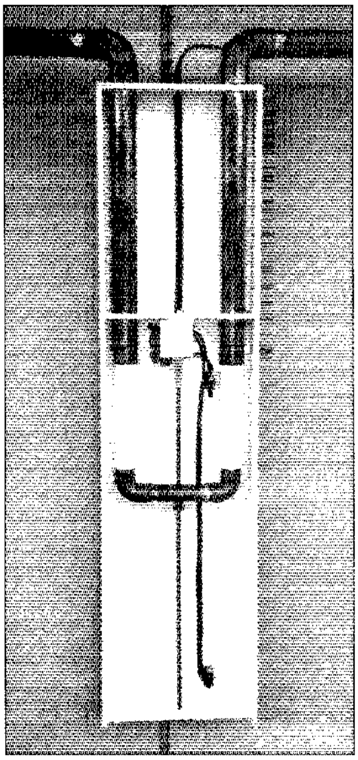


Figure 4—Close-up view of the KD7S loop-tuning capacitor and drive mechanism.

Test Results

Several of these home-brew capacitors have been tested on a 10-foot circumference (3-ft diameter) loop made of $^3/_4$-inch copper pipe. The antenna is mounted within 4 feet of the ground and has been fed with power levels ranging from 5 to 100 W. I've made contacts on 10, 15, 17 and 20 meters with this arrangement. There's been no sign of capacitor heating or breakdown. Tuning is smooth and precise. During Field Day 1994, dozens of stations were worked, coast to coast, with good signal reports. DX contacts have included stations in South America, Europe, Japan and the South Pacific. In a nutshell, the antenna and capacitor combination works quite well!

I welcome your questions, comments and suggestions on this project. I'll reply to correspondents who supply a business-size SASE.

Acknowledgment

A special thank you goes to my wife, Merleigh Jones, who put up with my numerous weekends of "radio therapy" in the garage. Without her understanding and support, this project would never have become a reality.

Notes

[1] T. Hart, "Small, High Efficiency Loop Antennas," *QST*, Jun 1986, pp 33-36.

[2] See J. Rusgrove, "The Trombone Trimmer—Build Your Own Variable Capacitors," *QST*, Nov 1975, pp 22-23 and 34. The calculated capacitance achieved using Rusgrove's equation amounts to about 2 pF per inch more than that found by using Equation 1 of this text. Note also (on pages 23 and 34 of the referent) the reasons given for a rather large discrepancy between the calculated and measured capacitance of The Trombone Trimmer.—*N1FB*

[3] *The ARRL Handbook*, 1994 edition, p 2-14.

[4] *The ARRL Handbook*, 1994 edition, p 2-13.

First licensed at age 13, Bill Jones wasn't satisfied with the restricted operating privileges his Novice ticket allowed, so he upgraded to General three months later. He's held an Extra Class license since they first became available.

High-speed CW, QRP and building his own equipment from scratch are Bill's primary Amateur Radio interests. According to Bill, his junk box looks like part of the freebie box from every hamfest ever held.

A Systems Analyst for the City of Fresno, California, Bill deals strictly with personal computers and is part of a team responsible for supporting several hundred city employees with hardware and software expertise. Mostly self-taught, he professes to be a graduate of McGraw-Hill, SAMS and TAB Universities...

Just Enough Radio—
The SP-750 Spider Junior

Fun things come in small packages...

By Mike Agsten, WA8TXT
405 W Bogart Rd
Sandusky, OH 44870
(Photos by Kirk Kleinschmidt, NT0Z)

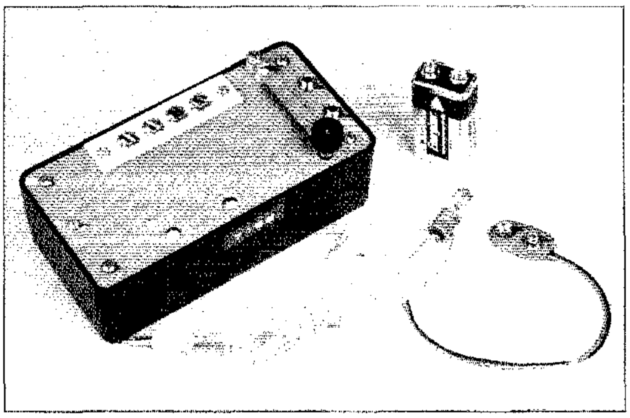
Assuming there's a typical 100-W transceiver at the other end, how much radio hardware is actually needed to provide you a fair chance at a QSO? Not much! Under favorable conditions, the rig described here can make contacts on 80 or 40 meters when powered with just a 9-V battery! Of course, you must forego luxuries such as SSB and AM, all-band operation and a digital read-out VFO or frequency synthesizer. *This is radio in the raw!*

This minimal-hardware approach has the benefits of low cost, easy construction and compactness. The SP-750 Spider Junior[1,2,3] is a 1-W, crystal-controlled CW transceiver with sidetone, RIT, audio bandpass filter and QSK TR switching—all in a 1⅝×5×2⅞-inch (HWD) box. To minimize vulnerability to backpacking abuse (and keep down the cost), RIT and volume controls are recessed, screwdriver adjustments, There's no on/off switch—the rig is on whenever power is connected, and it transmits whenever the key (built-in or external) is closed. Jacks for an external key (or keyer) and low-impedance earphone are included, as is a 4-lug terminal board for antenna, and dc power connections.

Nominal power requirements are 13.8 V dc at 25 mA when receiving, and 200 mA during transmit, but the supply voltage can range from 14 V all the way down to 7 V with little sacrifice except for power output. RF output falls to 750 mW at 12 V. (See the sidebar "Powering the Spider Junior.") A fresh 9-V alkaline battery should yield 250 mW output with a 16-mA current drain during receive and 95 mA on transmit. Battery endurance of an hour or more is possible (if you're not too long-winded!) and your signal will be far better than Tommy Rockford's grid-dip oscillator in *SOS at Midnight*![4]

Circuit Description

The Junior is basically a two-stage crystal oscillator/power amplifier transmitter (Q1-Q2) with a direct-conversion receiver (U2) and an audio amplifier (U1) tacked on. Direct-conversion means the receive signal is converted straight to audio without any intermediate signal processing, Performance is sacrificed for simplicity.



Needing only an operator and antenna, this 40-meter package is ready to go: transceiver, earphone, battery and clip and a no. 47 bulb dummy antenna.

U2—the MC3359—is usually seen in FM scanner receivers, the application for which it was designed. I chose it for a number of reasons. It has a double balanced mixer for converter use, an op amp for audio bandpass filtering and a shunt switch for key-down audio muting—just what Junior needs—all under one roof! The clincher was availability: I already had some MC3359s in my junk box and it's still listed in supplier catalogs, costing little more than the beloved (but nearly extinct) 40673 dual-gate MOSFET.

Construction Notes

Consult the parts list, schematic diagram and your junk box, then make up your shopping list. Don't forget any supplies you may need if you decide to make the PC board (see Note 2). Parts suppliers are listed in the sidebar on page 35. If you need help getting organized, read Bruce (KB1MW/7) Hale's comprehensive tutorial on how to tackle projects like this—good advice for all.[5]

Crystals

The Junior uses miniature HC-18 or HC-49 solder-in crystals. There's sufficient room on the PC board to accept a socket for HC-25 plug-in crystals. If you want to use the socket, you need to enlarge the two outer holes in the Y1 area of the PC board to accommodate it. You can use FT-243 or HC-17 crystals too, but they won't fit on the PC board. I recommend that you use a larger enclosure and run short leads from the existing Y1 PC-board holes to an off-board crystal socket. If you enjoy operating on the 80-meter band, you're in luck. Surplus catalogs abound with HC-18 crystals for 3579 and 3686 kHz. Put your Junior on 80 meters and join the frugal Color Burst Gang or the Novice 86ers! Ocean State Electronics and JAN crystals (among others) sell crystals for more commonly used frequencies.

Assembly

When installing the parts, observe proper orientation of the diode bands, transistor

Table 1
Band-Sensitive Part Values

Component	*80 m*	*40 m*
C1	390 pF	68 pF
C2	18 pF	5 pF
C3	680 pF	390 pF
C4	820 pF	680 pF
C11	820 pF	390 pF
C12	820 pF	820 pF
C13	680 pF	Not used
C14	820 pF	390 pF
C15	39 pF	10 pF
C16*	Not used	Not used
C17	0.001 µF	0.001 µF
C25	390 pF	68 pF
L1	40 t #30	23 t #28 (on FT-37-61 core)
L2, 3	22 t #24	14 t #24 (on T-50-2 core)

*This capacitor location is for future experimentation and not presently used.

Capacitors are 50-V (or more) disc-ceramic units. L1-L3 are wound with the appropriate size enameled wire.

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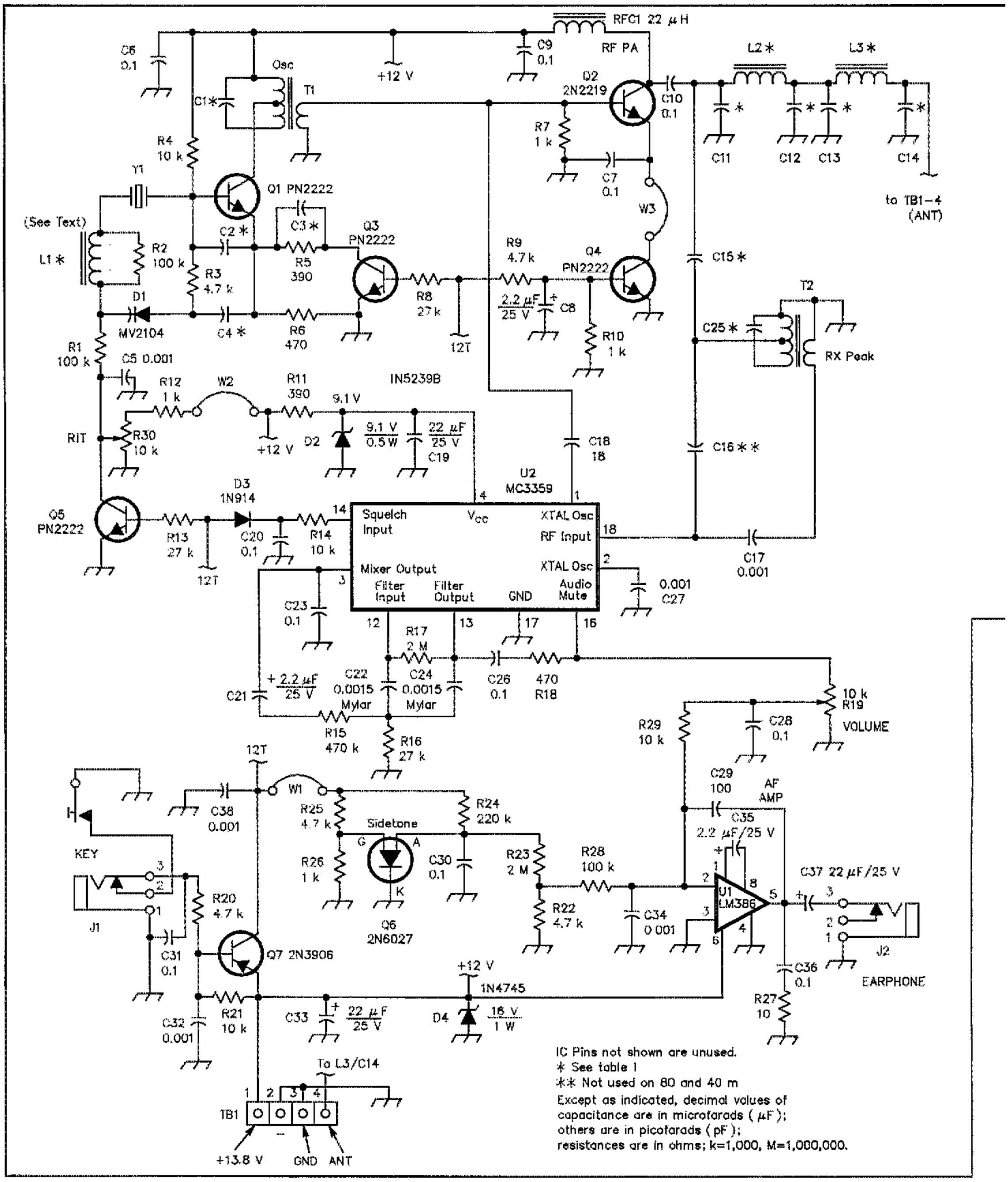
[1] Notes appear on page 36.



IC Pins not shown are unused.
* See table 1
** Not used on 80 and 40 m
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.

Figure 1—Schematic of the SP-750 Jr. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. Disc-ceramic capacitors are 50-V units.

flats, etc, and you should have no trouble. Begin assembly by installing the board-hugging components such as the resistors and diodes. Many resistors—and some diodes—must be installed vertically because of close component spacing. Don't overlook wire jumpers W1 through W3. W2 runs from just below T1 to R12 near the **RIT** control. Install the **RIT** (R30) and **VOL** (R19) potentiometers on the *bottom* (foil side) of the PC board. Q2 needs no heat sink for normal CW duty cycles.

Only three common drill sizes are needed for the holes, unless the parts you select vary from the norm. Alter your metal work to fit the parts you have. TB1's cutout can be made

C1-C4, C11-C18, C25—See Table 1. Note: C16 is not used.
D1—Varactor tuning diode, MV2104, ECG612, 1S2687.
D2—9.1-V, ½-W Zener diode (1N5239B or equiv.).
D4—16-V, 1-W Zener diode (1N4745 or equiv.).
J1, J2—3.5-mm phone jack (Mouser 16PJ137).
L1-L3—See Table 1.
Q1, Q3, Q4, Q5—PN2222, ECG123A.
Q2—2N2219A.
Q6—2N6027, ECG6402.
Q7—2N3906, PN2907 or ECG159.
R19, R30—10-kΩ horizontal-mount PC potentiometer (Radio Shack no. 271-282).
RFC1—22 µH , epoxy coated (Mouser 43LS275).
T1, T2—RF transformer (Mouser 42IF123).
U1—LM386 (Circuit Specialists).
U2—MC3359P (Circuit Specialists).
Y1—Fundamental crystal, 32-pF load, HC-18 or HC-49 holder (Petersen type Z-13-P). Note: The chosen crystal frequency should be 1 kHz above the desired operating frequency.

Misc: PC board (see Note 2); #24, #28, #30 enameled wire; enclosure (Radio Shack no. 270-233 used here); 4-position terminal strip; brass strip (for case-mounted key); hardware.

Powering the Spider Junior

In your shack, where size and weight do not restrict, your power supply can range (in order of increasing endurance) from a battery of 8 or 9 AA, C, or D cells—or a pair of 6-V lantern batteries—to an ac-operated, electronically regulated, 13.8-V supply. Shoot for a 12 to 14-V dc supply with a current rating of 200 mA or more. To interconnect the power supply and rig, dc power cords (such as the Radio Shack #270-025, or equivalent) provide quick-disconnect, polarized connectors and an in-line fuse to minimize the risk of applying reverse polarity to the rig. If you want to test your mettle against the band at reduced output, a 9-V alkaline battery will do the trick.

For backpackers seeking minimal size and weight, a good compromise is to install 9 cordless-phone, NiCd cells in the case bottom. They'll provide 10.8 V dc with a capacity of 280 mAh and about ½ W of RF output. Sked a sharp operator back home to improve your "bit error rate!" Cordless phone cells are typically purchased in battery packs of three. They must be reworked to fit the case. The only difficulty I encountered was soldering together the rearranged cells. Apparently, the end straps (of the original pack) are specially plated for soldering, but the middle straps aren't; they're spot welded to the cells. I ended up reconnecting the cells with #22 solid wire, making a few tight wraps around each cell strap and then soldering the wire to hold it physically tight. (We need a Hints & Kinks item from someone on how to condition these "unsolderable" straps to take solder.) Try to minimize the height of the battery pack. Clearance in the case bottom is adequate, but not abundant. For ruggedness, I glued the cells permanently into the case bottom. When this pack eventually fails, I'll move the transceiver assembly to a new case bottom and battery pack rather than attempt salvage.

Excellent versatility is possible with an internal battery pack; just make the following inside changes. Remove the wire between TB1 lugs 2 and 3. Transfer the lug 2 ground wire from TB1-2 to TB1-3. Now, only lug 3 of TB1 is grounded and lug 2 is free. Connect the negative side of the internal pack to lug 3 and the positive to lug 2. Reinstall the transceiver in the case.

To charge the internal battery, connect the charging source to TB1-3 (–) and TB1-2 (+). Be sure to include a current-limiting resistor in the positive lead. Adjust the resistor's value to obtain 25 to 50 mA of charge current. I use a 47-Ω, 1-W resistor when recharging from the shack's 13.8-V workbench supply. Let's see... is there enough room on the Junior case for a solar panel?

To operate from the internal battery, simply install a jumper between TB1-1 and TB1-2. This jumper acts like an on-off switch: The rig is on when the jumper is installed. For operation with an external power source, connect to TB1-3 (–) and TB1-1 (+). In all cases, the antenna system connects to TB1-3 (ground or coax shield) and TB1-4 (coax center or antenna). As you can see, there will be two connections to TB1-3 when an external power source is used: dc (–) and antenna system ground.—*Mike Agsten, WA8TXT*

Part Suppliers List

All Electronics Corp.
PO Box 567
Van Nuys, CA 91408
Tel: 800-826-5432; 818-997-1806
Fax: 818-781-2653

Circuit Specialists, Inc
PO Box 3047
Scottsdale, AZ 85271-3047
Tel: 800-528-1417; 602-464-2485
Fax: 602-464-5824

JAN Crystals
2341 Crystal Dr
PO Box 06017
Fort Myers, FL 33906-6017
Tel: 800-526-9825; 813-936-2397
Fax: 8130-936-3750

Mouser Electronics
2401 Highway 287 N
Mansfield, TX 76063-4827
Tel: 800-346-6873; 817-483-4422
Fax: 817-483-0931

Ocean State Electronics
PO Box 1458
6 Industrial Dr
Westerly, RI 02891
Tel: 800-866-6626; 401-596-3080;
Fax 401-596-3590.

Petersen Radio Company
(crystals only)
2735 Avenue A
Council Bluffs, IA 51501
Tel: 712-323-7539

with a nibbling tool, or by hacksawing the ends and lightly chiseling the bottom so that it fatigues and breaks when you work it up and down. As a last resort—for this or any other difficult opening—use the "multitude of holes" method. Drill a series of small holes inside the perimeter of the desired opening, then enlarge them until they join and the unwanted piece falls out. File smooth the jagged edges and deburr the rough holes prior to light sanding and painting the panel to suit your taste. (I've had good luck with Wal-Mart "Fashion Satin" spray paint.)

A ¼-inch-wide, 2¼-inch-long strip of 0.032-inch-thick brass mounted on quarter-inch spacers makes up the arm of the built-in telegraph key. The key knob is a rubber equipment foot epoxied to the threads and hex nut of the upper-contact machine screw. The lower contact, mounted in the top panel beneath the arm , is a ½-inch-long #4-40 brass machine screw with its head filed flat. This screw must be insulated from the metal top panel, so use a shoulder washer above and fiber washer below, followed by flat metal washer, solder lug and hex nut to complete the assembly. Bend or shim the key arm for best feel. You may need this key someday, so I recommend installing it. (At least you won't forget to pack it...)

The PC-board mounting posts are ¾-inch-long #4-40 machine screws held by hex nuts. Install ¼-inch metal spacers, then swing the board back and down onto the posts. Secure the board with hex nuts. Dress interconnecting leads from the PC board to the top panel via the corner notch left of C30; other routes may interfere with installing the enclosure's bottom. Congratulations! You've built an SP-750 Spider Junior!

Tune-Up and Operation

The earphone can be any low-impedance type with a 3.5-mm plug. Temporarily substitute a wattmeter and dummy antenna for the antenna system and peak T1 for maximum RF output. Then, with your antenna connected, adjust T2 for best signal reception. If your test equipment cupboard is bare, you can substitute a no. 47 (6.3 V, 150 mA) lamp for the wattmeter. Solder short wire leads to its base and connect it to TB1 terminals 3 and 4. A 9-V battery can be used instead of a heftier station power supply; just

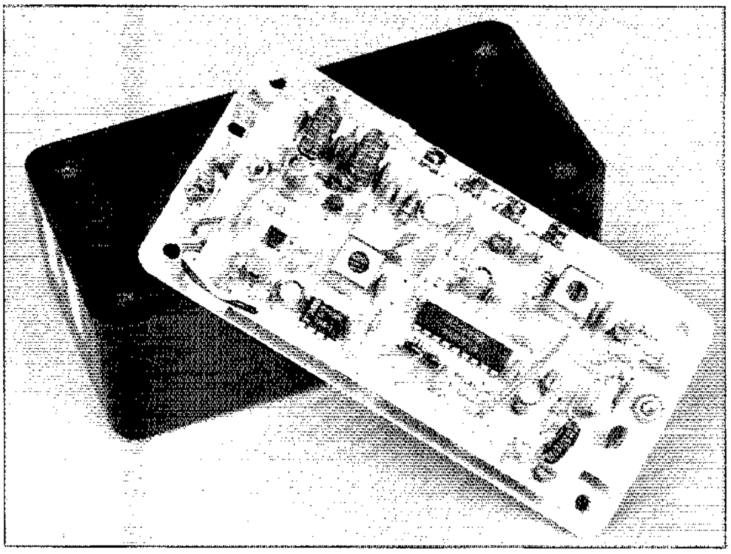


Figure 2—This completed transceiver is mounted on the enclosure cover. Simply turn it over, insert four screws and the unit is finished.

be certain you've got the power-supply polarity correct, especially if you opt to dispense with the fuse. Close the key and peak T1 for maximum lamp brilliance, which should be about one-third normal (at 9 V).

Received signals that zero-beat (or go to a very low pitch) as you rotate the **RIT** counterclockwise, are close to your crystal frequency. Other signals may be too far up or down the band to hear your signal unless the operator is tuning his receiver for replies. As a friend once told me: "Any rig that can light a no. 47 bulb can make a QSO," so keep trying!

Summary

When I hit the key on my big-bore 160-meter rig, and lights dim as the plate transformer groans, needles torque and tube plates sweat blood, it's not surprising that the RF is heard halfway 'round the world. Getting a *QRZ?* from Minnesota while using a dinky 9-V battery—now that's a *real* challenge!

Notes

[1] The Spider Junior is a direct descendant of the SP-1 Spider; see M. Agsten, "The SP-1 Transceiver," *73 Amateur Radio Today*, Jan 1993, pp 24-30.

[2] A kit for the SP-750 Spider Junior is available from Lectrokit, 401 W Bogart Rd, Sandusky, OH 44870 (no telephone). The kit includes a detailed, step-by-step assembly manual, PC board, all parts (except for the crystal and plastic case). The case parts, mini telegraph key and earphone are included. SP-750QK price: $34.50 each plus $4 shipping and handling in the US. Price and availability are valid within six months of publication; please inquire thereafter.

[3] A PC-board and mechanical layout template package is available from the ARRL. Address your request for the AGSTEN SP-750 JR TEMPLATE to the Technical Department Secretary; please include a business-size SASE with 52 cents (First Class) postage affixed.

[4] W. A. Tompkins, *SOS At Midnight*, (Newington: ARRL); available from the ARRL Bookshelf, no. 5005.

[5] B. Hale, "Build It Yourself from *QST—Part 1*," *QST*, Apr 1992, pp 31-36; *—Part 2*, May 1992, pp 35-39; *—Part 3*, Jun 1992, pp 42-45;*—Part 4*, Jul 1992, pp 31-34.

New Books

EASY TARGET

By Cynthia Wall, KA7ITT

Published by the ARRL, 225 Main St, Newington, CT 06111-1494; tel 203-666-1541. First edition, 1994, softcover, 8½×5½ inches, 168 pp, $5.95.

Reviewed By Si Dunn, K5JRN
1916 Parkside Dr
Denton, TX 76201

As a child, I loved adventure stories, especially ones where radio or Morse code sent by flashlight played a pivotal role in saving lives or capturing the bad guys.

I remember tales where cars plunged into canyons. The injured passenger would assemble a spark-gap transmitter from ignition pieces and send SOS by tapping two wires together. In variations of that plot, a light plane would crash in dense woods and the young passenger would save the critically injured pilot by patching up the radio and radiating a weak "Mayday!"

None of it was sophisticated fiction, just life triumphing over death or good triumphing over evil with a little help from RF. But reading those stories at an impressionable age helped shape my values, my commitment to Amateur Radio and my willingness to volunteer for public service.

Cynthia Wall, KA7ITT, has continued this action-adventure tradition with her Amateur Radio fiction series published by the ARRL. But as her newest novel, *Easy Target*, engagingly demonstrates, her works are a solid cut above those simplistic tales of the 1940s and '50s. Her key characters, Kim, KA7SJP, and Marc, KA7ITR, must deal with life problems, including their own strained relationship, and current social and environmental issues. When danger suddenly confronts them, Amateur Radio doesn't become their *deus ex machina*; it becomes their electronic Swiss Army knife. They still have to use their wits and skills to survive.

In *Easy Target*, Kim is now a resident student at the Oregon State University Marine Science Center. Under the guidance of a leading whale expert, she becomes involved in figuring out why some of the giant mammals are being killed by rifle bullets during their Pacific migratory journey. When Kim helps a local radio club run a transmitter hunt, she accidentally witnesses a drug smuggling operation. The men carrying it out see her and Marc and recognize Kim from a previous encounter. The smugglers track her down and start closing in, convinced Kim has stolen their shipment. That's when they get some unexpected and clever surprises by way of Amateur Radio.

Some people think Cynthia Wall's four books are written specifically for young teenagers—and the cover art on two of the books could give that impression. I've seen kids as young as eight devour these works and I've watched adults eagerly snap them up at hamfests, not only as gifts, but for their own reading pleasure.

I'm 50, my wife says I'm an old grump and I've reviewed "serious, literary" fiction for several newspapers and magazines over the years. I enjoy the heck out of Cynthia Wall's novels.

Easy Target is 60 dB over S9 seaside adventure. It has serious social and environmental issues to ponder. The portrayals of Amateur Radio are positive and realistic. Most of all, *Easy Target* is smooth, solid fiction that springs from the interactions of characters as they wrestle with their own limitations, weaknesses, hopes, dreams and fears. The fact that some of them carry hand-held transceivers or operate HF just adds to enjoyment of the plot and the writing.

If you haven't yet read any of Kim's and Marc's adventures, you can start with *Easy Target*, then move back to *Night Signals*, *Hostage in the Woods* and *Firewatch!* Or start with *Night Signals* and work forward. No matter which way you approach this series, you can enjoy some good stories and gain fresh appreciation for the Amateur Radio service and its roles in emergency communication.

A Single-Board Superhet QRP Transceiver for 40 or 30 Meters

Superhet performance in a small, inexpensive package—just the thing for traveling or low-profile operating.

By Dave Benson, NN1G
80 E Robbins Ave
Newington, CT 06111

I suspect it's been a while since most of us have truly understood what goes on behind the front panel of our transceivers. Perhaps the dazzling array of features at our disposal mutes the recognition of elegant simplicity as a design philosophy. With all this convenience at hand, why would anyone wish to build his own gear? Because there's a certain satisfaction to being able to say "I did it myself." You can't get that satisfaction in a box!

This article describes the "40-40," a single-board superhet transceiver project designed with ease of construction in mind. This design was initially developed under the auspices of the New England QRP (NE-QRP) club for use as a club project (see the sidebar for details). Far from being a "one-of-a-kind" special, nearly 200 of these transceivers have been built to date! Members of the Boston Amateur Radio Club have also built about a dozen of these transceivers.

NE-QRP emphasizes homebrewing to encourage member participation. We recognize that this isn't everyone's cup of tea, though, and sought easily reproduceable designs. We insisted that the product be as simple as possible (consistent with good performance), be modestly priced, and offer the builder a high probability of success. This project's nickname is symbolic of its original band coverage (40 meters) and cost to members ($40).

With these goals in mind, the club settled on a "bare-bones" superhet transceiver board. When you get right down to it, a workable superhet design needn't be more complex than a good direct-conversion (D-C) transceiver. This approach pays off in relative freedom from received QRM and a stable transmitted signal. We did cut several corners to hold cost and complexity to a minimum:

- We deleted the receiver's intermediate-frequency (IF) amplifier to save current drain and shave the parts count. As a result, standby (quiet-signal) current for this transceiver is only 21 mA! This approach has been described by DeMaw previously[1] and reflects our club's interest in solar/battery operation. Consequently, this design is intended for headphone use only. It provides more-than-adequate audio levels with a good set of low-impedance phones.
- We left out RIT (receiver incremental tuning) for much the same reason. The result of this minimalist philosophy is a radio with only two controls: gain and tuning! The design lends itself to adding RIT as an outboard option.

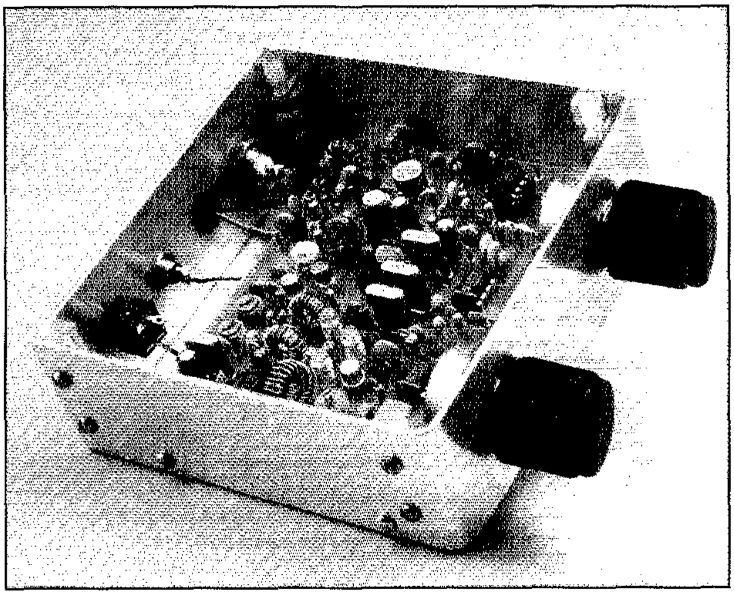


Figure 1—Mounted in an enclosure, the 30 and 40-meter transceivers make handy portable rigs.

The design efforts resulted in a single printed-circuit board measuring only 2.8 × 4 inches lending itself readily to packaging in any of a number of commercial enclosures (Figure 1). Packaged as a complete board kit, the offering included the PC-board, all on-board parts and wire for toroid winding, and a comprehensive instruction manual.[2] Builders provided the enclosure, assembly labor, and controls and connectors. These last items are all available from Radio Shack, so rounding up the necessary parts to complete the project needn't be an exercise in frustration! Let's take the 25-cent tour through the design:

So How's It Work?

Refer to Figure 1 for the following discussion. Although this text describes the 40-meter version, the 30-meter version is identical except for the choice of IF (8.0 MHz) and local oscillator (2.1 MHz). The receiver RF input is applied to U1 through T1, C1, and C2, which provide a bandpass filter tuned to the band in use. This circuitry attenuates out-of-band sig-

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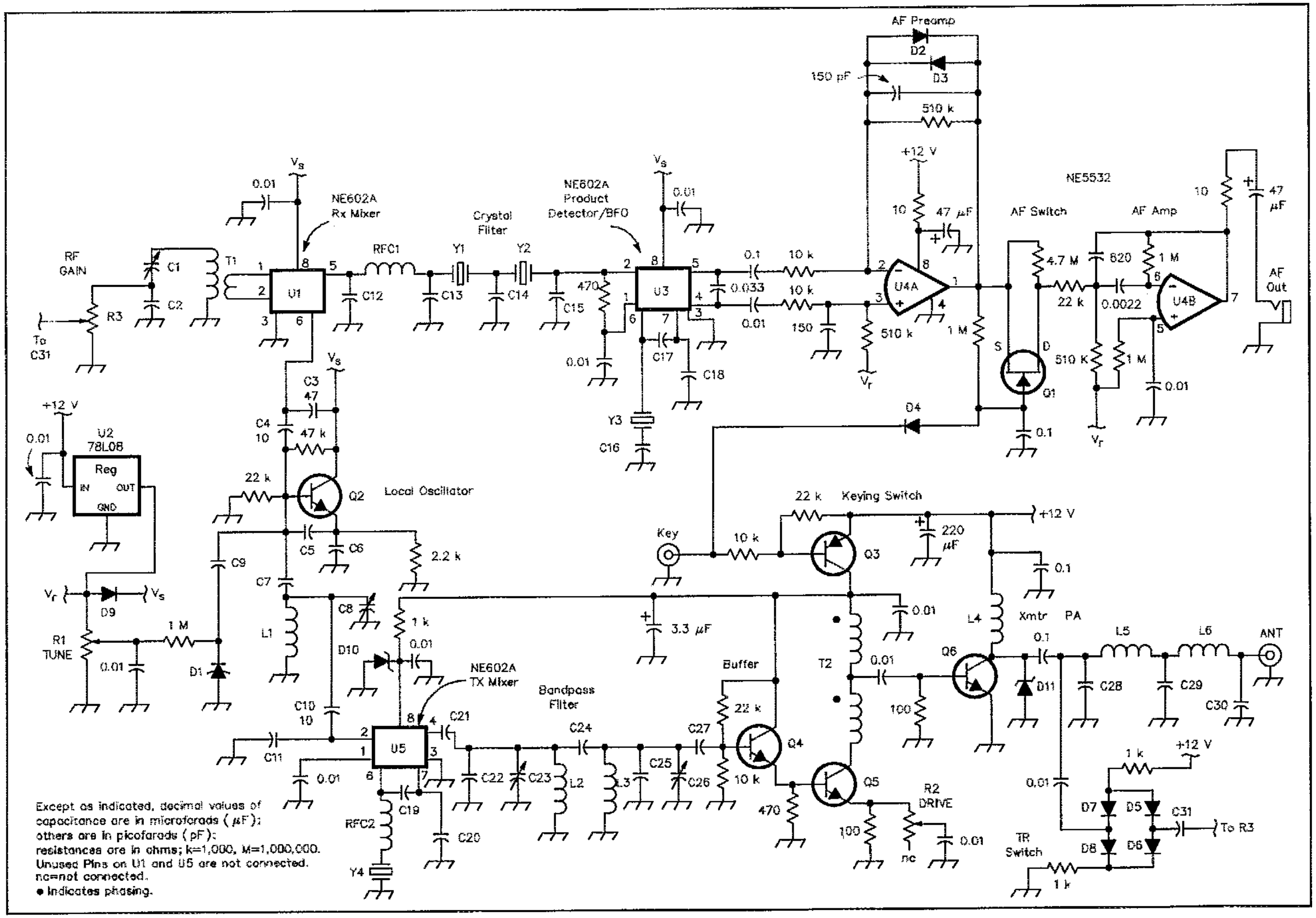
[1] Notes appear on page 41.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000. Unused Pins on U1 and U5 are not connected. nc=not connected.

● Indicates phasing.

Figure 2—Schematic diagram of the transceivers. Table 1 is the parts list. See the text for additional details. Etched circuit boards and kits of board-mounted parts are available. See Note 2.

Anatomy Of A Club Homebrew Project

A well-organized project can attract new members to your club, and entice old members to return to meetings. Here are some suggestions to help other clubs repeat the success of the 40-40:

• Identify Your Goals

Your organization probably has a central theme, whether it be Public Service, DXing or any of the many specialties that draw us together. By focusing on your needs and interests, your group can identify that perfect building project to serve as a focal point for boosting club camaraderie.

• Choose Wisely

A review of the literature will turn up candidate kit projects suited to your group's needs. The project's complexity, of course, hinges on the skill level of the membership and the availability of technical guidance. For the club lucky enough to boast an experienced builder, you might even consider "rolling your own." This will be especially true of the specialty groups (eg, ATV, VHF) who may be more construction-oriented to begin with.

• Seek the Guru

If you're going to develop your own project, you need a "Guru"! This doesn't imply a doctorate in engineering with a labful of gear, but your candidate should have a track record of successful homebrew work. This person understands the pitfalls inherent in designing an item that can be duplicated, not just once, but dozens of times! These folks aren't as rare as you might think—you might be surprised at the depth of experience among your club's ranks!

• The Project Leader

It's imperative to establish a key player to "pull it together." Unless your guru enjoys bringing an idea to production, this is a function distinct from the technical role. The project leader establishes and maintains a schedule, orders and receives parts, and enlists the aid of other necessary helpers.

A few words about scheduling: Much as we all dislike deadlines, once you've advertised a schedule for getting a project on the street, stick with it! The project leader needs to identify milestones early-on for such elements as completing tests, ordering parts and lining up help. Remember that Murphy doesn't take vacations, so allow extra time for parts orders to arrive!

• "Beta" Testing

Whether the project is a commercial kit or your own handiwork, you want to be assured that the kit goes together smoothly. You need to discover schematic errors or parts kitting shortages *before* your entire group warms up their soldering irons! You need one or more volunteers who build up the project solely from the furnished kit materials and instructions. If you have a "Guru," leave him out of this! He knows too much, and you're better off having people more representative of the average club member. This assures that the designer didn't overlook anything or leave crucial information out of the document package.

• The "Parts Team"

If you're purchasing a kit, this step is easy—send money to your chosen supplier and he sends you pre-packaged kits in return. You may expect the occasional missing, defective, or "I plugged it in backwards" parts problems to occur. If you're ordering in quantity, consider buying a spare from which to cannibalize parts. This avoids the inevitable delays involved in calling the supplier and waiting for replacements to arrive.

If it's your club's own brainchild, though, you've got a little more work ahead! The project leader needs several volunteers who can bag parts, reproduce and staple instruction sheets, etc. Once all the parts are in hand, the "kitting party" becomes part work and part social occasion. If your club is far-flung geographically, you may be able to find several willing members within reasonable driving distance to serve this function.

Allocate at least a half-day, preferably on a weekend, to get all the kitting done in one session. One suggestion: You don't want to be kitting parts while you're fatigued, so avoid late evenings. It takes a surprising amount of concentration to count out parts from their containers (we used cereal bowls). Interruptions should also be kept to a minimum—take the phone off the hook and turn off the H-Ts!

• Spreading the Word

Be sure to keep the membership apprised of the project's status. This helps maintain a high level of interest and anticipation in the outcome, and if snags arise, you may get additional help by keeping people informed. Your club's newsletter is the ideal vehicle for disseminating this information. The project leader should be working closely with the newsletter's editor to get the message across and to assure sufficient publicity. One other tip—when you're ready to release the project to the membership, make sure the point of contact for handling orders, collecting checks, etc is clearly identified. This should be one person, not a committee. Nothing beats confusion as a source of unnecessary work!

• Elmers

Every club has Elmers—experienced people willing to share their time and expertise with others. Make use of them! Provide your members with names of a couple of experienced hams who can answer questions and suggest troubleshooting tips if necessary. There's nothing like a "mentor" when you need assistance!

• Follow-Up

Put that new project to work! This step is critical to maintaining member interest and to ensure that the new gizmo doesn't sit unused on the back of the workbench! If appropriate to the project, your club might sponsor an operating event featuring the completed items. A "troubleshooting session" is also a good idea for local organizations. This lends members the impetus to give their work the "smoke test." And finally, follow up your club's results by publishing summaries in your newsletter. Don't forget to keep your ARRL Section Manager informed, so that your club's good name shares its rightful spot in the limelight!

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nals to minimize the risk of overloading U1. U1 converts the RF input to the IF of 4.000 MHz and provides about 13 dB of gain. The L network (C12 and RFC1) following the mixer matches that stage's output impedance to the crystal filter's design value.

The crystal filter itself uses only two crystals instead of the usual four. This works well because of the choice of a low intermediate frequency. (The 8-MHz IF used for the 30-meter version is somewhat broader, but the result is still usable.) Loss through the filter is less than 2 dB, and with the component values shown, the –6 dB bandwidth is about 500 Hz. Despite the filter's low parts count, performance is adequate when combined with the AF section's selectivity. The unwanted sideband image is down about 45 dB at the audio chain's 800-Hz peak response frequency.

The filter output is terminated in a 470-Ω resistor at the input to U3, the product-detector stage. U3 converts the 4.0-MHz IF signal to audio and contributes another 13 dB of gain. BFO crystal Y3 has been selected to match the IF filter frequency, so there's no BFO frequency trimming needed. The 0.033-µF capacitor across pins 4 and 5 of U3 provides the first measure of audio low-pass filtering.

The two sections of U4 each provide roughly 30 dB of amplification. The first section is configured as a differential amplifier to make use of U3's differential output, and rolls off the audio response

Table 1

Parts List For The Transceivers

Designation *(See Figure 2)*	*40-Meter Version*	*30-Meter Version*
C1, C23, C26	8 to 70-pF, 6-mm trimmer	8 to 70 pF, 6-mm trimmer
C2	220 pF	150 pF
C3	47 pF	47 pF
C4	10 pF	10 pF
C5 to C7	0.0022 µF, 5% Polystyrene, radial lead	0.0047 µF, 5%, Polystyrene, radial lead
C8	2 to 27 pF (Digi-Key SG3004)	2 to 27 pF (Digi-Key SG3004)
C9	68 pF	150 pF
C10	10 pF	10 pF
C11	220 pF	220 pF
C12	47 pF	22 pF
C13	150 pF, 5%	270 pF, 5%
C14	150 pF, 5%	270 pF, 5%
C15	150 pF, 5%	270 pF, 5%
C16	68 pF	none (replace with jumper)
C17	47 pF	68 pF
C18	47 pF	150 pF, 5%
C19	47 pF	47 pF
C20	150 pF	220 pF
C21	10 pF	10 pF
C22	150 pF, 5%	150 pF, 5%
C24	5 pF	5 pF
C25	150 pF, 5%	150 pF, 5%
C27	47 pF	47 pF
C28	470 pF	330 pF
C29	0.001 µF	680 pF
C30	470 pF	330 pF
C31	68 pF	47 pF
L1	3.65 µH (27 t #22 on T-50-2 toroid)	3.65 µH (27 t #22 on T-50-2 toroid)
L2, L3	2.5 µH (25 t #26 on T-37-2 toroid)	1.2 µH (20 t #26 on T-37-6 toroid)
L4	10 µH (5 t #22 on FT-37-43 toroid)	7 µH (4 t #22 on FT-37-43 toroid)
L5,L6	1.0 µH (16 t on T-37-2 toroid)	0.68 µH (15 t #26 on T-37-6 toroid)
RFC1	22-µH epoxy RF choke	10-µH epoxy RF choke
RFC2	22-µH epoxy RF ckoke	4.7-µH epoxy RF choke
T1	Primary 16 t #26; secondary 4 t on FT-37-61 toroid	Primary, 11 t #26; secondary 6 t on FT-37-61 toroid
Y1 to Y4	4.000 MHz (see text)	8.000 MHz (see text)
(Remaining parts are identical for both bands)		
D1	MV1662 Varicap diode	
D2 to D9	1N4148 or 1N914	
D10	7.5-V Zener, 0.5 W, 5% (1N5236 or equiv.)	
D11	33-V Zener, 0.5 W, 5% (1N5257 or equiv.)	
R1	100-kΩ linear-taper pot	
R2	200-Ω Cermet trimmer pot (Digi-Key 36C22)	
R3	5-kΩ linear-taper pot	
Q1	2N5486 (2N5485 or MPF102 substitutes)	
Q2, Q4, Q5	2N2222A metal	
Q3	2N3906 or equivalent PNP GP Switch	
Q6	2N3553 (MRF-237 substitute)	
T2	4 bifilar turns on FT-37-43 toroid (use 3-inch piece of two-conductor ribbon cable)	
U1, U3, U5	NE602A(N) Signetics mixer/oscillator IC	
U2	78L08 voltage regulator IC	
U4	NE5532	

Notes

Small-value capacitors are 10% tolerance ceramic disc, unless otherwise noted.
Decoupling capacitors are ±20% ceramic or monolithic types.
Electrolytic capacitors are minimum 16 V dc.
All resistors are 1/4-W, 5% tolerance.

above 1.5 kHz. Diodes D2 and D3 limit the audio swing during transmitter key-down to reasonable values. Without these diodes, this stage saturates and upsets the operation of the FET switch section that follows.

The AF mute function is the familiar series-FET switch popularized by W7EL.[3] Despite its relative simplicity, it's hard to beat this circuit for click-free audio switching. In the key-up condition the FET is zero-biased and acts like a resistance of several hundred ohms. In the key-down condition the FET is in cutoff (because the gate is now 7 to 8 V below the source) and acts like an open-circuit, preventing audio from getting to U4B, the audio final stage. The 4.7-MΩ resistor across the FET serves to "leak" a sample of the received signal during key-down, and thus provides a built-in sidetone.

The audio final stage is configured as a bandpass filter centered at 800 Hz. The high gain of the two NE5532 stages allows a design with no IF amplifier stage. The audio output level is adequate to drive headphones, but it won't cut the mustard for loudspeaker applications. The AF-output stage is internally overcurrent limited on loud signals, to provide ready-made ear protection. (In this respect, it accomplishes what you'd want an AGC for anyway.) If you're interested in saving a few milliamperes on receive, either TL072 or an MC1458 may be substituted, although at reduced maximum AF output levels. Use good quality low-impedance headphones for best results. Walkman headphones are fine, but remember—you get what you pay for. The $3 cheap imitations are distinctly inferior!

Diodes D5 through D8 and associated components provide full break-in (QSK), to eliminate the annoyance of a switching relay. This "bridge" circuit presents a low resistance to small signals (in the receive mode), yet limits the receiver input voltage swing during key-down to several volts peak-to-peak.

The local oscillator uses the familiar Colpitts configuration. Polystyrene capacitors aren't the usual candidate of choice when used alone in a VFO design, but I used them here because they're compact and available in 5%-tolerance values. The resulting temperature stability is more than adequate at 3 MHz. Coverage is about 40 kHz on 40 meters and 25 to 30 kHz on 30 meters. Both transceivers cover the parts of the subbands most used by QRPers, and allow use of a single-turn tuning pot. If you want broader frequency coverage, the value of C9 may be increased. Use NP0 or silver-mica capacitors if you experiment with this circuit. The design uses a varicap tuning diode for tuning. While a smooth ball-bearing tuning capacitor and vernier reduction drive are the preferred approach, that choice drives up the cost and mechanical complexity of a transceiver considerably.

The transmitter portion of this design consists of mixer/oscillator U5, driver stages Q4 and Q5, and the power amplifier (PA) stage, Q6. The output power is about 1.5 W and a drive control (R2) at the emitter of Q5 sets the output level. The L-C components between U5 and Q4 form a bandpass filter that serves to filter the relatively complex waveform at U5's output into a clean sine wave.

Performance Summary

Testing conducted in the ARRL Lab confirmed that this design met FCC requirements for transmitter spectral purity. All harmonics were down more than 34 dB from the carrier for the 40-meter version and 32 dB for the 30-meter version. Spurious outputs were down at least 35 dB. While additional filtering would improve the characteristics, these values meet present FCC requirements. Transmitted-waveform rise and fall times are 1 to 2 milliseconds. The receiver performance was also evaluated. The two-tone dynamic range (90 dB) and blocking dynamic range (115 dB) are fine for a receiver of this simplicity.

Construction

Table 1 lists parts for both versions of the transceiver. A few tips on constructing this project are in order. Although a printed-circuit board is available, there's nothing to prevent you from duplicating this design using "dead-bug" or "ugly" construction. Remember to keep all leads and connections short and direct if you're employing this approach.

Soldering

Use a small (25-W) iron (such as a Radio Shack 64-2070) and keep the tip clean. Use a moistened sponge or paper towel to clean the tip periodically as you work. Apply only as much heat as is needed to get a good joint. The use of a high-wattage soldering gun is not recommended, as it will tend to lift traces off the board.

Unsoldering

Sooner or later, you're going to need to remove a part installed in the wrong location, or perhaps pull a part for troubleshooting purposes. There's an easy way: get yourself a roll of desoldering braid (Radio Shack 64-2090B). Lay the end of the braid down on the joint to be cleaned and press the soldering iron tip over the braid. Within a second or so, you'll see the braid begin to wick up solder from the joint. Remove the braid and reapply a new section as needed until the joint is clean. This is good stuff—with care, it's possible to remove ICs from the board without damaging them.

Checking Your Work

Get into the habit of checking your work each time you put down the soldering iron. It's much easier and ultimately less frustrating to find solder bridges sooner rather than later.

If you're rolling your own version of this transceiver from scratch, select the crystals (Y1 through Y4) for identical frequency characteristics. The design makes uses of readily available microprocessor crystals to keep the cost low. A batch of 8 to 10 of these should yield four crystals that oscillate within 50 to 100 Hz of each other, and the most closely grouped crystals are the ones you want! A "what-have-you" test oscillator may be kluged up on a scrap of circuit-board material. Its characteristics are noncritical, as you're interested in frequency consistency rather than absolute accuracy. A frequency counter or main-station receiver is used to perform this grading "by the numbers" or "by ear," respectively. Incidentally, the choice of 4.000 MHz as the 40-meter IF yields a strong "birdy" at 7.000 MHz. If your plans include the use of the extreme low end of the band, 4.032-MHz crystals are a better choice.

Alignment

Alignment of this little rig is a piece of cake. For test equipment, you'll need a main-station rig and SWR meter or wattmeter. Here's how:

Receiver

- Double-check your power supply polarity *before* you apply power!
- Apply dc power and connect a pair of headphones to the receiver AF output.
- Set tuning pot R1 to full counterclockwise (0 V on R1's center terminal). Using your main-station rig (at minimum power into a dummy load, please!), transmit a string of dots at the frequency you desire for the low end of your band coverage. Adjust trimmer cap C8 until you hear your main-station rig's transmitted signal. You may need to remove a turn from L1 to get the tuning range up into the 40-meter band. If the band coverage is too high in frequency, squeeze the turns on L1 together to reduce the operating frequency.
- Connect a matched antenna. Using a *nonmetallic* tuning tool, peak C1 for maximum signal (or noise). If you don't have one of these tools, you can make one from a narrow strip of PC board material with the tip filed down to fit the slot in the tuning cap. The use of an insulated tool is important because body capacity will make adjustment difficult otherwise. Once this step is complete, background noise will be noticeably higher with an antenna connected than without.

Transmitter

- Set drive adjustment pot R2 to its full counterclockwise position.
- Add a clip lead to the case (collector) of Q5 to serve as a short antenna. (Leave the other end of the clip lead unconnected.) Ground the "key" input and tune in your transmitted signal on your "big rig" receiver. Adjust trimmers C23 and C26 for maximum indicated signal on your receiver's S-meter. The capacitors will be somewhat interactive, so alternate between the two as you tune for maximum output.

Remove the clip lead. Connect the RF output to a wattmeter. If you lack this luxury, you can use your SWR meter in the **FORWARD** position, connected to a dummy load. Adjust R2 as needed for 1.5 W key-down power into a dummy load or the test circuit. More is *not* better—you'll coax a little more power out at the expense of efficiency!

That's it! You're on the air. Feedback to date from other builders has been highly favorable. You might be surprised at what you can work on either 40 or 30 meters with only a watt or two. I've had the pleasure of ragchewing with stations from all over the US and have worked about a half-dozen European countries with the 40-meter version alone.

Acknowledgements

Special thanks are in order to Jim Fitton, W1FMR, for his guidance; to Paul Kranz, W1CFI; Wayne Burdick, N6KR; and others for design suggestions and other support. This rig was first described in the April 1994 issue of *72*, the newsletter of the New England QRP Club.

Notes

[1] D. DeMaw, "A Four-Stage 75-Meter SSB Superhet," *QST*, May 1989, pp 25-28.

[2] Kits of parts for the 40-40 and 30-40 are available for $50 postpaid from Small Wonder Labs, 80 East Robbins Ave, Newington, CT 06111.

Be sure to specify either the 40 or 30-meter version. Includes selected crystals, all on-board parts, wire for toroids and manual. Allow three weeks for delivery. (The ARRL and *QST* in no way warrant this offer.)

Circuit boards (only) for this project are available for $6 (plus $1.50 postage) from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118.

Dan's Small Parts and Kits (1935 S 3rd West No. 1, Missoula, MT 59801) carries many of the parts needed for this project. Remaining items may be obtained from Jameco, Mouser Electronics, and Digi-Key Corp. See the List of Suppliers in the ARRL *Handbook* for more information.

[3] R. Llewellyn, "An Optimized QRP Transceiver for 7 MHz," *QST*, Aug 1980, pp 14-19 (an updated version of this project appears in the 1994 *ARRL Handbook*).

New Products

PROGOLD CONTACT CLEANER/ ENHANCER

◊ Caig Laboratories' ProGold contact and connector cleaner/enhancer/conditioner is now available in five packages. ProGold deoxidizes and cleans surface contamination on electrical connectors and contacts, and actually penetrates plated surfaces and bonds molecularly to the base metal. This bonding action, according to the manufacturer, is unique to ProGold.

Treating contacts and connectors can reduce arcing, RFI, wear, abrasion, and intermittent connections. Use it on switch contacts, connectors, edge cards, plugs, sockets, relays—even on bare circuit boards.

ProGold is available in spray, liquid, precision dispenser, wipes and pen applicators. Contact Caig Laboratories Inc, 16744 West Bernardo Dr, San Diego, CA 92127; tel 800-CAIG-123, fax 619-451-2799.

QST

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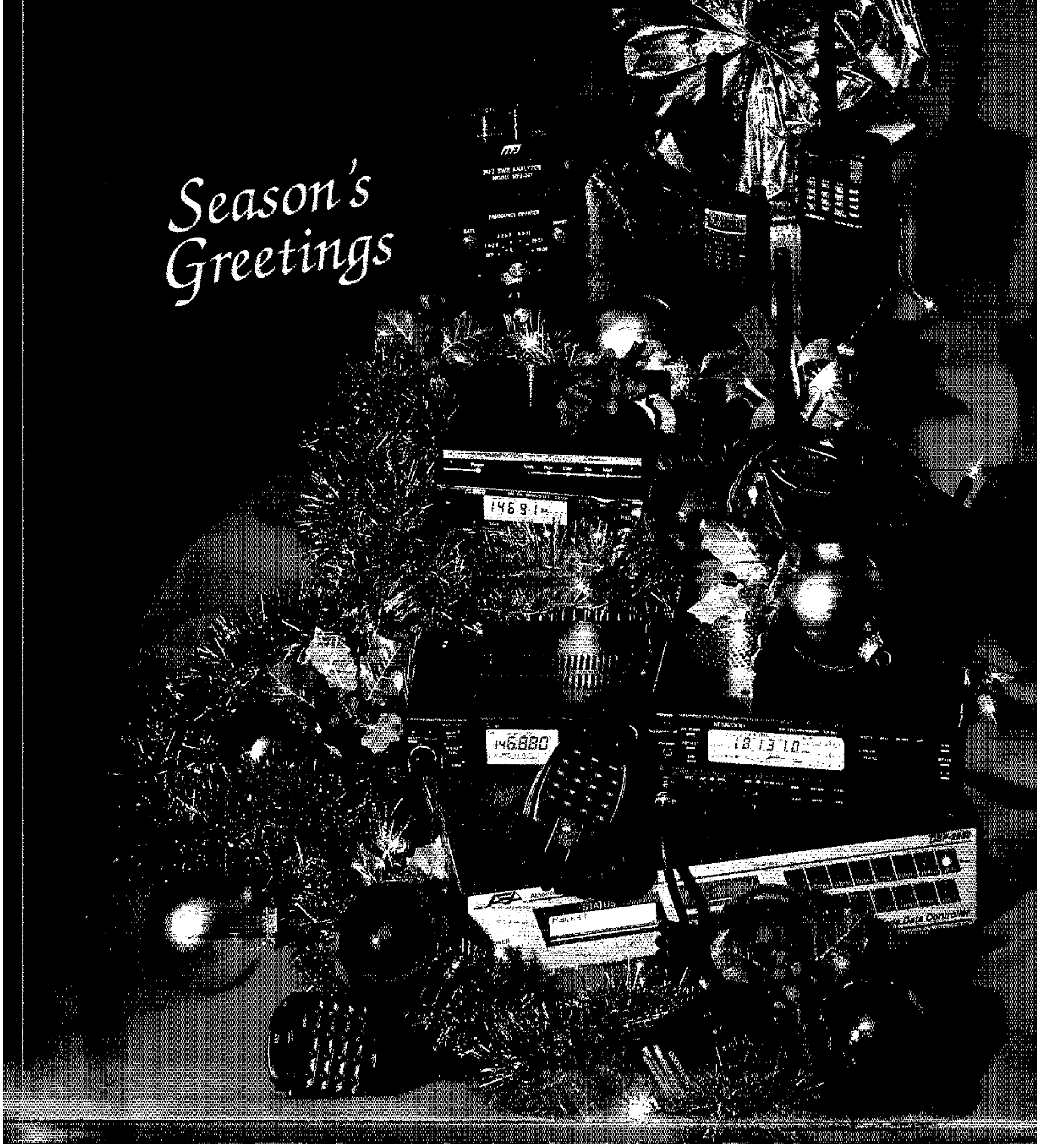

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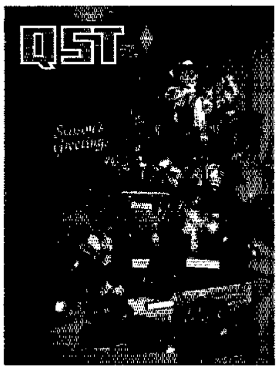

OUR COVER

Is it real, or is it a dream? This tantalizing holiday vision includes an HF radio, several 2-meter rigs and even a 6-meter transceiver (can you find it?). Plenty of accessories are in attendance to make your mouth water. Forget the gift list—just photocopy this cover and give it to your spouse, friends or whomever. Then they'll *really* know what you want!

CONTENTS

December 1994
Volume LXXVIII Number 12

TECHNICAL

NEWS AND FEATURES

NEW HAM COMPANION

OPERATING

DEPARTMENTS

SSTV

An Inexpensive System Continues to Grow

A good system gets better...

My software-based SSTV system[1] received a fairly good reception, and a few weeks after its publication there were pictures showing up on 14,230 kHz with the program's characteristic red (or green) "shades of gray" stripe at the top. Figure 1 shows some of these as copied at this location. One advantage of having a sizable body of users is the feedback you get related to problems experienced and additional features desired. A pleasant surprise is the substantial number of people who are now using SSTV on the VHF bands and have organized weekly nets to trade pictures. This move has added a new set of user priorities different from that of the old HF SSTV gang. As this is written, the software is up to revision F, each revision being a step toward filling users' requests. Some of these revisions were described,[2] but a number of them need pictorial support, so they're described here.

Program Structure

The BASIC programming built around a series of machine-language modules (which do all the high-speed work) has proved to be even more flexible than I initially realized. Its greatest virtue is that it allows many users to modify the programs to fit their own operating preferences, but it has also simplified my task of adding features.

One user-instigated feature is the menu screen shown in Figure 2. Dennis, KY1S, decided he wanted a more attractive menu and went to considerable effort to make an attractive personalized screen with icons, using some heavy BASIC programming. Believing other SSTVers would like to do the same thing—but probably wouldn't have Dennis' programming skill—I modified the original *RT* program so it accepts any user-generated SSTV picture file as a menu screen. Now you can feature your favorite landscape, girlfriend—or whatever—on the menu screen. If (or when) you get tired of it, you can change it easily. The *LABEL* program adds the menu words to the picture in a few minutes. This is but one example of a user-initiated addition that I would never have thought of! There are many more described in the oversize system manual file, *TVINFO.TXT.* Please—*read it!*

By Ben Vester, K3BC
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Ellicott City, MD 21043
Photos by the author

3-D Pictures

Many of the additions were obvious, such as adding the vertical-interval signal (VIS) code detector, and then expanding to add automatic picture saving (SAVE) so you can leave your computer unattended to detect and save the pictures received while you're away. Noting that a number of people were sending 3-D—or in many cases, *synthetic* 3-D pictures—I took a look at doing that. I was amazed to find that the machine-language module in the *LABEL* program is sufficiently general purpose that with a few judicious POKEs, it generates 3-D pictures, both the synthetic type and those composed of two pictures of the same scene displaced by eye separation. Although no one had asked for this feature, it was too easy not to include it! It's most useful when you're using a TV camera and frame grabber to generate images. You merely feed in a "right eye" and a "left eye" picture and the program outputs a 3-D image. Synthetic generation is probably too

Figure 1—Early examples of the Vester SSTV system. The top two pictures are by WB7PAP and the dog by N9EJJ.

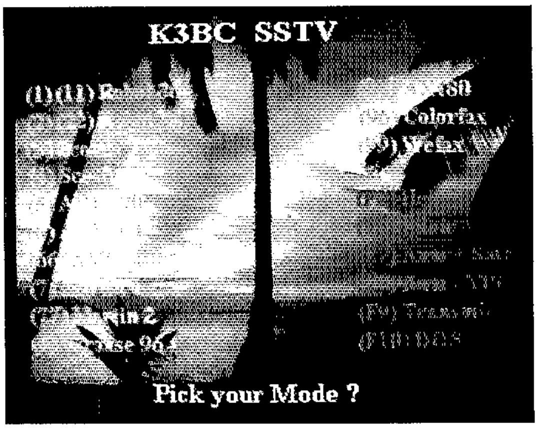

Figure 2—With the latest version of the software, you can customize the Mode menu of your copy of the software.

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[1] Notes appear on page 24.

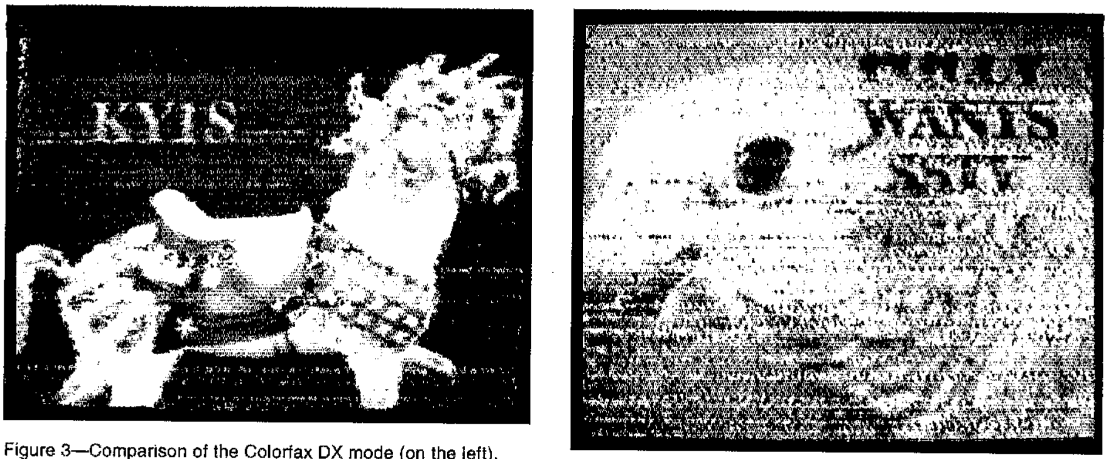

Figure 3—Comparison of the Colorfax DX mode (on the left), with a Scottie 1 mode picture (on the right) received under severe QRM (see the text).

laborious for most, but can be done with this program. Typically, you'll synthetically insert your call sign to be at the picture depth you choose.

File-Handling Convenience

The picture files are rather large, and if not compressed, soon fill up your available disk space. There are also a number of the associated programs that generate picture files that you may choose to save. Each of these has a SAVE routine added now so you can save pictures in full or compressed format right where it is generated. This is particularly convenient when you're using *SLIDESHO* or *TRUSHO* to thumb through a bunch of pictures collected while you were gone. The transmit program, *VT*, is now configured to accept files in full or compressed formats.

Alignment

Because the frequency tolerances of computer clocks vary, line-timing calibration is necessary. In revision A, released January 30, 1994, I included a simple arithmetic formula that allowed users to calculate the line timing for all of the modes once they'd experimentally determined the line timing for any single mode. Months later, there were still people on 14,230 kHz talking about the labor involved in aligning all the modes. Since some users wouldn't (or couldn't) do the simple arithmetic, I created *LINETIME* to do it for them. You crank in a corrected value for Scottie 1 mode and the program spits out all the line timings (LTs) required for the other modes.

Trucolor

More recent additions have taken advantage of the new Trucolor video cards that offer 640×480-pixel resolution with 16.7 million colors and are available for as little as $70! Such cards not only give you much brighter pictures than the 256-color boards, but the increased information content per screen allows you to display *four* SSTV pictures per screen at full resolution and color content. This not only makes for more interesting slide shows, but offers a test vehicle for side-by-side comparison of transmitted files with the original file for calibration of the transmit program.

Also, as described under **New Modes**, such cards make it possible to display higher-resolution color pictures in full color. When the normal SSTV pictures (bounded by 320×240-pixel resolution) are expanded to fill the Trucolor screen, the "square-pixel" effect becomes visible at normal viewing distances. To ameliorate this, we've used fractal techniques in the most recent versions of *TRUSHO*. Although this improves the picture quality, *it can't improve the intrinsic resolution of the picture*. This can *only be done* by using higher-resolution modes to transmit the images.

New Modes

I considered adding the Scottie DX mode, but after studying it, decided it didn't use the redundant pixels as efficiently as it might. Basically, Scottie DX sends the same pixel-byte value three times in a row and then combines these three for a more noise-free single-pixel byte. This is great for reducing white noise, which is not correlated in the three samples. But you only need look at a few pictures to see that many of the noises which pollute our pixel values (QRM, QRN, multipath propagation, carriers, etc) are often correlated over three successive pixels, or more.

So I began looking at inserting the redundancy with time separation between samples—like repeating the same line three times in a row. Each pixel byte would have three samples separated by the time of one line length. The good stability of the crystal-locked receiving system allows combining time-separated pixels quite precisely (systems using re-sync on each line would not combine so precisely!). This would decorrelate the samples against more than white noise and give better results.

The more I thought about it, the more foolish it seemed to use completely redundant pixel bytes. Four pixel bytes grouped together in a square (ie, two adjacent in line 1 and the two directly below these in line 2) could be sent with exactly the same value and give a redundancy improvement against noise in preserving that value. The same four pixel bytes could be used to improve the spatial resolution by a factor of two in each axis. If conditions are good, you see a resolution improvement of *four!* If conditions are bad, your eye averages those same pixels and gets at least as much improvement as if the pixels were all sent with the same value.

Being time separated, the two groups of two give further improvement over the Scottie DX mode. So, if you are going to send more bytes to improve poor-condition reception, it's better to use the bytes for better resolution and let your eyes do the averaging. Signal strength varies and you naturally get the better resolution when the signal strength is high. Also, because most SSTV gatherings have several stations at different distances copying the signal, some get extraordinarily good, high-resolution pictures; stations receiving weaker signals getting less. This process is illustrated daily when people switch modes from Martin 2 to Martin 1. Here, the double resolution is in one axis only, but the difference as to how fast each mode deteriorates under worsening conditions is well known.

I decided to make a full-color DX mode with double resolution in *both* axes. To avoid start-up problems getting everybody locked to the same line timing, I chose to use the same timing as Ralph (WB8DQT) Taggart's FAX480 system.[3] Three successive 546-byte "lines" are used for the red, green, and blue data bytes that form the first full picture line. This is close to a double-size Scottie 1—double in both axes. The Australian street scene in the Up Front section (page 13) shows the typical picture resolution. This double-resolution mode I call Colorfax, and it uses a VIS code of &H6A (the ampersand H

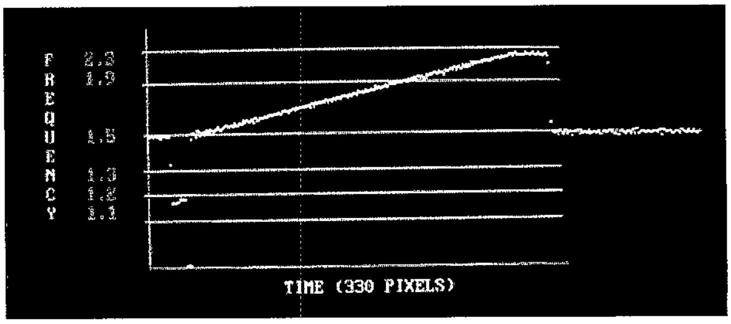

Figure 4—The frequency versus time display is used to look at incoming signals. Here you can see that the receiver is mistuned about 15 to 20 Hz from the transmitted (incoming) frequency. If perfectly tuned, the sync pulse would be located at the 1200-Hz mark instead of at 1180 Hz as shown. This frequency difference is small enough to still receive pictures well. This display is from the gray-shades portion of the incoming picture.

denotes hexadecimal notation).

For evaluating the mode, I enlisted the help of KY1S. To get a relatively constant level of interference and noise, both stations zero-beat the same 40-meter broadcast station. Then, KY1S sent me a Colorfax picture followed immediately by a Scottie 1. The resulting pictures are shown in Figure 3. It's worth mentioning that because of the strong broadcast-station signal, we had trouble communicating verbally during this test. Clearly, a mode that consumes this much time per picture (over 6 minutes) should be used only sparingly on busy frequencies such as 14,230 kHz. With the growing army of slow-scanners, the use of less-busy frequencies is becoming popular, however. And for many VHF bands, Colorfax fills the need for a really high-resolution, full-color mode. It should turn out to be an excellent DX mode.

I replaced the long acquisition header of FAX480 with a short VIS code (&HAA) to make it compatible with all the other SSTV formats. This makes both black and white and high-resolution color available with automatic VIS. The picture files for the B&W fax are interchangeable with FAX480 and can be transmitted with either the VIS or the FAX480 header. This B&W fax mode uses 64 shades of gray on both transmit and receive, unlike the 16 gray shades of FAX480. The format-conversion programs allow generation of pictures in all of these formats from *any* color image.

Format Conversion

Although the original programs were aimed at using a TV camera and frame-grabber for generating most of the picture files, I found that a number of people are using widely available GIF files for pictures. These pictures have only 256 colors, but often do a good job of matching the colors used to the picture color content. Users have located several shareware programs that convert GIF to TIFF files to plug into *TIFCONV* for SSTV conversion, but the procedure is time-consuming. Recently, KY1S located a shareware command-line format-conversion program called *Alchemy*, which is ideally suited to chaining with other programs. With one DOS line command, *Alchemy* converts images from a number of possible formats to any one of another set of formats. Using a simple batch file to chain together the *Alchemy*, *TIFCONV*, and *VT* (transmit) programs, you now can simply type the batch-file name, *TR*, followed by the path and file name for any GIF, PCX, or JPG file and—except for announcing which mode you're going to transmit—no further action on your part is required except to press the **G** key to start transmitting!

CD-ROM Pictures

After surveying the CD ROMs available at the Dayton HamVention this year, I found a number of 24-bit Trucolor pictures available in TGA format. To convert this new source of high-quality images, I created another conversion program, *TGACONV*. Using batch-file chaining of *TGACONV* with the transmit program allows *direct transmission of pictures from the CD ROM in whatever mode you choose for each picture.* Since the PATH, a portion of each file name and its suffix are common to all pictures, you need only type **TE 2905** to get the file E:\TGA\IMG2905.TGA into the transmit buffer, visible on the transmit screen and ready for **G**(o). The manual, *TVINFO.TXT*, gives you enough information to allow you to create your own minimum-keystroke batch files like this.

Real Time

One of the most-requested features was a line-by-line paint of the picture as it came in. This is easy to do with a line-synched system (which has more jitter than the crystal-locked system) and where you don't want to eliminate the picture gap, which occurs on the occasional picture that is copied out of sync. Cramming the screen paint at the end of each line just is too large a gap. Interleaving the screen paint with the measurement of each pixel byte's frequency can be done, but it requires a fast computer. Since the population of 80286 computer users seems to be decreasing rapidly, I put this in as a switchable option for the people with faster machines.

The newer Trucolor video cards don't require VESA drivers, are faster, and require a minimum of code to write to, so they work better. I also added a real-time option for 256-color boards. One negative aspect of this is I no longer manage to read my daily newspaper while pictures are coming in!

Tuning Aids

I resisted adding any frequency displays to the software for some time since the *HAMCOM* software is freely available, can use the same interface as our programs and has an excellent spectrum-analyzer display that can serve this purpose. The display I added differs in that it is a frequency versus time presentation. It displays in sequence the frequency of every byte sample received and looks much like the oscilloscope display you get looking at a TV video signal. With the same scroll routines as used in picture viewing, you can scroll the frequency-versus-time viewing window through a complete picture file. Figure 4 shows a typical shades-of-gray picture line. This display is useful for looking at VIS codes, AVT and FAX480 headers and more.

As another aid to get everyone on the same operating frequency, I added a pulsing 1200-Hz tone to the transmit program, *VT*. This tone can be keyed on when needed by holding down the T(one) key.

Comment Line

In most SSTV modes, there are 16 picture lines devoted to shades of gray, and many systems overlay your call sign here. In the latest revision, I added the ability to include a short comment in this area on each picture as it is transmitted. Or, you can default to your call sign and/or name, or whatever you choose to put in the SYSTEM CONFIGURATION list on any particular day. You can have a little fun with the Robot users, since (I'm told) that system doesn't display these lines—although the lines often do show up on the Robot's *next transmission.*

Summary

"I haven't had so much fun in years!" is by far the most common comment I've received about the system. And, I've heard at least one young man describe the system as "not user-friendly." Both comments apply. They're the natural reactions to the software's "sweat factor" that is *designed in specifically* to get some experimental juices flowing again!

The latest software revision is available on the ARRL BBS: 203-666-0578 as VESTER_F.ZIP. A number of other BBSs—and Internet—have picked it up, so you may find it on your local BBS. Let's see you on the air!

Notes

[1] B. Vester, "An Inexpensive SSTV System," *QST*, Jan 1994, pp 27-29.

[2] B. Vester, "Vester SSTV/FAX480/Fax System Upgrades," Technical Correspondence, *QST*, Jun 1994, pp 77-78.

[3] R. Taggart, "A New Standard for Amateur Radio Facsimile," *QST*, Feb 1993, pp 31-36.

QST

Exploring Intermodulation Distortion in RF Switching and Tuning Diodes

How bad are distortion-prone switching diodes, and how good are those designed for low distortion? Straightforward lab testing can provide the answers.

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A recent article by Ulrich Rohde discussed the generation of intermodulation products by the PN-junction diodes used for switching transceiver front-end filters.[1] Some of the older literature warns of this problem but only discusses it qualitatively.[2] To measure these effects and compare diode performance, I constructed the diode switch shown in Figure 1 and measured it using the setup in Figure 2.

Background

The difference between the level of the test-signal power (P_S) and the spurious response decreases as P_S is increased. (At the intercept point for an IMD product of a given order, a level that we must calculate by extrapolation, the test- and spurious-signal levels would be identical.) It follows that test equipment with a more modest dynamic range can be used to measure spurious-signal levels if P_S can be made as high as possible. But a high P_S will strain the capability of the measuring device. P_S must impinge upon the device under test, but the measuring device need see only the spurious signal, so filtering that attenuates the test signals while passing the IMD product can extend the system's measurement capability.

The spectrum analyzer I used has a third-order intermodulation-distortion intercept point (IP3) of +30 dBm, and the IMD product I wanted to measure (f_S) was always higher in frequency than my two test signals (f_1 and f_2). Adding a high-pass filter to knock down f_1 and f_2 while passing f_S increased the system's IP3 to greater than +50 dBm. Adding the filter also increased system's second-order intermodulation-product intercept point (IP2) from +50 dBm to +80 dBm.

The choice of test frequencies determines how steep the filter's slope needs to be. For IP2 measurements, I chose f_1=6.1 MHz and f_2=8 MHz, which produce a second-order spurious at f_1+f_2, or 14.1 MHz.

Normally, we measure third-order IMD by using two test signals close together in frequency. For example, using 14.2-MHz and 14.3-MHz signals, we could quantify third-order IMD by measuring the level of the resulting 14.1-MHz spurious [(2×14.2)−14.3=14.1]. Using a high-pass filter as shown in Figure 2 would not improve the test setup's dynamic range with such closely spaced test signals, however, because the filter cannot differentiate the test signals from the spurious. To avoid this problem, I did my third-order measurements with test signals of f_1=8 MHz and f_2=1.9 MHz. This still produces a third-order spurious at 14.1 MHz, but the fundamentals are much lower in frequency than the spurious and can be attenuated by the filter. (This approach is

[1]Notes appear on page 27.

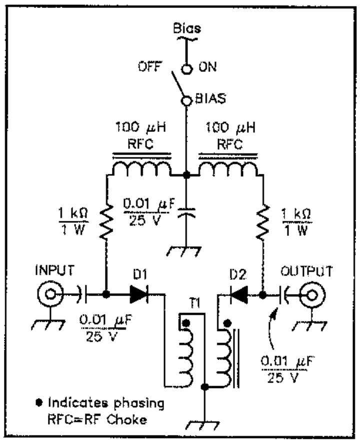


Figure 1—The diode switch used for the tests. D1 and D2, the diodes under test, included PN and PIN (power-rectifier and RF-specified) types. The capacitors are disc or monolithic ceramics. T1 consists of 11 bifilar turns of #28 enameled wire on an FT-37-43 ferrite toroidal form; the inductance of each winding is approximately 50 µH.

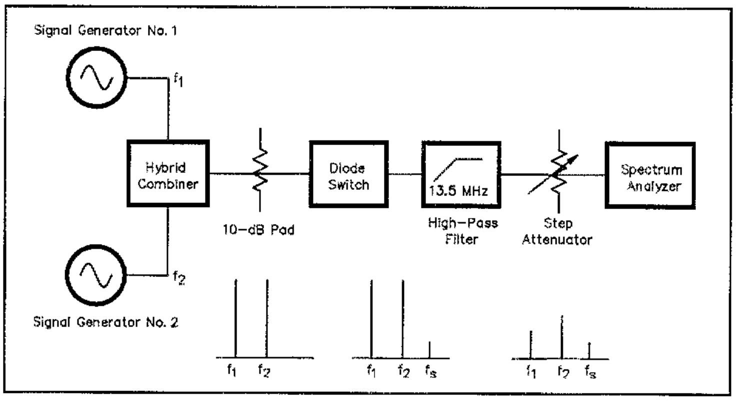


Figure 2—Setup for measuring the diode switch's second- and third-order intercept points. The high-pass filter increases the spectrum analyzer's second- and third-order intercept points as described in the text. Page 26.38 of the 1995 *ARRL Handbook* shows how to build a hybrid combiner.

valid only if the device under test responds equally well to f_1 and f_2.) My filter (Figure 3) is a five-element Chebyshev design with 3 dB of attenuation at 13.5 MHz and 30 dB of attenuation at 8 MHz.

Table 1
Diode Switch Insertion Loss (dB) at 10 MHz

Diode		*Bias Conditions Per Diode*			
Type	*Reverse*	*0 mA*	*5 mA*	*10 mA*	*20 mA*
1N4153	75	75	2	1	0.5
MPN3700	70	55	0.1	0.1	0.1
BAR17	75	70	0.3	0.1	0.1
1N4007	35	20	0.1	0.1	0.1

Table 2
Diode Switch Second-Order Intercept Point (IP2), dBm

Diode		*Bias Conditions Per Diode*			
Type	*Reverse*	*0 mA*	*5 mA*	*10 mA*	*20 mA*
1N4153	>80	>80	18	30	42
MPN3700	>80	80	66	70	72
BAR17	>80	>80	60	70	75
1N4007	>80	40	>80	>80	>80

Table 3
Diode Switch Third-Order Intercept Points (IP3), dBm

Diode		*Bias Conditions Per Diode*			
Type	*Reverse*	*0 mA*	*5 mA*	*10 mA*	*20 mA*
1N4153	>50	>50	20	30	37
MPN3700	>50	47	>50	>50	>50
BAR17	>50	50	>50	>50	>50
1N4007	>50	35	>50	>50	>50

Findings

I tested several types and makes of diodes. The 1N4153 is a generic PN switching diode like the 1N914; both have been around for years. The MPN3700 is a Motorola PIN diode intended for RF switching, and the BAR17 is a glass-envelope PIN switching diode made by Siemens. Some readers may wonder why I tested the 1N4007, a generic 1 kV PIV, 1 A rectifier diode. I tested it because it contains a PIN structure, and this has resulted in its use for RF switching by some experimenters and manufacturers. Because the 1N4007 is not intended for RF-switching use, I tested 1N4007s of several makes, finding no significant difference in the results.

Table 1 shows switch isolation and loss as functions of frequency. Table 2 shows the second-order intercept points, and Table 3 shows the third-order intercept points. The trends shown in Tables 2 and 3 are clear. We should certainly use RF-characterized PIN diodes whenever we want to maintain high second- or third-order intercepts. Surprisingly, however, many Amateur Radio transceivers do not use them. For example, the Kenwood TS-940S uses PN switching diodes, biased at 27 mA, for switching its input filters. Whereas that bias level will provide an IP2 greater than +42 dBm and a IP3 greater than +37 dBm, PIN diodes would provide much better intercepts at considerably lower bias currents.

It's conceivable that transceivers designed for mobile use might bias their switching diodes at levels much lower than the TS-940S's 27 mA to reduce power consumption. I therefore did tests at 5 mA, and found that PN switching diodes yielded IP2s on the order of +18 dBm and IP3s on the order of +20 dBm. These intercept points are dangerously close to being the limiting cases.

PIN diodes are more expensive than the PN diodes normally used. Nonetheless, the saving in bias current and the performance improvement afforded by PINs makes them well worth installing unless the signal levels present are guaranteed to be low.

Making Your Own Measurements

Even though I used a spectrum analyzer to perform these measurements, you can make similar measurements using your communications receiver in place of the spectrum analyzer in Figure 2. Adjust the attenuator for a midscale reading on your S meter while the receiver is tuned to a spurious response. With the receiver tuned to one of the test-signal frequencies, remove the high-pass filter and switch in attenuation until your S meter indicates the same midscale reading. The difference in the two attenuator readings is the system's third-order IMD dynamic range. Assuming you know P_S, you can determine IP2 and IP3.

Be careful not to overload your receiver's front end, or you might end up measuring your receiver's IP2 and IP3 instead of the IP2 and IP3 of the device under test!

Switching Diodes Versus Tuning Diodes in Tracking Preselectors

The practice of using a transceiver's automatic antenna tuner or other motor-driven tuned circuits[3] in a transceiver front end to increase its second-order intercept brings up an interesting point. Relatively few Amateur Radio receivers and transceivers of the 1960s and early 1970s used switching diodes in their front ends, and many had tunable preselectors that excluded signals capable of producing second-order spurious responses. These preselectors used manual band switching. Modified for electronic tuning—with tuning diodes replacing variable capacitors, or with switching diodes and fixed capacitors synthesizing variable capacitances—such a preselector can easily be made to track a radio's tuning if a microprocessor is used for system control. Which of these approaches would be better? Would tuning diodes generate spurious products? If so, would PIN-switched fixed capacitors (or inductors) be a better solution? To check this, I built the circuit shown in Figure 4.[4]

For these measurements to make sense, f_1 and f_2 cannot be as far apart as the test frequencies used in evaluating switching diodes. The tuned circuit steps up the voltage only of signals close to its resonant frequency. Stepped up and applied to the diodes, they intermodulate to create spurious signals. Both test signals must therefore be close enough in frequency to fall within the tuned circuit's bandwidth. I decided to set f_1=14.2 MHz and f_2=14.3 MHz. This produces third-order products of $2f_1-f_2$= 14.1 MHz and $2f_2-f_1$=14.4 MHz. Note, however, that since f_1 and f_2 are very close in frequency to the system's possible *third*-order spurious responses, the high-pass filter no longer increases the spectrum analyzer's third-order dynamic range. Intercept point measurements close to +30 dBm must therefore be considered to be suspect during this test.

The tuned circuit is decoupled using the capacitance dividers such that its loaded Q is 88 and its insertion loss is 8 dB. With a variable capacitor resonating the inductor, the filter's IP3 was +21 dBm. Adding a PIN-diode switch in series with the capacitor also resulted in an IP3 of +21 dBm, so using PIN-diode switching in a tracking preselector's tuned circuit(s) seems to involve no performance trade-off. Using a tuning-diode pair to resonate the inductor brought IP3 down to +11 dBm.

The IP3 measurement of +21 dBm may be suspect, depending on whether the spectrum analyzer itself was overloading. If so, the actual number may well have been *higher* than +21 dBm. The +11 dBm measurement for the tuning-diode pair is a valid measurement, however, and clearly illustrates that using tuning diodes to tune high-impedance circuits in the presence of strong signals is not a good idea.[5]

Since a PIN switch in series with the tuned circuit's capacitor does not seem to contribute to spurious-signal generation,

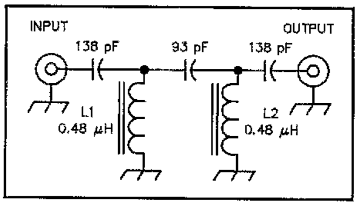


Figure 3—The 13.5-MHz high-pass filter. L1 and L2 each consist of 12 turns of #28 enameled wire on a T-25-6 toroidal powdered-iron core. The filter will work fine if capacitances of 130 pF (120 pF paralleled with 10 pF) and 92 pF (82 pF paralleled with 10 pF) are used instead of the nonstandard values shown.

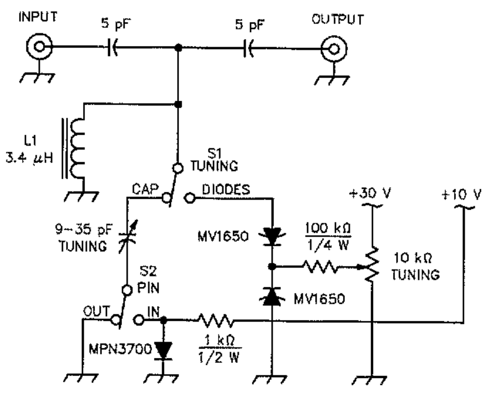


Figure 4—A loosely coupled tuned circuit for testing IMD production by PIN and tuning diodes in a narrow-band preselector. S1, **TUNING**, selects whether C1 or a pair of back-to-back MV1650 tuning diodes resonate L1. S2, **PIN**, adds or removes an MPN3700 PIN diode in series with C1. L1 consists of 33 turns of #28 enameled wire on a T-37-6 toroidal powdered-iron core. The MV1650, a "20-volt" tuning diode, exhibits a nominal capacitance of 100 pF at a tuning voltage of 4 volts.

a tracking preselector based on PIN-switched fixed-value components appears to be a viable alternative to the switched, sub-octave bandpass filters common in today's radios. If, however, the +21 dBm IP3 measurement shown for the test case is valid, an air-core inductor may be necessary to raise this intercept even further.[6]

Conclusion

RF-specified PIN diodes are the devices of choice for low-distortion switching at HF and above, for both bandpass filter selection and LC switching in a narrow-band preselector.

Although the presence of a PIN structure in the 1N4007 makes it seem attractive as a low-cost alternative to RF-specified PIN diodes, its insertion-loss performance when unbiased and reverse-biased, and its IMD performance when unbiased, is demonstrably inferior to RF-specified PINs.

The manually switched and tuned front-end filters of the 1960s and 1970s had much to offer in terms of second-order IMD, but we need not retrogress to those techniques to achieve improved IP2 and IP3 performance today. More attention paid to front-end filtering in general can produce the improvement we need.

Notes

[1] U. Rohde, "Key Components of Modern Receiver Design," Part 1, *QST*, May 1994, pp 29-32; Part 2, *QST*, Jun 1994, pp 27-31; Part 3, *QST*, Jul 1994, pp 42-45.

[2] W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur* (Newington: ARRL, 1986).

[3] U. Rohde, "Recent Advances in Shortwave Receiver Design," *QST*, Nov 1992, pp 45-55.

[4] This circuit was suggested by U. Rohde in private communication.

[5] U. Rohde's "Key Components of Modern Shortwave Receiver Design: A Second Look," elsewhere in this issue, describes a diode-tuned filter designed to minimize IMD.—*Ed.*

[6] The assumption being that intermodulation in the coil's ferromagnetic core is then the principle IMD contributor.

Further measurements done in the ARRL Lab, however, indicate that the filter's IP3 is quite a bit higher than +21 dBm. Measuring the small amount of IMD caused by ferromagnetic material is difficult—not only does a spectrum analyzer generate spurious signals, but signal generators can couple through the measurement setup's hybrid coupler and upset each other. (The leveling circuit in the Hewlett-Packard HP 8640 generator, two of which we use in the ARRL Lab, is a well-known IMD source.) The tiny ferrite cores used in commercial summer circuits are worthy of consideration as well. Unfortunately, there is no HF equivalent of a VHF/microwave isolator.—*Ed.*

First licensed in 1955 as K5BHB in Odessa, Texas, Tom Thompson has held W0IVJ since the early 1960s. He works as an electrical engineer for NOAA in Boulder, Colorado, where he does instrumentation for stratospheric atmospheric research. His ham interests include test equipment and HF transceiver construction.

New Books

BEHIND THE FRONT PANEL
The Design and Development of 1920s Radios

By David Rutland

Wren Publishers, PO Box 1084, Philomath, OR 97370; tel 503-929-4498. 1994, 159 plus viii pp; B&W diagrams and photos; 5x8 inches. Retail price $18.95 plus $2 s/h.

Reviewed By Jim Kearman, KR1S
Assistant Technical Editor

If you open a textbook today, you're likely to think radio was invented only 25 years ago. Yes, modern technology is wonderful, but there's no reason we can't admire and appreciate the work done in the past. If you're new to radio, a casual perusal of this book will amaze you. There are schematic diagrams and photos, and from the controls you can tell they're of radios—but they sure don't look like the radios of today.

David Rutland is a retired electrical engineer and a member of several vintage radio collector's societies. Although his book primarily covers broadcast receivers manufactured in the 1920s, he appropriately begins with a discussion of spark transmission and crystal receiving sets. This technology set the stage for the broadcast boom of the 1920s, brought on by the availability of relatively inexpensive vacuum tubes, a growing economy and the public's fascination with radio. Patiently and clearly, the author explains each development in the radio art, aided by excellent illustrations.

Radio was somewhat hamstrung by patent fights in its early years, and many circuits were developed for no other reason than to get around having to pay royalties. Back when there were only a few broadcasters nationwide, it didn't matter if you had to make three or four tuning adjustments to listen on another frequency. The result was increasingly complex radios, not all of which were technically sound. What lovely furniture they made, though. Looking at the photos, you can imagine the thrills Atwater Kents, Freed-Eisemanns, Grebes, Hazeltines, RCAs and others brought to their users, as they made the final adjustment that brought in the local broadcast station.

Later chapters (there are 13) cover the numerous circuits in use until Armstrong's superheterodyne made them all obsolete, mechanical and electrical tuning innovations, audio amplifiers and the ac-operated radio. Once manufacturers figured out how to power radios off the ac line and the superheterodyne was established, there wasn't much left to learn about radios, and that's where the author closes his narrative.

Rutland's writing is easy to read; you'll learn painlessly. Within this compact book is a vast amount of information that will interest anyone concerned with the history of radio or technology, professionally or as a hobby. Small-press books are often plagued with stilted writing and muddy photographs, but this one is a notable exception. Everyone interested in early radios should have a copy.

An ATV Station for 915 MHz

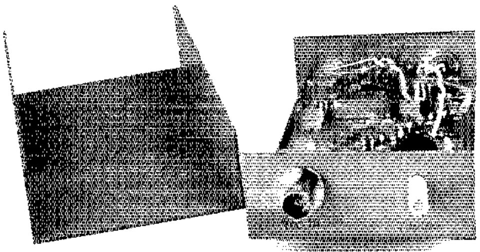
Part 2: A Tunable ATV Downconverter

Last month, you built the 915-MHz transmitter. This month, we present a companion downconverter to round out your 915-MHz station.

By Rudolf Graf, KA2CWL
111 Van Etten Blvd
New Rochelle, NY 10804-2321

and

William Sheets, K2MQJ
PO Box 200
Hartford, NY 12838



ATV operation is permitted on the 33-cm band (902 to 928 MHz), but, unfortunately, this frequency range isn't covered by today's cable-ready TV receivers. The current UHF TV spectrum ends at channel 69—800 to 806 MHz. Older receivers that tuned to channel 83, covered up to 890 MHz, but most such receivers had mechanical tuners, which often lacked RF amplifier stages. So we need a tunable, low-noise downconverter of reasonable cost to round out our ATV station. This simple downconverter for ATV reception at 902 to 928 MHz is quite effective—and it's easy to build.[4] Because the downconverter has a tunable LO, it's not recommended for narrowband SSB or CW work, but it's ideal for ATV, WBFM, or other wideband modes.

General Description

Refer to Figure 5 on the next page. This downconverter converts signals in the 902 to 928-MHz band to a 61.25 or 67.25-MHz IF (channel 3 or channel 4) to enable ATV reception on a standard VHF TV receiver or monitor. The downconverter features a low-noise RF amplifier feeding a Schottky diode double balanced mixer, a tunable LO and one IF preamplifier stage.

The RF amplifier is a low-noise, dual-gate MESFET followed by a second RF stage using a monolithic microwave integrated circuit (MMIC). This feeds a Schottky diode mixer assembly for good dynamic range and reduced susceptibility to intermodulation distortion and strong-signal interference—troublesome monsters in metropolitan areas or other strong-signal areas. The on-board local oscillator (LO) is voltage tuned and (if desired) can be set up for remote tuning—the remote tuning circuits are on board. This allows you to mast-mount the downconverter and minimize feed-line losses associated with operation on these frequencies. A separate dc feed is unnecessary since the coax (RG-59 is recommended) carries the downconverter's dc power, tuning voltage and IF signal. A dc-blocking capacitor is used to separate the supply and tuning voltages. This allows a cable run of several hundred feet if needed.

For portable operation, use in a packaged ATV transceiver or a video H-T, the on-board tuning potentiometer is used. The LO is voltage tuned and is fed from a regulated-voltage supply. A stage of IF post-amplification is included at 60 to 72 MHz. The downconverter PC-board footprint is 2.5×4 inches and matches the companion 915-MHz ATV transmitter in size. The power-supply requirements are 11 to 20 V dc at approximately 55 mA, or a variable supply from 11 to 19 V for combined dc and IF feed on a single coaxial cable for remote tuning. Overall gain is nominally 34 dB with an RF bandwidth greater than 30 MHz.

By using this downconverter and transmitter, a physically small 915-MHz ATV station—or even a video H-T—can be built by stacking the 2.5×4×1-inch units. A small PC-board camera, battery pack, relay, and a pocket TV receiver can be used with these units to make a very compact portable ATV station. Also, 915-MHz antennas are physically small and lightweight. Downconverter construction is easy and alignment accomplished simply by peaking for maximum received-signal strength, with no tricky alignment procedures or specialized RF test gear necessary for good results and performance.

Circuit Description

Refer to Figure 6. Incoming signals are routed from J1 to a tap on microstrip inductor Z1, which is tuned to resonance at 915 MHz (the band center frequency) by trimmer C1 and the gate-input capacitance of MESFET Q1, the first RF amplifier stage. Q1's drain current is about 8 mA. Chip capacitors C2, C3, C4, C6, C7 and C8 are RF bypass capacitors. (Chip capacitors are necessary for effective bypassing at 900 MHz because disc-ceramic capacitors are too inductive for this application.) Q1's drain feeds coupling network C5, Z2, C9, and Z3. Z2 and Z3 act as a two-winding transformer whose inductance and coupling coefficient is a function of the PC-board geometry and is fixed.

The overall gain so far is approximately 13 to 16 dB, depending on alignment. The intended band pass is 900 to 930 MHz at the −1 dB points, but by trading gain for bandwidth, more gain can be realized. This isn't always desirable for best dynamic range. The noise figure (NF) is 1.5 to 2 dB. (Although the MESFET itself has a rated noise figure of about 1 dB, unavoidable circuit losses make 2 dB a realistic figure.) It makes little sense to try to improve this, as cable and connector losses set the limit of achievable system NF. Furthermore, this converter is not intended for narrowband, weak-signal work anyway, and a 2-dB NF is far better than any cable-ready TV tuner or UHF TV tuner yields without an external preamplifier.

The second RF amplifier stage uses an MAR-1 MMIC, IC1. This device has 50-Ω input and output impedances and a gain of about 15 dB at 1 GHz. Dc is fed to IC1 via R6. A prepackaged Schottky diode mixer (a Mini-Circuits SBL-1X) is used for signal conversion to IF. The coupling network and mixer have an overall conversion loss of

[4] Notes appear on page 32.

about 10 to 11 dB. Total gain at this point is 19 to 22 dB, depending on RF alignment. This gain is typical for a relatively flat passband (±2 dB) from 900 to 930 MHz. Gain may vary if alignment is changed.

The output from mixer M1 contains the sum and difference products of the RF and LO signals—only the difference frequencies (60-72 MHz) are desired. The LO signal frequency varies from 820 to 870 MHz, depending on the desired IF and received frequency. IF output can be either VHF channel 3 or 4 (61.25 or 67.25 MHz) and if desired, the LO can be set up for channel 2 (or even channel 5 or 6) to avoid interference. However, some slight modifications may be needed at the IF-output network. The design center is channel 3 or channel 4. Mixer output is fed to low-pass filter L2-C15. The collector load is tuned to 60 to 72 MHz and feeds a 75-Ω cable that connects to the TV or monitor tuned to channel 3 or 4. L4 can be made larger if a channel 2 IF is to be used, or decreased if a channel 5 or 6 IF is required. This stage has a gain of 12 to 15 dB. Therefore, the overall converter gain is 34 dB with ±4 dB expected variation due to alignment, circuit and device tolerances. Nothing is critical.

The LO signal is supplied by Q3 and associated circuitry. Z6 is a resonant line tuned by the output capacitance of Q3, and the series connection of trimmer C20 and varactor diode D1. D1 has a nominal capacitance of 10 pF at 4 V bias and varies about ± 40% from 1 to 9 V. By properly proportioning C20 and Z6, a tuning range of 20 to 30 MHz, centered at 855 MHz, can be obtained. A parallel arrangement would require unwieldy circuit constants at this frequency. RF feedback from Q3's collector to emitter is via internal transistor capacitances. A tap on resonator Z6 supplies about 0.5 V RMS LO injection to mixer M1. Oscillator frequency stability is adequate for TV reception or other relatively wideband use. The LO is supplied regulated +8 V from regulator IC2.

Assembly

When assembling the PC board, solder all trimmer capacitors and grounded resistor leads to *both sides of board.* This is essential for good RF grounding. Near-zero lead length is a must in the RF circuits. Potentiometer R14 can be mounted in any of two positions; this allows some freedom in mounting the PC board in an enclosure.

Next, install all capacitors except for the chip capacitors. Install all transistors except Q1. *Observe proper pin orientation.* Mount all transistors 1/8 inch from the PC-board surface. Solder Q3's emitter lead to the top of the PC board where it passes through ground foil. Note that on the board Q3's pinout is base, emitter, collector (BEC) rather than emitter, base, collector (EBC).

Tack mixer M1's case to the ground plane in several places. Use a #8-32 screw to hold L1, L2, and L3 during installation. After installing the coils (see Table 2), re-

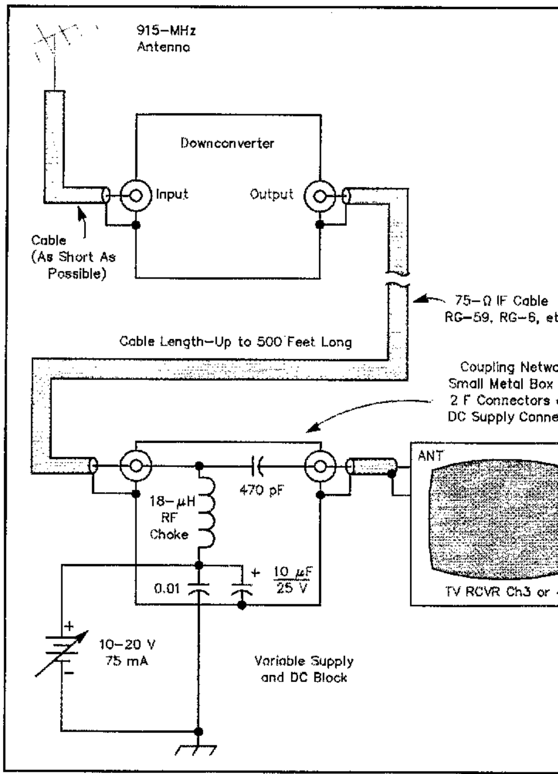
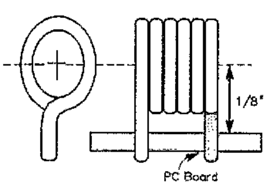


Figure 5—Block diagram of the 915-MHz downconverter hook-up.

Table 2
Coil Construction Information

Note: All inductors are wound with #22 enameled wire using a #8-32 screw as a coil form. Remove the screw from the inductor after winding.

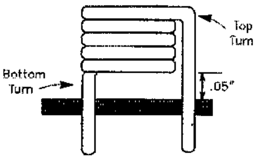
L1



L1—7 turns

Strip and tin the wire ends and mount the inductor 1/8-inch above the PC board.

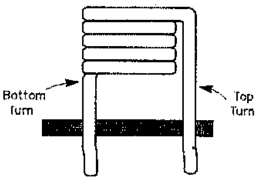
L2



L2—6 turns

Strip and tin the wire ends. Mount the inductor about 0.05 inch above the PC board as shown. (You can use a #8-32 screw to hold the inductor in place while soldering.)

L3



L3—11½ turns

nsert a #8-32 ferrite slug into the inductor (see text). Strip and tin the lead ends and mount it like L2.

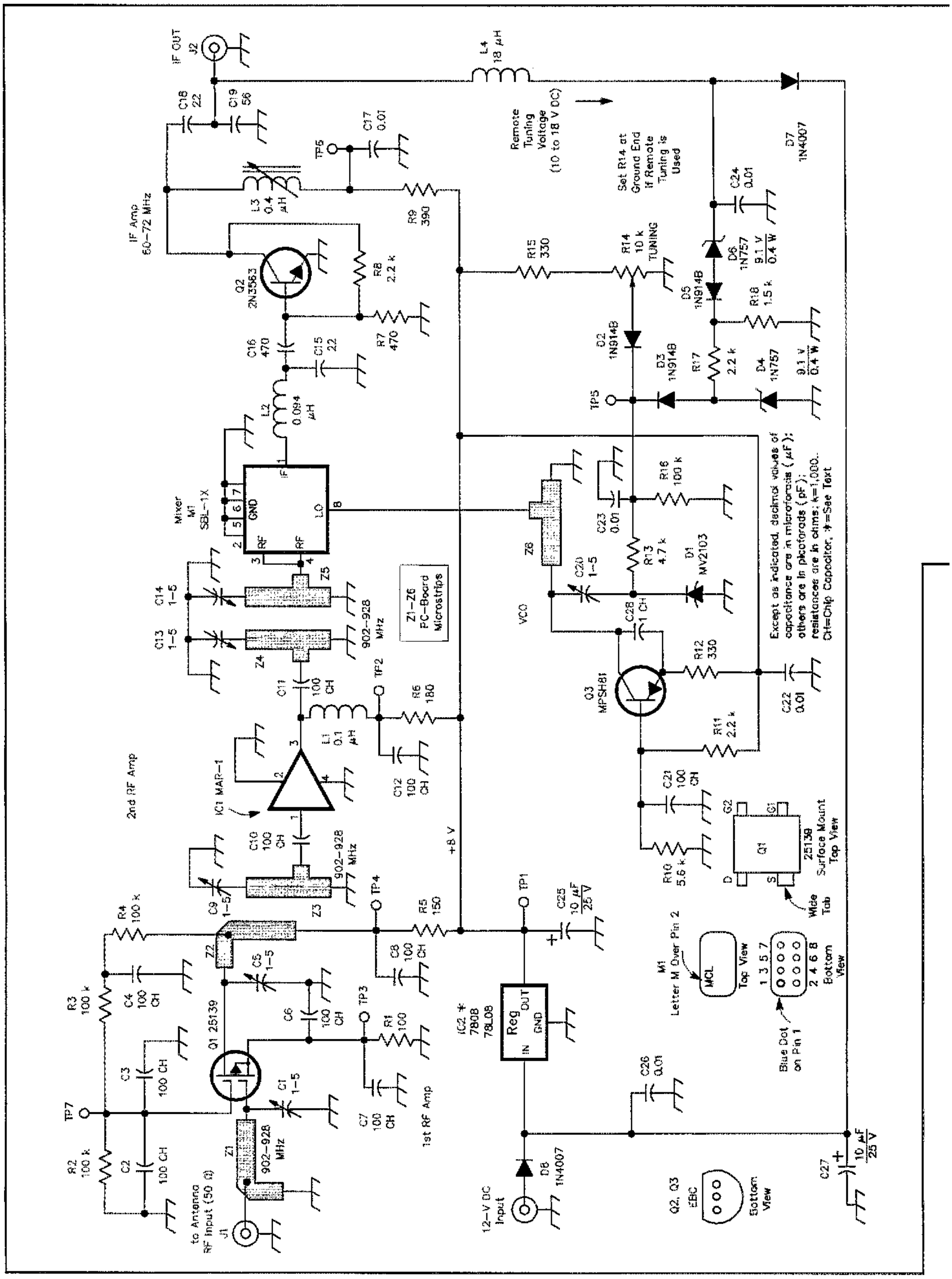

to Antenna RF Input (50 Ω)
J1
Z1 902–928 MHz
R2 100 k
C2 100 CH
TP7
C3 100 CH
R3 100 k
C4 100 CH
R4 100 k
Q1 2S139
C1 1–5
C5 1–5
C6 100 CH
TP3
R1 100
C7 100 CH
1st RF Amp
Z2
C9 1–5
Z3 902–928 MHz
TP4
C8 100 CH
R5 150
C10 100 CH
IC1 MAR-1
2nd RF Amp
+8 V
L1 0.1 µH
C12 100 CH
TP2
R6 180
C11 100 CH
Z4
C13 1–5
Z5
C14 1–5
902–928 MHz
Z1–Z6 PC-Board Microstrips
Mixer M1 SBL-1X
GND
RF
LO
IF
L2 0.094 µH
C16 470
C15 22
R7 470
Q2 2N3563
R8 2.2 k
IF Amp 60–72 MHz
L3 0.4 µH
R9 390
TP6
C17 0.01
C18 22
C19 56
IF OUT J2
L4 18 µH
Remote Tuning Voltage (10 to 18 V DC)
Set R14 at Ground End if Remote Tuning is Used
R15 330
R14 10 k TUNING
D2 1N914B
D3 1N914B
TP5
R17 2.2 k
D4 1N757 9.1 V 0.4 W
R18 1.5 k
D5 1N914B
D6 1N757 9.1 V 0.4 W
C24 0.01
D7 1N4007
Z6
C20 1–5
VCO
R16 100 k
C23 0.01
R13 4.7 k
D1 MV2103
C28 1 CH
Q3 MPSH81
R12 330
R11 2.2 k
C22 0.01
C21 100 CH
R10 5.6 k
G2 G1 D S
Q1
2S139 Surface Mount Top View
Wide Tab
TP1
C25 10 µF 25 V
IC2 ✱ 7808 78L08
Reg
IN OUT GND
C26 0.01
12-V DC Input
D8 1N4007
Q2, Q3 EBC
Bottom View
M1 Letter M Over Pin 2
MCL
Top View
1 3 5 7
2 4 6 8
Bottom View
Blue Dot on Pin 1
C27 10 µF 25 V
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000. CH=Chip Capacitor. ✱=See Text

Figure 6—Schematic of the 915-MHz downconverter. Unless otherwise specified, all resistors are ¹/₈-W units.

C1, C5, C9, C13, C14, C20—1 to 5-pF trimmer capacitor
C2, C3, C4, C6, C7, C8, C10, C11, C12, C21—100-pF NP0 chip capacitor
C15, C18—22-pF NP0 capacitor.
C16—470-pF disc-ceramic capacitor
C19—56-pF NP0 capacitor.
C25,C27—10-µF, 25-V electrolytic capacitors
C28—1-pF NP0 chip capacitor
D1—MV2103
D2, D3, D5 —1N914
D4, D6—1N757
D7, D8—1N4007
IC1—MAR-1
IC2—78L08 or 7808 8-V positive regulator (see text)
L1, L2, L3—See Table 2.
L4—18-µH RF choke
M1—SBL-1X mixer assembly
Q1—NEC 25139 MESFET
Q2—2N3563
Q3—MPSH81
R14—10-kΩ, PC-mount potentiometer
Z1-Z6, incl—Etched on PC board.

Misc: PC board, hardware, enclosure, BNC connector, F connector, dc-input connector, shaft for R14, Cambion green slug for L3, threaded #8-32×¹/₄-inch.

Test Point	**Expected (V)**	**If expected result is not obtained:**
Jct D8, IC2	+11.4	Check D8
Jct IC2, C25 (TP1)	+ 8.0	Is IC2 in backwards?
Jct R5, C8, Z2 (TP4)	+6.2 to +6.8	Q1 correctly installed? R5, C8?
Jct R1, C6, C7 (TP3)	+0.6 to +1.2	Q1 correctly installed? R1, C6, C7?
Jct R2, R3, C2, C3 (TP7)	+1.7 to +2.3	Q1? R2, R3, R4, C2, C3, C4?
Jct R6, C12, L1 (TP2)	+5.0 to +5.5	IC1 installed properly? R6, C12, L1?
Jct R9, C17, L3 (TP6)	+3.0 to +4.2	Q3 installed properly? R8, R9, R10?
Jct D2, D3, C23 (TP5)	0 to > +7.0 as R14 is rotated	R15 okay?
Jct R13, C20, D1	0 to > +7.0 as R14 is rotated	Is D1 installed correctly?
Q3 base	+5.5 to + 6.5	Q3, R10, R11, R12, C21
Q3 emitter	0 to 0.7 V more than base Q3	Is Q3 installed correctly?

move the screw and insert a ferrite tuning slug into L3. Coat L1 and L2 with a clear lacquer or nail polish. Install all ground-plane jumpers. *This is very important!*

Install all chip capacitors on the PC-board *bottom*. After installing them, avoid flexing the PC board as this may crack the chip capacitors. Install four grounding jumpers in the four holes around IC1's location. *These grounding jumpers are required.* When installing IC1, the brown dot on the case indicates pin 1's location. This lead may also have a diagonal cut. Use a magnifying glass to be sure. Q1's thicker lead is oriented as the source lead. Be careful—Q1 is static sensitive!

Tune-Up

Successful downconverter tune-up requires either a known signal in the 902 to 928-MHz band, or access to a frequency counter and a signal generator covering this range. A sweep generator is an excellent aid if you have access to one. If you have the 915-MHz ATV transmitter finished, it can be used as a signal source, but remember to use a dummy antenna at the transmitter output and keep it at some distance from the downconverter to prevent overloading it. You'll need a suitable TV or monitor tuned to either VHF channel 3 or 4, (a B&W portable TV will do), a variable power supply delivering 10 to 19 V dc at 75 mA or more (preferably with built-in metering), a VOM or DVM and clip leads and cables. A frequency counter that is reliable to at least 1000 MHz is a great help in setting up the LO and finding out where you are during tune-up.

Connect the dc power supply to the downconverter's 12-V input. (No damage to the converter will occur if the supply polarity is accidentally reversed.) Note the current drawn from the supply—it should be around 50 to 60 mA. If it's higher than 75 mA, check for short circuits. If the current drawn is less than 45 mA, something may be open or missing. When the current drawn is between 45 and 75 mA, proceed with following steps:

1) Refer to Figure 6 while performing the checks listed in the caption. Unless otherwise stated, the negative meter lead is connected to the negative power-supply terminal.

2) Set R14 at its midpoint and set C20 so its plates are meshed about 25%. Measure Q3's emitter voltage while rotating R14. There should be some perceptible voltage change at Q3's emitter as R14 is rotated end to end. Placing a metallic screwdriver blade on Q3's collector should also produce some change in this reading if Q3 is oscillating. A change of 0.03 to 0.1 V is about what you'll see if all is okay. This checks out the LO circuit.

3) Nothing should be getting hot. If you experienced some deviation in the voltage checks but they weren't much different than what's called for, all may still be okay, but remember this and identify it as suspect if any further problems are encountered. If the following results are successful and the downconverter works well, those minor variations can be ignored. Any major variation should be investigated as there is a problem. Make sure that your test equipment is set properly.

4) If Steps 1 to 3 have been completed successfully, connect a variable-voltage supply to the junction of C18, C19 and L4. The positive lead is hot, negative lead goes to ground. Make sure the wiper of R14 is at ground (extreme CCW as viewed from shaft side). Set the supply to +11V and check for the following voltages:

TP1 +8 V dc as before. If not, check L4, D7, C24, C26, C27

TP5 < +1.2 V dc. If not, check D6, D5, R18, R17, D4, D3

5) Increase the supply voltage to 19 or 20 V. Repeat the Step 4 readings. The voltage at TP1 should read the same as before. The voltage at TP5 should be about +9 V. If not, check all components mentioned in Step 4.

6) Connect the center conductor of a length of 75-Ω coaxial cable to the junction of C18, C19, L4 and attach the shield braid to ground. Terminate the end of the cable in a connector suitable for your TV or monitor. Tune the TV to channel 3 or 4—whichever is unused in your area—and set the controls for normal reception. If you have a sensitive RF millivoltmeter, use it instead of the TV set as an output indicator; it's easier to interpret.

Connect a suitable 915-MHz signal source between the downconverter input and ground. The source can be either a generator or received signal. Preset all trimmer caps (there are six) to 25% mesh. Preset R14 to the center of its rotation. Connect the power supply's positive lead to D8, negative lead to ground. Keep the PC board bottom at least ¹/₂ inch above any kind of material that may detune the microstrip elements on the PC board bottom. Four ¹/₂ or ³/₄-inch standoffs in the four corner holes are ideal for this purpose. Use only a nonmetallic tuning tool for all adjustments. Insert the slug halfway into L3.

7) Activate the signal source and slowly rotate C20 until some indication of reception is seen on the monitor or RF voltmeter. Use a fairly strong signal at first. On confirmation of reception, decrease the signal level until it's just enough for a reliable indication. As you remove the tuning tool from C20, undoubtedly some detuning will occur. You'll have to compensate a little to get the tuning correct when the tool is removed.

If you have a frequency counter good at 1000 MHz, you can use it to simplify this step. Loop-couple the counter to Q3, staying at least ¹/₂ inch away from C20 and Q3, and set C20 for 845 MHz (±3 MHz) with R14 centered. If you use the counter, rotate R14 full CW and then CCW and verify that the LO frequency goes below 845 MHz and above 867 MHz if you intend to use a channel 3 IF, or below 839 MHz and above 861 MHz for a channel 4 IF. Reset C20 as needed. (If no counter is available, skip this step.)

Peak C13 and C14 for maximum response. Decrease the signal amplitude and peak C5 and C9. Next, reduce the signal amplitude until it's just enough to know it's present and then peak C1. Decrease the signal amplitude further and repeak C5, C9, C13 and C14 for best output. Now, adjust the slug in L3 for maximum output and secure it in place using clear lacquer, glue or nail polish.

8) If you cannot obtain sufficient tuning range with the LO, you can try removing chip capacitor C28, but this may affect LO operation. Removing C28 is only necessary in one out of ten cases, for very wideband coverage, when one is trying to obtain a tuning range greater than 30 MHz. (At the time of this writing [Spring 1994], most all ATV activity takes place between 910 and 928 MHz following the ARRL band plan, so a tuning range of only 18 MHz is really needed.) Restricting the tuning range also makes tuning of the received signals easier and improves the LO stability.

9) Repeat the adjustments to further improve performance. If you have access to a sweep setup, align the downconverter for a flat response within ±2 dB from 900 to 930 MHz. The alignment is straightforward, so experiment to get best results. If everything is operating properly, you should have 30 to 38 dB gain with an NF of under 3 dB. For remote tuning, make sure your variable supply is clean. Any noise or hum causes FM of the LO signal and a noisy received picture, since the tuning voltage in the downconverter is not regulated. Be sure also to set R14 fully CCW when remote tuning is used.

10) If you're contemplating downconverter operation from supply voltages over 20 V, replace IC2 with a 7808 TO-220 unit as the smaller 78L08 may run too hot.

Summary

This downconverter (See Figure 7) has proven to be highly satisfactory for ATV reception in the 915-MHz band. It should prove useful as well for reception of many other signals in the 900 to 960-MHz range. For best reception, mount this unit close to the antenna as feed-line loss is astronomical at 900 MHz when using small-diameter coaxial cables. Although not intended for such use, this downconverter and an FM broadcast receiver tuned to around 88 to 92 MHz as an IF (pick a quiet spot) enables reception of miscellaneous services using this region, such as police, business, STL links, and various 900-MHz wireless video links.[5] You may have to reset the LO for such use.

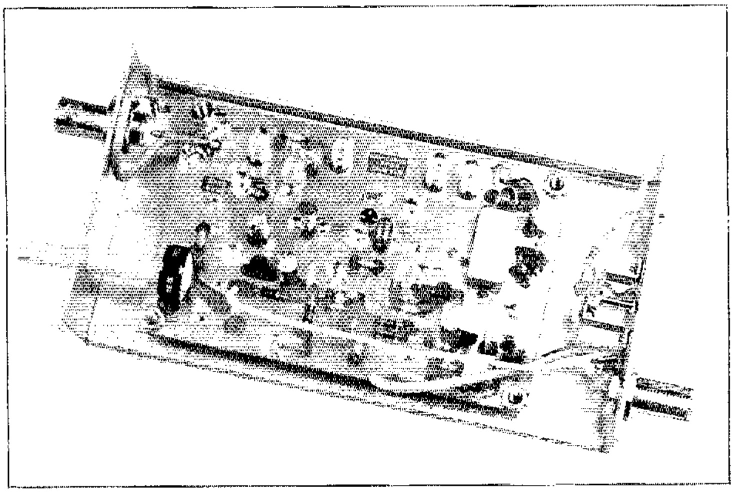


Figure 7—An inside view of the 915-MHz downconverter.

This downconverter can be used to copy narrowband signals, such as 5 or 15-kHz deviation FM, or even AM, but because the LO is free-running, the stability and tuning rate leave much to be desired. Remember: Some of the transmissions you hear may be protected under the Communications Act as to legality of monitoring, so use some discretion and never divulge anything you've heard to a third party.

Notes

[4] An etched and drilled PC board and some parts for this project are available from:
North Country Radio, PO Box 53, Wykagyl Station, New Rochelle, NY 10804-0053.
Contact them for details. Tel 914-235-6611; fax 914-516-6051. Catalogs can be purchased for $1 and a no. 10 envelope bearing 75 cents in First Class postage.

[5] While monitoring our own transmission on 910.25 MHz, we picked up the security video system of one of the local mall stores while parked about a block away. Evidently, the store is using a 900-MHz link.

QST

QEX:

The ARRL Experimenter's Exchange

The November issue of *QEX* includes:

☐ DSP is the obvious choice for a modern HF modem design. This month, Johan B. Forrer, KC7WW, presents "An Adaptive HF DSP Modem for 100 and 200 Baud" that provides state-of-the-art performance by using state-of-the-art technology.

☐ "A Better A/D and Software for the DDC-Based Receiver," by Peter Traneus Anderson, KC1HR, updates the author's March, 1994, *QEX* article, "A Simple SSB Receiver Using a Digital Down Converter" with a 10-bit A/D, providing more than 60 dB of dynamic range.

☐ Part 3, the final part of "Practical Microwave Antennas," by Paul Wade, N1BWT, discusses lens antennas you can build and shows how to make effective microwave antenna measurements.

☐ In this month's "RF" column, Zack Lau, KH6CP/1, shows how to improve the performance of 10-GHz wideband FM systems that use Gunnplexers.

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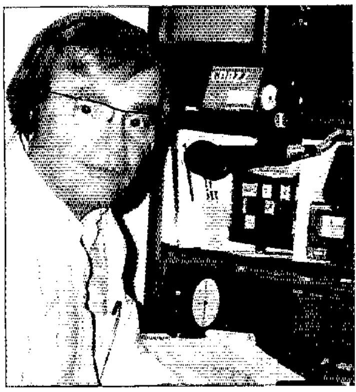
Strays



Dave Bell, W6AQ, of Hollywood, California, has worked on several Amateur Radio films, and will be keynote speaker at the 7th annual joint luncheon meeting of the Society of Wireless Pioneers (SOWP), Pacific Southwest Chapter, and the Quarter Century Wireless Association (QCWA) Arizona Barry Goldwater Chapter 16, at the Safari Resort in Scottsdale, Arizona, on Saturday, December 10.—*Bill Jackson, W6HDP, Director, SOWP Pacific Southwest Chapter, 4930 N Hobo Cir, Prescott Valley, AZ 86314.*

The Quick Powerhouse

Tired of missing a new grid square while waiting for your amplifier to warm up? Build this almost-instant-on, 1.2-kW linear power amplifier for 2 meters.

By Russ Miller, N7ART
12041 SW Peninsula Drive
Crooked River Ranch, OR 97760

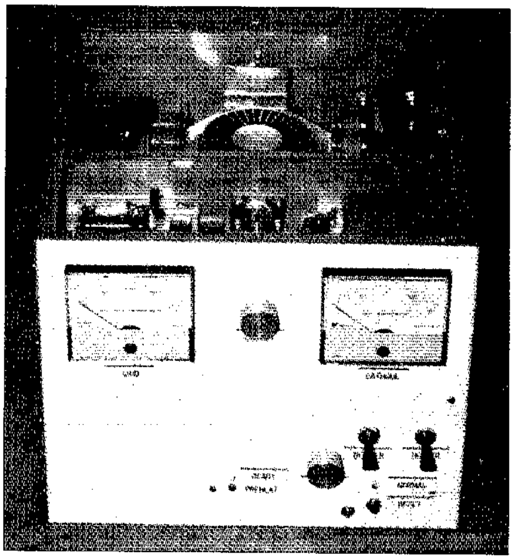


Figure 1—This table-top 2-meter power amplifier uses a quick-warm-up tube, a real plus when the band suddenly opens for DX and you want to join in.

It's that typical, early summer afternoon, when, out of the blue comes co-channel interference to the low-band (channels 2 through 6) TV channels. Time to turn on the 2-meter receiver. Whoops! There's a sporadic-E opening, and stations are booming in from over 1000 miles away. Quickly, I hit the switches to warm up my linear amplifier. Time marches on, and now the amp is ready—only took 3 or 4 minutes. But that juicy, far-away clatter via sporadic E is gone! Woe is me!

Having had this happen more than once, I was delighted to find that Eimac was producing a new tube with an almost-instant-warm-up filament that functions well at 2 meters—the 3CX1200Z7.[1] The Z7 is different from the 3CX1200A7 by virtue of its external grid ring, redesigned anode assembly and a 6.3-V ac filament. One advantage to the 3CX1200Z7 is the wide range of plate voltages that can be used, from 2000 to 5500. This amplifier looks much like the easily duplicated W6PO design,[2] except for the plate collet and the addition of some control circuitry. The plate collet is adapted from the W6PO 222-MHz design.[3] The RF deck is a compact unit, designed for table-top use (Figure 1). Table 1 gives some data on the 3CX1200Z7; Table 2 lists CW operating performance for this amplifier.

Input Circuit

The tuned-filament T network matches the 50-Ω drive source to the filament input impedance, providing a very low input SWR. Tuning is easy and docile. Grid bias is provided by an 8.2-V, 50-W Zener diode. Cutoff bias is provided by a 10-kΩ, 25-W resistor. A relay on the control board shorts out the cutoff-bias resistor, to place the amplifier in the **TRANSMIT** mode.

I didn't use a tube socket. Instead, I bolted the tube directly to the top plate of the subchassis, using the four holes (drilled to clear a #6 screw) in the grid flange. Connections to the heater pins are via drilled and slotted brass rods. The input circuit is contained within a 3½×6×7¼-inch (HWD) subchassis (Figure 3).

Control Circuit

The control circuit (Figure 4) is a necessity. It provides grid overcurrent protection, keying control, and filament surge control. To protect the tube filament from stressful surge current, a timer circuit places a resistor in series with the primary of the filament transformer. After four seconds, the timer shorts the resistor, allowing full filament voltage to be applied. C2 and R4 establish the time delay.

Table 1
3CX1200Z7 Specifications

Maximum Ratings

Plate voltage: 5500 V
Plate current: 800 mA
Plate dissipation: 1200 W
Grid dissipation: 50 W

Table 2
CW Operating Data

Plate voltage: 3200 V
Plate current (operating): 750 mA
Plate current (idling): 150 mA
Grid current: 165 mA
DC Power input: 2400 W
RF Power output: 1200 W
Plate dissipation: 1200 W
Efficiency: 50%
Drive power: 85 W
Input reflected power: 1 W

Another timer inhibits keying for a total of 10 seconds, to give the internal tube temperatures a chance to stabilize. C1 and R3 determine the time constant of this timer. After 10 seconds, the amplifier can be keyed by grounding the keying line. When the amplifier is not keyed, it draws no plate current. When keyed, idle current is approximately 150 mA, and the amplifier only requires drive to produce output. A safety factor is built in: the keying circuit requires +12 V from the high-voltage supply. This feature ensures that high voltage is present before the amplifier is driven.

The grid overcurrent circuit should be set to trip if grid current reaches 200 mA. When it trips, the relay latches and the **NORMAL** LED extinguishes. Restoration requires you to press the **RESET** switch.

Plate Circuit

Figure 5 shows an interior view of the plate compartment.The anode collet is patterned after the one used in the W6PO 222-MHz amplifier. The differences are small. A 4×2¼-inch tuning capacitor plate and a 2×2-inch output coupling plate are centered on the collet.These parts are the same size and shape as those used on the 2-meter W6PO amplifier. The remaining difference is the diameter of the hole for the 3CX1200Z7 anode. Sufficient clearance must be left for the fingerstock. The hole diameter will be approximately 3⅝ inches. Figure 6 is a drawing of the plate

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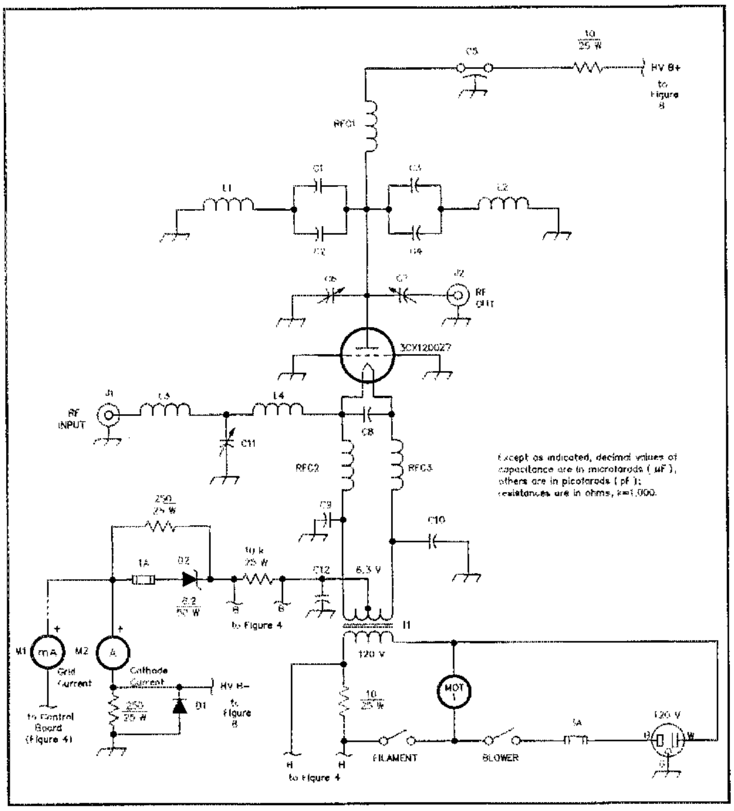
[1] Notes appear on page 37.



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

Figure 2—Schematic diagram of the 2-meter amplifier.

C1-C4—100 pF, 5 kV, type 850
C5—1000 pF, 5 kV
C6—Anode-tuning capacitor; see text and Figure 5 for details
C7—Output-loading capacitor; see text and Figure 7 for details
C8-C10—1000-pF silver mica, 500 V
C11—30-pF air variable
C12—0.01 µF, 1 kV
D1—1000 PIV, 3-A diode, 1N5408 or equiv
D2—8.2-V, 50-W Zener diode, ECG 5249A
J1—Chassis-mount BNC connector
J2—Type-N connector fitted to output coupling assembly (see Figure 7)
L1, L2—Plate lines; see text and Figure 6 for details
L3—5 t no. 14, 1/2-inch diameter, close wound
L4—3 t no. 14, 3/8-inch diameter, 1/4-inch spacing
RFC1—7 t no.14, 5/8-inch diameter, 1 1/8 inch long
RFC2, RFC3—10 t no. 12, 3/8-inch diameter, 2 inches long
T1—Filament transformer. Primary: 120 V; secondary: 6.3 V, 25 A, center tapped Available from Avatar Magnetics (Ronald C. Williams, W9JVF, 240 Tamara Trail, Indianapolis, IN 46217, 317-783-1211); part number AV-539
M1—Grid milliammeter, 200 mA dc full scale
M2—Cathode ammeter, 2 A dc full scale
MOT1—140 free-air cfm, 120-V ac blower, Dayton 4C442 or equivalent.

Sources for some of the "hard to get parts" include:

Fair Radio Sales, 1016 E Eureka, Lima, OH 45802, tel 419-227-6573

Surplus Sales of Nebraska, 1502 Jones Street, Omaha, NE 68102, tel 402-346-4750.

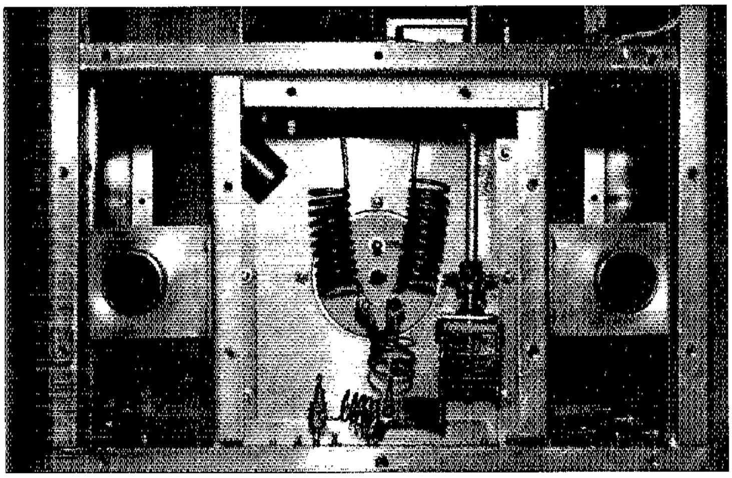


Figure 3—This view of the cathode-circuit compartment shows the input tuned circuit and filament chokes.

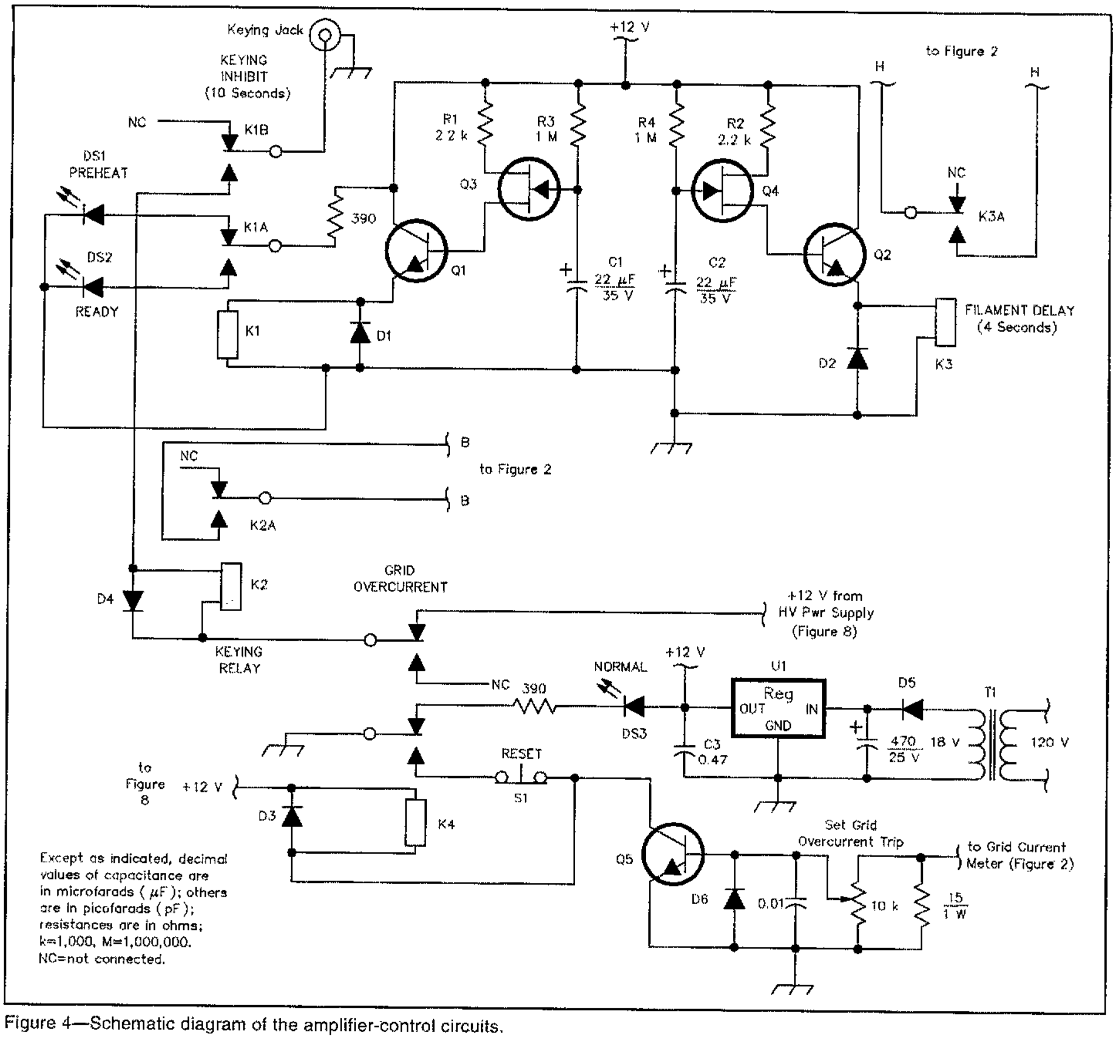
line and Figure 7 is a drawing of the output coupling asssembly.

Cooling

The amplifier requires an air exhaust through the top cover, as the plate compartment is pressurized. You can fashion a chimney from a 3 1/2-inch waste-water coupling (black PVC) and a piece of 1/16-inch-thick Teflon sheet. The PVC should extend down from the underside of the amplifier cover plate by 1 1/8 inches, with the Teflon sheet extending down 1/4 inch from the bottom of the PVC.

The base of the 3CX1200Z7 is cooled by using bleed air from the plate compartment, which is directed at the tube base, through a 3/8-inch tube set into the subchassis wall at a 45° angle.

The recommended blower will supply more than enough air for any temperature zone. A smaller blower is not recommended, as it is doubtful that the base area will be cooled adequately. The 3CX1200Z7



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000. NC=not connected.

Figure 4—Schematic diagram of the amplifier-control circuits.

C3—0.47-µF, 25-V tantalum capacitor
D1-D5—1N4001 or equivalent
D6—1N4007 or equivalent
DS1—Yellow LED
DS2—Green LED
DS3—Red LED
K1—Keying-inhibit relay, DPDT, 12-V dc coil, 1-A contact rating (Radio Shack 275-249 or equivalent)
K2—Amplifier keying relay, SPDT, 12-V dc coil, 2-A contact rating (Radio Shack 275-248 or equivalent)
K3—Filament delay relay, SPST, 12-V dc coil, 2-A contact rating (Radio Shack 275-248 or equivalent)
K4—Grid-overcurrent relay, DPDT, 12-V dc coil, 1-A contact rating (Radio Shack 275-249 or equivalent)
Q1, Q2, Q5—2N2222A or equivalent
Q3—MPF102 or equivalent
Q4—2N3819 or equivalent
S1—Normally closed, momentary pushbutton switch (Radio Shack 275-1549 or equivalent)
T1—Power transformer, 120-V primary, 18-V, 1-A secondary
U1—+12 V regulator, 7812 or equivalent

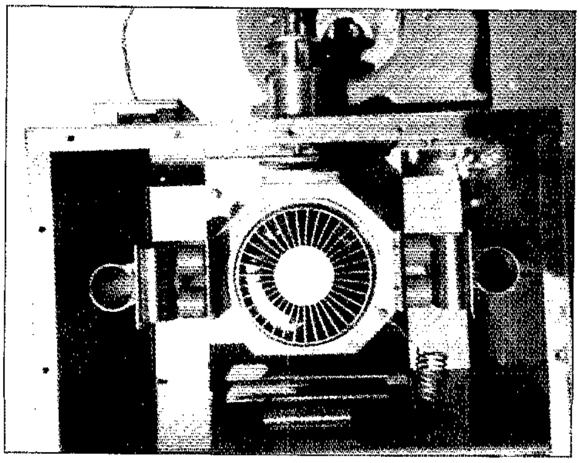


Figure 5—This top view of the plate compartment shows the plate-line arrangement, C1-C4 and the output coupling assembly.

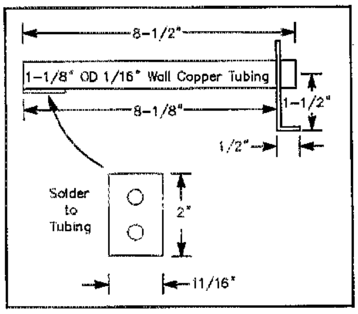


Figure 6—Plate line details.

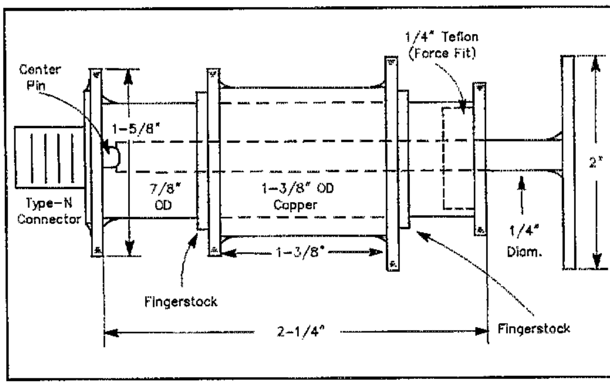


Figure 7—Details of the output coupling assembly.

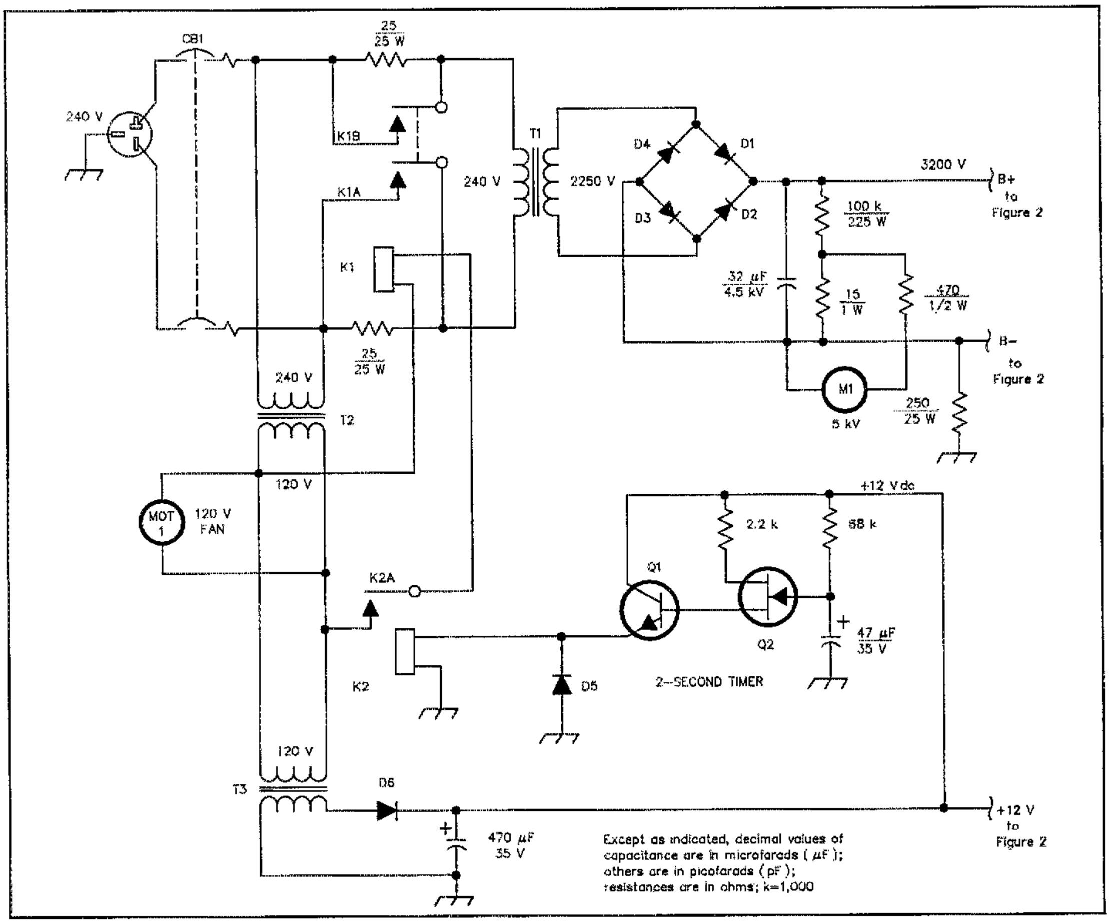


Figure 8—Schematic diagram of the high-voltage power supply recommended for use with the power amplifier.

D1-D4—Strings of 4 each, 1000-PIV, 3-A diodes, 1N5408 or equivalent

K1—DPST relay, 120-V ac coil, 240-V-ac, 20-A contacts (Midland Ross 187-321200 or equivalent)

K2—SPDT miniature relay, 12-V dc coil (Radio Shack 275-248 or equivalent)

M1—High-voltage meter, 5 kV dc full scale (1-mA meter movement used with series resistors shown in drawing)

MOT1—Cooling fan, Torin TA-300 or equivalent

Q1—2N2222A or equivalent

Q2—MPF102 or equivalent

S1—20-A hydraulic/magnetic circuit breaker (Potter and Brumfield W68X2Q12-20 or equivalent)

T1—High-voltage power transformer, 240-V primary, 2250-V, 1.2-A secondary (Avatar AV-538 or equivalent)

T2—Stepdown transformer, Jameco 112125, 240-V to 120-V, 100 VA

T3—Power transformer, Jameco 104379, 120-V primary; 16.4 V, 1-A secondary (half used) filament draws 25 A at 6.3 V! It alone generates a great deal of heat around the tube base seals and pins, so good air flow is critical.

Construction

The amplifier is built into a 12×12×10-inch enclosure. A 12×10-inch partition is installed 7¼ inches from the rear panel. The area between the partition and the front panel contains the filament transformer, control board, meters, switches, Zener diode and miscellaneous small parts. Wiring between the front-panel area and the rear panel is through a ½-inch brass tube, located near the shorted end of the right-hand plate line.

High voltage is routed from an MHV jack on the rear panel, through a piece of RG-59, just under the shorted end of the left-hand plate line. The cable then passes through the partition to a high-voltage standoff insulator made from nylon. This insulator is fastened to the partition near the high-voltage feed-through capacitor. A 10-Ω, 25-W resistor is connected between the insulator and the feed-through capacitor.

The plate lines are connected to the dc-blocking capacitors on the plate collet with 1¼×2-inch phosphor-bronze strips. The bottom of the plate lines are attached to the sides of the subchassis, with the edge of the **L**-shaped mounting bracket flush with the bottom of the subchassis.

When preparing the subchassis top plate for the 3CX1200Z7, cut a 2$^{11}/_{16}$-inch hole in the center of the plate. This hole size allows clearance between the tube envelope and the top plate, without putting stress on the envelope in the vicinity of the grid flange seal.

Exercise care in placing the movable tuning plate and the movable output coupling disc, to ensure they cannot touch their fixed counterparts on the plate collet.

Operation

When the amplifier is first turned on, it cannot be keyed until:

- 10 seconds has elapsed
- High voltage is available, as confirmed by presence of +12 V to the keying circuit

Connect the amplifier to a dummy load through an accurate power meter capable of indicating 1500 W full scale. Key the amplifier and check the idling plate current. With 3200-V plate voltage, it should be in the vicinity of 150 mA. Now, apply a small amount of drive and adjust the input tuning for maximum grid current. Adjust the output tuning until you see an indication of RF output. Increase drive and adjust the output coupling and tuning for the desired output. Do not overcouple the output: once desired output is reached, do not increase loading.

When you shut down the amplifier, leave the blower running for at least three minutes after you turn off the filament voltage. I found the 3CX1200Z7 to be an excellent tube. I tried it with exessive drive, plate-current saturation, excessive plate dissipation—all the abuse it's likely to encounter in amateur applications. I had no problems, but I don't recommend you repeat these tests!

Table 3
Power Supply Specifications

High voltage: 3200 V
Continuous current: 1.2 A
Intermittent current: 2 A
Step/Start delay: 2 seconds

A Companion Power Supply

A good, solid-state high-voltage power supply is a necessity to ensure linearity in SSB operation. Specifications of the power supply I built are given in Table 3. Figure 8 is a schematic diagram of the supply. A power supply for a high-power linear amplifier should operate from a 240-V circuit, for best line regulation. I have specified a special, hydraulic/magnetic circuit breaker that doubles as the main power switch. I don't recommend you substitute a regular switch and fuses for this breaker, as fuses won't operate quickly enough to protect the amplifier in case of an operating abnormality. The bleeder resistor dissipates about 100 W, so I included a small fan to remove the excess heat.

Power Supply Construction

The power supply can be built into a 17×13×10-inch cabinet. The power transformer is quite heavy, so use ⅛-inch aluminum for the cabinet bottom, and reinforce it with aluminum angle for extra strength. The diode bridge consists of four legs, each containing five diodes.

Power Supply Operation

When the front-panel breaker is turned on, the two, 25-Ω resistors in the primary circuit limit inrush current as the filter capacitor charges. After two seconds, K1 activates, shorting both resistors and allowing full line voltage to be applied to the transformer.

As with all high-voltage power supplies, you must be extremely careful! Before opening the cabinet, remove the ac-line plug from its receptacle, and confirm that the filter capacitor is discharged before working on the supply.

Conclusion

This amplifier is a reliable and cost-effective way to generate a big 2-meter signal—almost as quickly as a solid-state amplifier.

To ensure that the output of my amplifier meets current spectral purity requirements, I use a high-power version of the half-wave output filter that appears as Figure 16 on page 39-10 of the 1993 and 1994 editions of *The ARRL Handbook*. Although I did not make spectral measurements of the output, I can run full output while my wife Mary Lou watches TV in a nearby room of our home.

Another suitable filter is the one that appears in the 1990 *ARRL Handbook* (Figure 150, on page 31-72) as part of the description of "A Legal-Limit 2-Meter Tetrode Amplifier."

Notes

[1] Suggested retail price of the 3CX1200Z7 is $625. You can obtain it from: Henry Radio, 2050 S Bundy Dr, Los Angeles, CA 90025, tel 310-820-1234; Richardson Electronics, 40 W 267 Keflinger Rd, La Fox, IL 60147, tel 708-208-2200; RF Parts, 435 South Pacific St, San Marcos, CA 92069, tel 619-744-0700.

[2] W. Orr, Editor, *Radio Handbook*, 23rd ed. (Indianapolis: Howard W. Sams and Co., 1987), pp 18-2 through 18-7.

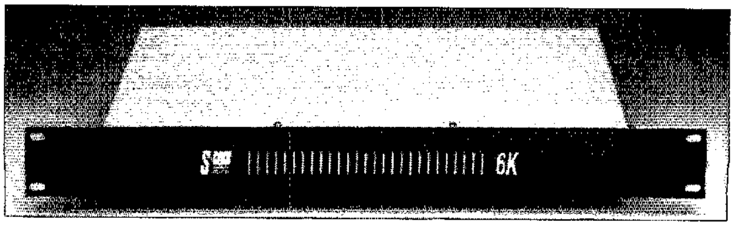
[3] This project is also described in the *Radio Handbook*, pp 18-11 through 18-15. QST

New Products

NEW SOFTWARE FOR S-COM 6K REPEATER CONTROLLERS

◊ S-COM Industries is now shipping a software upgrade with all of its 6K Repeater Controllers. New features include: a 100-setpoint scheduler for handling automatic events; an "antikerchunker"; improved DTMF routines compatible with the Long-Tone Zero (LiTZ) standard; the ability to monitor repeater audio from a support telephone line; more macros; and the ability to use run-time variables (for automatic CW time and date stamping in repeater messages).

The 6K Repeater Controller with Autopatch is priced at $395. Older 6K controllers can be upgraded to version 2.0 software with the 6K V2.0 upgrade kit, priced at $49.95 plus $3 s/h. For more information, contact S-COM Industries, PO Box 1718, Loveland, CO 80539-1718; tel 303-663-6000. QST



Key Components of Modern Receiver Design: A Second Look

More about front-end-filter switching diodes, dynamic-range testing, and why commercial transceivers cost more than the best high-end Amateur Radio gear.

By Ulrich L. Rohde, KA2WEU
52 Hillcrest Dr
Upper Saddle River, NJ 07458

This three-part series (May, June and July 1994 *QST*[1]) has probably had the best response of anything I have ever published in *QST*, which I attribute mostly to the fact that the community of ham-equipment users now is more educated and looks more critically at Amateur Radio products. The higher power densities caused by the large number of active hams on our MF and HF bands have made us more alert to these large-signal effects.

Diode Questions

Among the detailed responses and questions triggered by the series, front-end switching diodes were the Number One topic. Although Hewlett-Packard is a well-established company, most individuals encountered difficulty in finding the HP 5082-3081 diodes mentioned in the series. The majority of distributors either didn't carry them or considered them exotic. During the time I was evaluating these diodes, I was contacted by Tom Thompson, W0IVJ, who has since done extensive measurements with various diodes and has come up with some fascinating results. His findings are the subject of an independent article.[2] The immediate outcome of his efforts revealed that the Motorola MPN3700, an RF-switching PIN diode that comes in leaded and surface-mount packages, seems to work as well as the Siemens and HP PINs at a cost of less than $1 per diode.

Even though the MPN3700 costs significantly less than the HP 5082-3081, replacing all of a transceiver front end's 20-plus filter-switching diodes (usually two per filter—one at each filter's input and another at its output) can be expensive. Considering the likelihood that only the filters' *input* diodes cause the second-order IMD spotlighted in the "Key Components" series, several questioners asked which diodes must be replaced. Replacing only the filter-input diodes might make enough improvement, but an improvement of about 6 dB in *third-order* IMD dynamic range is also involved. Because all of diodes in the signal path likely play a part in setting third-order-IMD performance, I recommend changing *all* the diodes between the antenna and the first mixer, which includes the diodes on both sides of the bandpass filters of the transceiver of interest. (Special note for transceivers that use diode TR switching: It is generally *dangerous* to replace the transmit/receive switching diodes, which typically are already high-quality PIN types.)

Communication with Japanese transceiver manufacturers suggests that they may prefer domestic devices to the Siemens diodes I used for my tests, and to the Hewlett-Packard and Motorola types North American amateurs can use to replace the non-PIN switching diodes in their equipment. Of the Japanese diode types I've unearthed so far, the MI204 seems to duplicate the specifications of the Siemens diodes I used (the leaded version of which is the BAR17), but I understand that the manufacturers may consider it to be too expensive to be a general replacement for front-end-filter switching diodes. The 1SV128 may be a tolerable cost-versus-performance compromise between the "bad" diodes we have seen in the past and the MI204 if its minority-carrier lifetime is long enough.[3] I'm interested in hearing from radio amateurs who have on-air or lab-testing experience with equipment that uses 1SV128s as front-end-filter switches.

Antenna Tuners

A similar set of questions arose relative to our transceivers' antenna tuners. Although some of the "older" radios like the ICOM IC-781 and -765, and those with external antenna tuners, like the Kenwood TS-50S, keep the antenna tuner in line in both transmit and receive, others, such as the Kenwood TS-450S, Yaesu FT-890 and similar ones, must be rewired to use their tuners in receive and transmit. Some comments on how to accomplish this for the Kenwood TS-450S were written up by David DeCoons, KE2SL, in July 1994 Hints and Kinks, the same issue in which Part 3 of the "Key Components" series appeared.

Doing the Radio Modifications

The modifications were done for me by either the manufacturer as indicated, and verified, or by Amateur Electronic Supply (AES) Service, 5710 W Good Hope Rd, Milwaukee, Wisconsin 53223, telephone 414-358-0333, fax 414-358-3337. For the Kenwood transceivers tested, Kenwood made the actual two-tone measurements and confirmed my measurements as well.

Careful readers may have noted that the Ten-Tec Omni VI data came in at the low end of the second-order-IMD performance curve delimited by the measurements presented.[4] As we know it in the US market, the Omni VI uses PIN-diode TR switching—two MPN3700s and a 1N4007 rectifier diode, which alone in the 1N4000 series contains a PIN structure—and BA482 front-end-filter switching diodes. I had been contacted by one reader who ultimately was told by Ten-Tec that a few "European high-performance version" units have been using a relay or relays instead of PIN diodes. A relay is an improvement over even the best PIN diodes, of course, but I am convinced that the Omni VI can be significantly improved merely by installing appropriate RF PIN diodes.[5]

Other comments had to do with the prospect of exchanging front-end switching diodes with hot-carrier diodes. This would be a mistake! Hot-carrier diodes are designed to replace the germanium point-contact diodes long used for RF and microwave mixing, detection and frequency multiplication. Although they exhibit very low threshold voltages, they are very temperature-sensitive and do not handle static electricity well. A hot-carrier diode is a far better candidate for use as mixer than as a low-distortion switch, and its tendency to support intermodulation can only be calmed down by subjecting it to fairly high bias currents. Hot-carrier diodes are definitely the wrong choice for low-IMD *switching* diodes.

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[1] Notes appear on page 44.

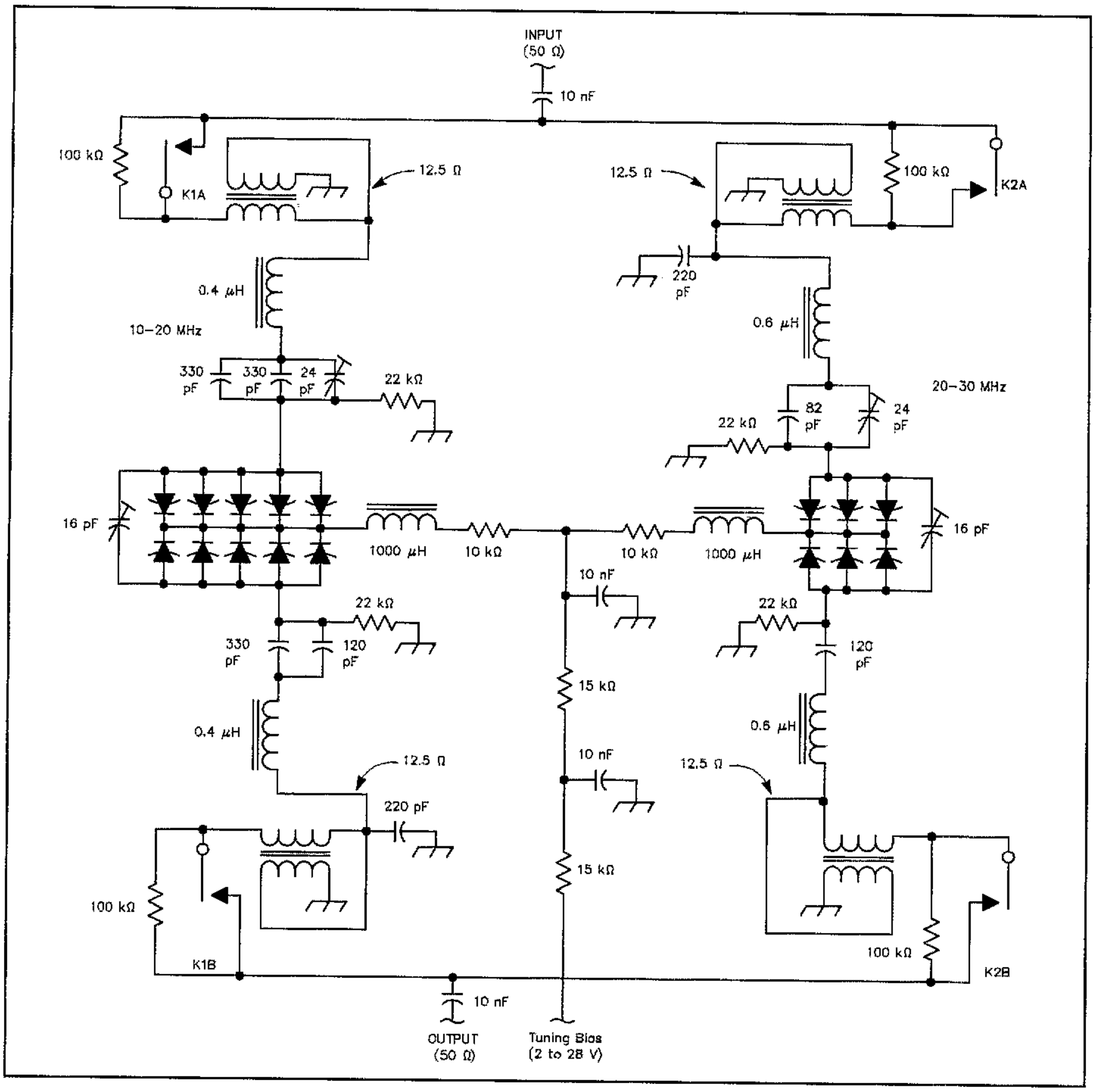


Figure 1—A voltage-controlled input filter stage that covers 10 to 30 MHz in two ranges. Energizing K1A selects the 10 to 20-MHz filter; energizing K2 selects the 20 to 30-MHz filter. The filter's impedance is 12.5 Ω, with 4:1 transformers matching them to the module's 50-Ω I/O impedance. The tuning diodes are MVAM125s, each of which consists of a common-cathode diode pair. This circuit comes from the Rohde & Schwarz ESH-2 receiver, the specifications for which guarantee a +30-dBm intercept point with this filter in line.

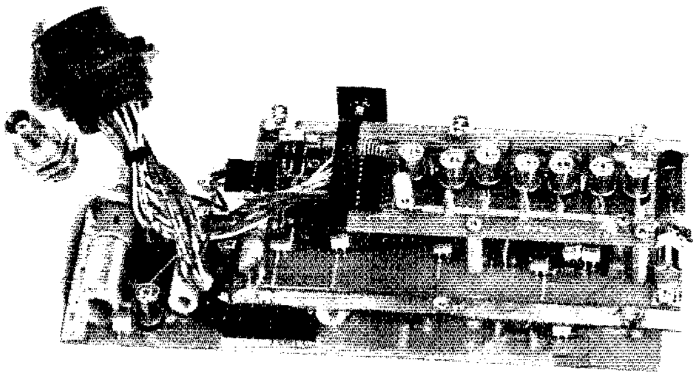


Figure 2—Inductor-capacitor assembly, including the post-filter amplifier. The cans are TO-5 relays; they switch the inductors and capacitors.

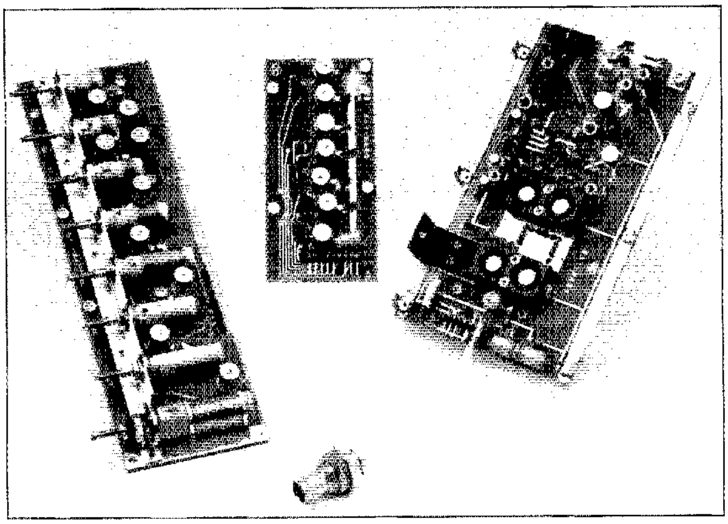


Figure 3—The component boards of the assembled filter shown in Figure 2.

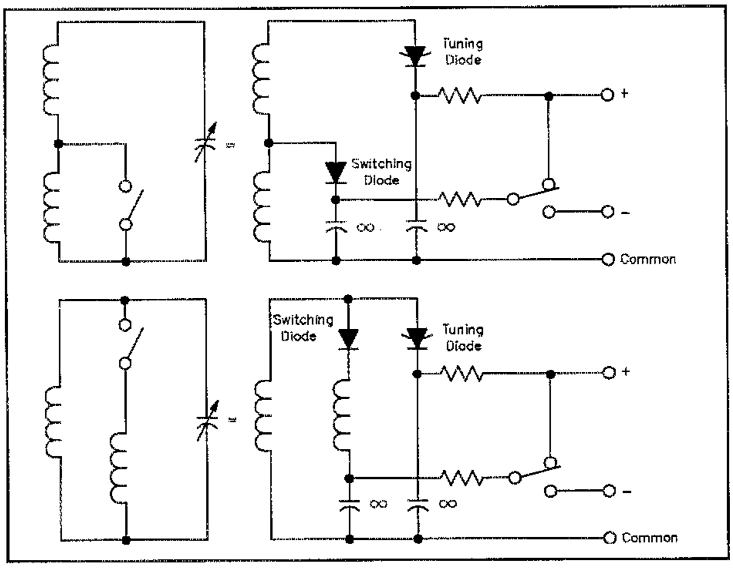


Figure 4—Mechanically tuned circuits and their electronically tuned equivalents. The infinity symbol (∞) near some of the capacitors indicates that these components act as RF short circuits. Although it's possible to turn off a switching diode merely by removing its forward bias, using reverse bias is preferable because it lessens the likelihood that signals in the tuned circuit can turn the diode on.

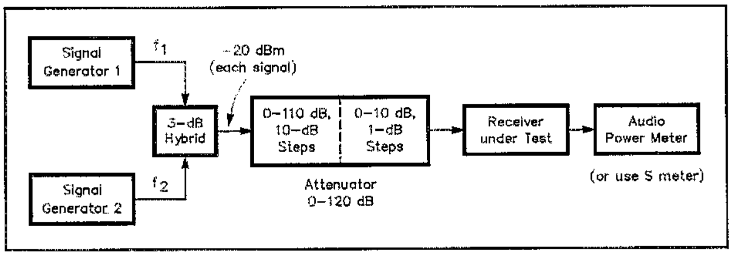


Figure 5—The intermodulation test setup as described in July 1994 *QEX*.

Input Filtering

The irony of the second-order IMD discussed in the "Key Components" series is that it occurs in the diodes used to switch front-end filters built into our radios for the express purpose of suppressing second-order IMD! Good front-end filtering, teamed with balanced RF amplifier and mixer circuitry to cancel out second-order responses, is the ultimate guarantor of high second-order IMD dynamic range.

Tuning-Diode Filter Tuning

In our discussions, Tom Thompson and I looked at the issue of input filtering. Our radios use switched fixed filters, but they are not the only solution. Figure 1 presents an implementation that uses tuning diodes to provide continuously tunable input filtering, like the preselectors available in the old tube radios and early dual-gate MOSFET radios. This circuit, which covers 10 to 30 MHz in two ranges, comes from the Rohde & Schwarz ESH-2 test receiver, which guarantees a +30-dBm third-order intercept point with this filter in line. The method used here is consistent with my approach to VCOs, in which I distribute the tank-circuit current through many diodes to improve linearity. In addition, the tuned circuits are designed to exhibit a higher-than-usual ratio of capacitance to inductance, which lowers their impedance and keeps the RF voltage across the diodes low. In the case of the ESH-2 input filter, using many more diodes results in a much higher intercept point. In an oscillator, using multiple tuning diodes results in better noise performance.

Binarily Stepped Filter Tuning

An alternative to this approach is the use of a PIN-diode-switched input filter, which is similar to an oscillator in which range switching is accomplished by switching fixed inductor and capacitors in binary values. (For example—the stray inductances and capacitances present in any real implementation aside for the moment—seven switches and seven capacitors [1, 2, 4, 8, 16, 32 and 64 pF] could be used to synthesize a 127-pF variable capacitor with 1-pF resolution.) Any resistance in series or parallel with a tuned circuit's inductances and capacitances lowers its Q and increases its loss, so low-loss diodes are a must in this application. (The MPN3700 and its equivalents are suitable.) Using relays instead of diodes would minimize losses and allow the filter to be used in transmit.

Figure 2 shows the inductor and capacitor assembly for just such a binarily tuned filter, and Figure 3 shows its mechanical assembly. A post-filter amplifier in this module compensates for the filter's insertion loss. To minimize second-order intermodulation, the amplifier consists of two devices, one PNP and one NPN, operating in complementary-symmetry, and the input stage acts as a single-tuned circuit, maintaining high operating Q.

The antenna line connects to the module at the center of the top edge of the rightmost module in Figure 3. A pre-ionized gas-discharge cell (the small half-white, half-silver cylinder at far right in Figure 2) provides lightning protection for the filter input. This device, which consists of a mixture of radioactive gases whose radiation level is significantly below critical level toxic to humans, clamps at voltage levels as low as 10 to 20 volts or more of RF and can handle several thousand amperes for short periods. Such devices can also be used to protect a receiver input against high-power transmitter signals. More detail on these tracking input stages appeared in November 1992 *QST*.[6]

Figure 4 compares means of tuning circuits mechanically and electrically, and shows the process whereby inductors can be diode-switched in series or parallel to extend the frequency range. Capacitors can also be switched using these techniques.

Measuring Intermodulation

The actual test setup for measuring intermodulation, Figure 5, was described in *QEX*,[7] but for those of you who don't regularly subscribe to *QEX*, here is a brief summary. Two signal generators, both of which must have little second-harmonic output, are required. Most modern signal generators have low-pass filters at the output of their amplifiers, and suppress their second harmonic by at least 40 (typically 60) dB.

The next item needed for dynamic-range testing is a hybrid 3-dB coupler. These couplers have been described in a variety of ARRL publications and are available commercially.

The remaining item is an attenuator, which takes the combined input from both signal generators and feeds it to the device under test. (One of the absolute requirements *before* trying to make any of these tests is to remove the microphone from the transceiver and make sure that it cannot transmit unless you hit its **MOX** or **TRANSMIT** button.)

The accuracy of these measurements depends largely on the fact that both signal generators can be set precisely at −17 dBm and that the 3-dB coupler indeed introduces 3 dB of loss. The best way to double-check is to confirm by measurement that the power of each single tone at the hybrid coupler's output is −20 dBm.[8]

The attenuator must be capable of 1-dB and 10-dB steps and an attenuation range of at least 100 dB. One from Hewlett-Packard or similar high-quality manufacturer would be ideal. Now that most modern attenuators are remotely controllable via a data bus, some manually controlled attenuators have become obsolete and can sometimes be bought at affordable prices.

What About the XK2100 High-Performance Transceiver?

Some correspondents asked about the Rohde & Schwarz XK2100 transceiver described in Part 3 of the series (Figure 6). One reader suggested that the ARRL staff should formally review it. Although this $18,000 box is not really intended for general ham use, I brought the unit up to ARRL HQ, and listened to it with some of the key players after connecting it to a multiband beam antenna (a DJ2UT 40 to 10-meter type) on the roof. W1AW, the Maxim Memorial Station, was on the air for part of the test period. Not being able to duplicate the multiple-strong-signals W1AW environment at home, I was quite pleased with the transceiver's performance with W1AW on the air, especially since the version we listened to does not have a tracking preselector (meaning that, in receive, the only input filtering between the

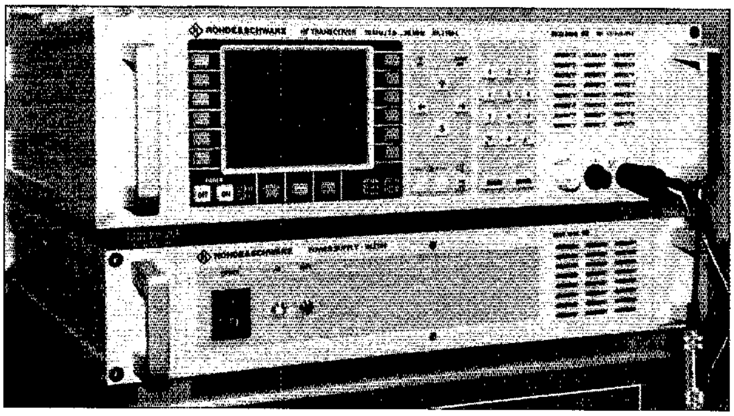


Figure 6—The Rohde & Schwarz XK2100 transceiver and power supply. The title-page graphics of Part 3 of the "Key Components" series show its liquid-crystal display in greater detail.

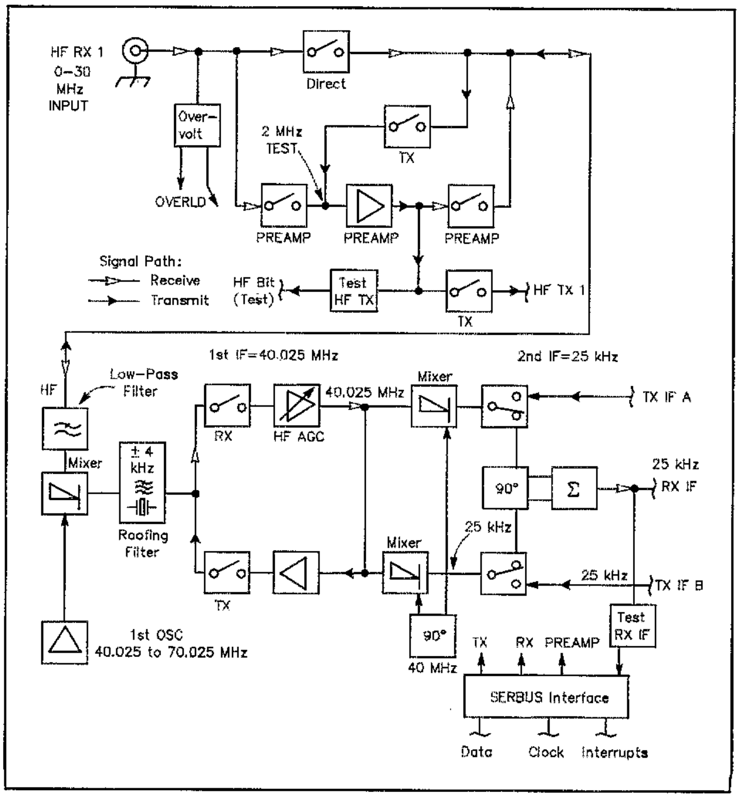


Figure 7—The XK2100 first upconverts incoming signals to 40.025 MHz. This first IF is then downconverted to 25 kHz in an image-reject mixer.

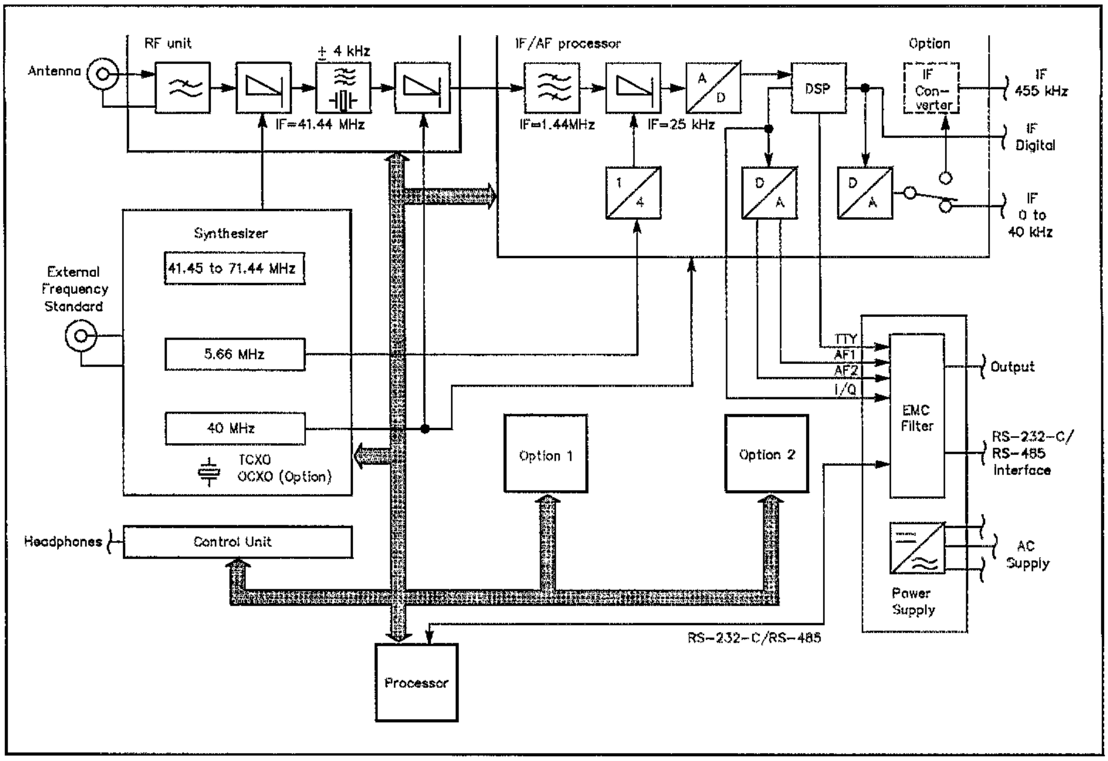


Figure 8—Partial block diagram of the "DSPized" EK890 receiver. In this arrangement, the oscillator that was the EK890 BFO now converts the radio's second IF (1.44 MHz) down to a new third IF (25 kHz) for digital signal processing.

transceiver's antenna jack and first receive mixer is a 32-MHz low-pass filter). With the XK2100's optional preamplifier switched out, we could hear weak signals quite close to W1AW. For highly demanding applications, such as multioperator contest operation, the input selectivity added by the XK2100's optional tracking input filter would be a must.

Additionally, we also tuned directly to W1AW's signal, which produced something like 0.5 volt at 50 Ω (114 dBμV) at the transceiver input. This was also a fairly harsh test, especially for the XK2100's AGC. We found the radio's AGC response was entirely in keeping with what hams expect of well-designed, high-end analog Amateur Radio gear—a significant finding, since the marriage of analog and digital technology can make the achievement of good AGC performance much more difficult than it is in an all-analog design.

The XK2100's IF section uses digital signal processing. As Figure 7 shows, the XK2100 first upconverts incoming signals to 40.025 MHz. An image-reject mixer downconverts this IF to 25 kHz for DSP.

Some readers are curious about how DSP might be added to an existing receiver design. Figure 8 shows the block diagram of such a scheme: the Rohde & Schwarz EK890, the BFO of which can be used to downconvert the radio's second IF to 25 kHz for DSP. The 25-kHz digital processor replaces the EK890's IF board with all of its discrete IF filters. This approach has the potential of making a radio smaller and more affordable, in addition to allowing a greater selection of operating modes and bandwidths.

I have also been asked to justify the XK2100's $18,000 price tag, which, of course, is extremely expensive compared to standard ham equipment—practically ten times more costly! For starters, commercial transceivers like the XK2100 are typically designed to meet very stringent environmental specifications, including temperature and vibration, and are also required to use components that are much more reliable than those in standard ham radios. Because commercial/military transceivers must be designed to meet government specifications, development costs are fairly high. Added to this is the fact that the number of units sold is fairly small compared to the equipment generally used by hams. Nonrecurring engineering costs are therefore fairly high for equipment like the XK2100.

High-end Amateur Radio transceivers such as the Kenwood TS-950SDX, Yaesu FT-1000D and ICOM IC-781 are really excellent products, but they overwhelm commercial users with their overabundant features. Compared to high-end ham gear, the XK2100 system affords a drastically reduced number of bells and whistles. On the other hand, if you take five different ham transceivers and evaluate their signal-strength meters, all will give different readings compared to an absolute scale. (The same is also true for a normal production run of a given ham equipment model.) In contrast to this, high-end commercial receivers and transceivers operate from −10 dBμV to +120 dBμV, both in AGC range and AGC indication.

Antennas

Finally, since I have been asked several times about the antenna system I use in my evaluations. At Upper Saddle River, the antenna consists of several levels, in descending order: The top level is a two-element beam for 40 meters (KLM); the second is a Force 12 antenna for 17 and 12 meters (three elements on each band); and the third is a Force 12 antenna for 20, 15 and 10 meters (three elements on 20, three elements on 15 and four elements on 10). The 40-meter beam antenna is approximately 90 feet above sloping ground.

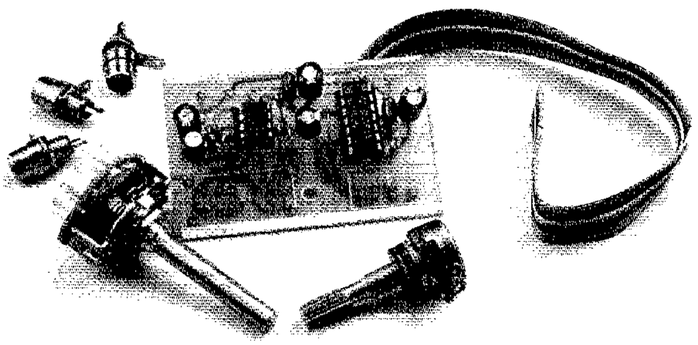
At my Marco Island, Florida, QTH, a

(continued on page 44)

A Simple, General-Purpose AF Amplifier

You're sure to find a million uses for this handy 1-watt audio amplifier. (Well...at least a few...)

By Ben Spencer, G4YNM
33 New King St
Bath, Avon BA1 2BL
England



This economical, general-purpose AF amplifier features a low-noise, high-input-impedance FET op amp input stage, with a switched input-gain control and an audio power amplifier with continuously variable gain control. All components are readily available, including a PC board.[1] Chances are, you can find most of the components in your junk box.

Circuit Description

Refer to Figure 1. C4 and C5 bypass the power-supply bus. Op amp U1 is a TL081 low-noise FET device with a 50-kΩ input impedance and initial gain controlled via S1 (**GAIN 1**). The voltage gain settings are 0, 20 and 40 dB for switch positions 1, 2 and 3, respectively. The op amp's lower cut-off frequency is set at 16 Hz by R3 and C2. In the 0-dB position, R3 and C2 have no effect, as the op amp is then operating as a unity gain buffer.

U1's output is fed via C3 to **GAIN 2** control R7 and then to the 34-dB audio power amplifier U2.[2] As the power-amplifier gain is fixed at 34 dB, the output level is further controlled by adjusting the input level via R7. Using S1 and R7, it's possible to obtain any gain value up to 74 dB—a considerable range. The AF output can be fed to an 8-Ω loudspeaker or other 8-Ω load. If you use a resistor, ensure it's rated at 2 watts. A LINE output (approximately 600 Ω) is provided via R6.[3] U2's output is short-circuit-proof and has internal thermal limiting.

Construction

Using the single-sided PC board designed for this project (see Note 1), construction is easy. With the exception of the loudspeaker, gain controls and input and output connectors, all components are mounted on the board. If you intend to run a sine-wave input into the amplifier for a prolonged period, attach a heat sink to U2.

To make the amplifier less susceptible to external signals (such as the RF from your transmitter), mount it in a metal case or an enclosure made of double-sided PC-board material. Use shielded wire (RG-174 coax or microphone cable) for the input and output wiring. Keep the input and output leads of the circuit at opposite ends of the enclosure to eliminate the possibility of crosstalk and positive feedback, which would cause unwanted oscillation.

You can mount a loudspeaker within the enclosure, or merely route the LS output line to a suitable jack. Or, route the LS line through a normally closed output jack that disconnects an internal speaker when an external load is plugged into the jack. That way, you'll always have a speaker available and have the option of using an external speaker (or other load) when you want to.

Test and calibration

Ideally, an AF generator[4] should be used to set the input level; however, a simpler, empirical method is to connect the amplifier's input to the audio output of an existing piece of equipment such as a radio receiver, cassette player, or similar source. Set S1 to the 0-dB position, the variable **GAIN 2** control (R7) to midrange, and adjust the audio source's volume control to an acceptable level. Adjust R7 and check to see that the audio level changes correctly.

Table 1

Required audio-level input for 1-W output into an 8-Ω load for various settings of S1.

S1 Setting	*Input Level (V_{RMS})*	*Overall Gain (Decibels)**
0 dB	0.0566	+34
20 dB	0.0057	+54
40 dB	0.0006	+74

*R7 set to maximum gain (34 dB)

Next, set R7 for a quiet level and without adjusting the audio source's volume control, set S1 to the 20-dB position—there should be a significant increase in the audio output level. Repeat the test for the 40-dB position. That's all there is to it!

Summary

Whether at the workbench, in the shack or elsewhere around the home, this amplifier is certain to come in handy.[5] It's inexpensive, easy to build and useful—what more could you ask for?

Notes

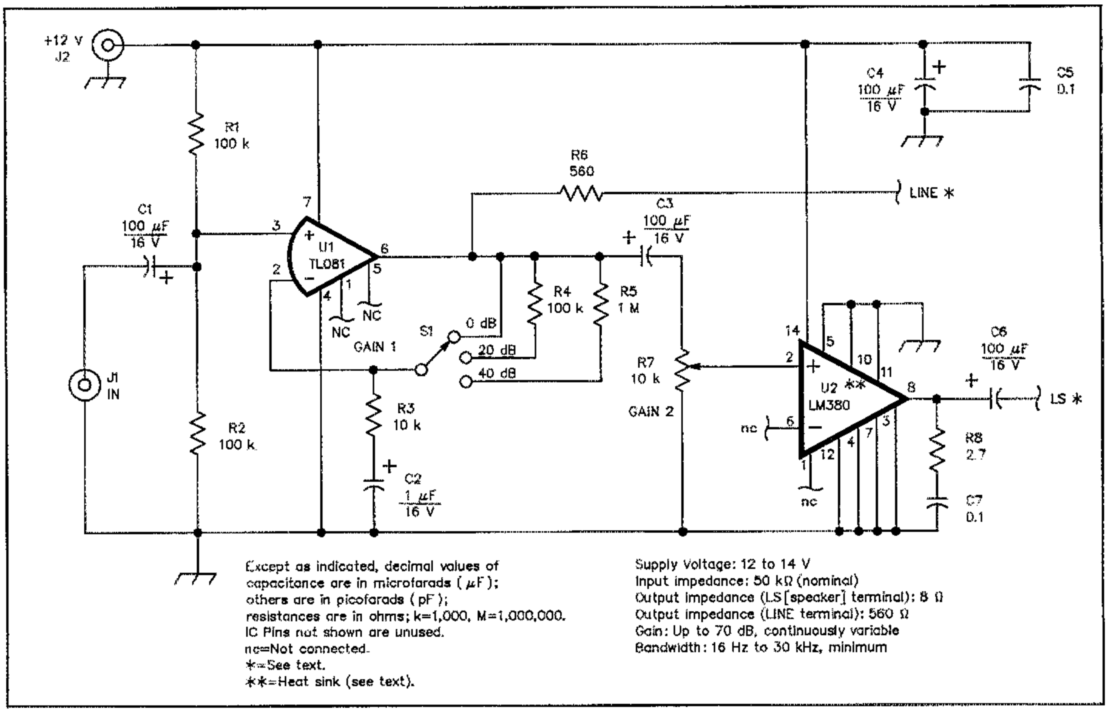
[1] PC boards are available from FAR Circuits, 18N640 Field Ct, Dundee, IL 60118-9269. Price: $4 plus $1.50 shipping.

[2] The LM380 power amplifier does have one nasty trait: On the negative-going half cycle of the output waveform, the IC can produce a spurious oscillation as current charges into it. The oscillation frequency is between 5 and 10 MHz. R8 and C7 are used to suppress this oscillation. This behavior is endemic to the device and not a function of this particular project.

[3] A low dc voltage appears at the LINE output. Use an external 100-µF, 16-V capacitor, if necessary, to dc-isolate the load.—*Ed.*

[4] See B. Spencer, "A Function Generator with a Frequency Counter Digital Readout," *QST*, Apr 1994, pp 35-39; see also Feedback on p 85 of this issue.

[5] This amplifier may be just what you need to build into some other project of yours that needs a bit of audio punch. By hard-wiring the S1 connection and using a *fixed-value* resistor for R7, you can preset the amplifier gain and not be concerned with variable controls.—*Ed.*



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.
IC Pins not shown are unused.
nc=Not connected.
*=See text.
**=Heat sink (see text).

Supply Voltage: 12 to 14 V
Input impedance: 50 kΩ (nominal)
Output impedance (LS [speaker] terminal): 8 Ω
Output impedance (LINE terminal): 560 Ω
Gain: Up to 70 dB, continuously variable
Bandwidth: 16 Hz to 30 kHz, minimum

Figure 1—Schematic of the general-purpose AF amplifier. All resistors are 1/4-W, 5%-tolerance carbon-composition or metal-film units. Equivalent parts can be substituted. General-purpose IC replacements are shown in parentheses.

C1, C3, C4, C6—100 µF, 16 V radial electrolytic
C2—1 µF, 16 V radial electrolytic
C5, C7—0.1 µF disc ceramic
J1, J2—Phono jack
R7—10-kΩ, log taper, PC-mount potentiometer
S1—Three-position slide or rotary switch
U1—TL081 low-noise FET op amp (NTE857)
U2—LM380 audio power amplifier (NTE740A)

Note: NTE general-purpose replacement components are available from many suppliers including Ocean State Electronics, PO Box 1458, 6 Industrial Dr, Westerly, RI 02891; tel: 800-866-6626 and 401-596-3080; fax 401-596-3590. See the ARRL Parts Suppliers List in the 1995 edition of the *Handbook.* (Don't have a recent *Handbook?* Get a *new* one by simply calling Publication Sales at 203-666-1541, tell 'em what you want and put it on your charge card!—*Ed.)*

Misc: PC board (see Note 1), shielded audio cable (RG-174 coax can be used), enclosure, knobs, output connectors or jacks, hook-up wire, IC heat sink, loudspeaker.

Key Components of Modern Receiver Design

(continued from page 42)

penthouse atop an approximately 120-foot-tall building. I have erected a 40-foot tower atop the building. The tower supports a DJ2UT antenna (40 through 10 meters) at approximately 160 feet above sea level. The antenna can be fully nested in case of storms (not *if* they occur, but *when!*) and is supported sideways by the elevator shaft penthouse.

At both locations, a horizontal V antenna (55 ft per leg), matched by an SGC automatic tuner, serves as backup.

Conclusion

Reader response to the "Key Components" series has shown me that radio amateurs continue to show high interest in understanding and improving the performance of their already excellent radio equipment. There's still much that manufacturers of shortwave transceivers and receivers can do to cater to the needs of this alert and demanding audience.

Notes

[1] U. L. Rohde, "Key Components of Modern Receiver Design," Part 1, *QST*, May 1994, pp 29-32; Part 2, Jun 1994, pp 27-31; Part 3, Jul 1994, pp 42-45. The inductor labeled 9.2 nH in Figure 16 (Part 2) should be 92 nH.

[2] Tom Thompson's "Exploring Intermodulation Distortion in RF Switching and Tuning Diodes" starts on page 25 of this issue.—*Ed.*

[3] Of the PIN diodes I have considered so far, the Siemens and MI204 devices have the longest minority carrier lifetime and are therefore the components of choice for this application. The HP 5082-3081 (which costs more than the MI204) comes next. The MPN3700 (the least expensive of the group) has the shortest minority carrier lifetime of the group. So far, however, Tom Thompson's measurements indicate that all of these devices produce acceptably low IMD at HF and above.

[4] Dr Rohde's original manuscript did not include test data for a Ten-Tec product. Seeking rounded coverage of contest-grade transceivers from all the manufacturers who currently make them, we tested W1AW's Omni VI in the ARRL Lab and included the results with Dr Rohde's assent.—*Ed.*

[5] In an October 28, 1994, circular letter sent to all registered Omni VI owners, Ten-Tec reported that the Omni VI's second-order intercept point can be increased from +40 dBm to over +60 dBm by changing the value of two resistors on the transceiver's BPF/Front End board. An improvement of this magnitude brings the Omni VI's second-order intercept into line with its already excellent third-order IMD performance, and likely makes replacing its front-end diodes desirable only in extreme cases of second-order IMD.—*Ed.*

[6] U. L. Rohde, "Recent Advances in Shortwave Receiver Design," *QST*, Nov 1992, pp 45-55.

[7] U. L. Rohde, "Testing and Calculating Intermodulation Distortion in Receivers," *QEX*, Jul 1994, pp 3-4.

[8] Measuring the coupler's output with both tones switched on simultaneously can be misleading, depending on the measurement technique used. Using an RF microvoltmeter, I have measured something like −16.7 dB—a measurement 0.3 dB lower than the expected −17 dBm—because the phase sensitivity of its sensing element causes it not to sum the powers of multiple signals perfectly. A true-power meter, such as one using a thermally based sensing element, should indicate −17 dBm with two −20-dBm tones present.

arrestor *must* be used.

Whether you use balanced line or coax arrestors, they should be mounted at the entry point to your shack—on the *outside* of the building—using a secure grounding connection. The easiest way to do this is to install a large metal (preferably copper) enclosure as a bulkhead and grounding block. This bulkhead serves as your last line of defense by keeping the lightning energy from entering your home, so it's critical that it be installed properly. You can homebrew a bulkhead panel from ⅛-inch copper sheet, bent into a box shape. Position the bulkhead on the building exterior, 4 to 6 inches (minimum) away from nearby combustible materials. Install a separate ground rod for this panel and connect it to the bulkhead with a short, direct connection. Also, bond this ground rod to the rest of the ground system. Mount all protective devices, switches and relay disconnects on the outside wall of the bulkhead.

Q: What about my rotator control lines?

A: You'll find multi-wire surge suppressers for these and other wires leading to your antenna system. They are available from a number of manufacturers, including ICE and PolyPhaser. As with the feed-line arrestors, these should be mounted on the grounded entrance panel.

Q: Can I do anything else to keep a strike from reaching my equipment?

A: The only foolproof way to protect your station equipment is to completely disconnect each item from the antenna system and the wall outlets when you're not operating. Even when you're on the air, however, you could encounter the rare "bolt from the blue."

Let's admit it; we sometimes forget to disconnect our station before we go off to our other activities. Don't let the lack of a thunderstorm forecast lull you into a sense of false security. Although lightning usually occurs during summer thunderstorms, it has also been known to strike during winter storms—and even when there are no storms at all. The strike that comes before the first clap of thunder may be the one that hits your station.

Q: Is there an easy way to disconnect my setup?

A: A quick, convenient way to disconnect your feed lines is through a feed-line switch. If you're using a coaxial feed line, you can use a manual, multiposition coax switch, a remote coax switch or an in-line coaxial relay.

Although you could also simply disconnect your coax by hand, it's awkward and the connectors will become worn after repeated connecting and disconnecting. Some coax switches also contain lightning arrestors.

For open wire or ladder line, you could install a knife switch or electrically operated remote 2-pole relay. Whichever method you choose, be sure to mount the switch on the *outside* of your entrance panel.

Q: Is there anything else I should do to protect my equipment?

A: One area often neglected is power line protection. Inexpensive multioutlet strips usually have little or no protection against surges or transients. Sensitive electronic equipment (modern digital radios, TNCs, computers, etc) sometimes need more protection than the factory provides.

Power line protectors use several different protection schemes, each of which solves a different power-line problem. *Inrush current limiters* keep the input current to the equipment's power supply from exceeding a fixed level. *Transient suppressors* (usually semiconductor-type devices) absorb voltage spikes that could damage sensitive digital ICs. *Surge suppressors* limit the input voltage on the line (usually by a clamping or "crowbar" method) to prevent damage.

To protect your equipment against transients caused by lightning-induced voltage surges on the ac line, unplug the power strip at the wall socket. Don't depend on the built-in switch or wall-outlet switch. A nearby strike can induce voltages that will easily jump the gap and overload the protective circuits.

Q: Now what about that list of manufacturers?

A: As promised, here they are:

Alpha Delta Communications
PO Box 620
Manchester, KY 40962
tel 606-598-2029
Note: Coax lightning arrestors, coax switches with surge protectors

Ameritron
921 Louisville Road
Starkville, MS 39759
tel 601-323-8211
Note: Remote coax switches, inrush ac current protector

Certified Quality (The Wireman)
261 Pittman Rd
Landrum, SC 29356
tel 800-727-9473 (orders)
803-895-4195 (Tech line)
Note: The Wireman stocks copper wire up to #4 AWG, 2-inch flat copper strap, 8-foot copper-clad ground rod and 1×¼ inch bus bar.

Cushcraft Corporation
48 Perimeter Rd
Manchester, NH 03108
tel 603-627-7887
Note: Coax lightning arrestors

Industrial Communication Engineers, Ltd.
PO Box 18495
Indianapolis, IN 46218-0495
tel 317-545-5412
fax 317-545-9645
Note: Coax lightning arrestors

Lightning and Noise Protectors
PO Box 380054
Birmingham, AL 35238-0054
tel 800-776-8357

MFJ Enterprises
Box 494
Mississippi State, MS 39672
Note: Model 1704 coax switch with lightning arrestor

PolyPhaser Corporation
PO Box 9000
Minden, NV 89423-9000
tel 702-782-2511
Note: Many lightning protection products for feed lines, towers, equipment, etc

Radioware Corporation
87 Belmont Street
North Andover, MA 01845
tel 800-950-9273
Note: Amateur products distributor; Radioware stocks grounding blocks,other grounding products and many ICE products

Rohn
PO Box 2000
Peoria, IL 61656
tel: 309-697-5612
Note: Copper strap and other tower grounding products

Zero Surge Inc
944 State Route 12
Frenchtown, NJ 08825
tel 908-996-7700
Note: Power line surge protector

Thanks to ARRL Technical Advisor John Bittinger, KI7GW, for his assistance during the preparation of this column.

We welcome your suggestions for topics to be discussed in *Lab Notes*, but we are unable to answer individual questions. Please send your comments or suggestions to: Lab Notes, ARRL, 225 Main St, Newington, CT 06111. QST