$1.75

QEX

ARRL Experimenters' Exchange

DECEMBER 1992

EX: The ARRL
xperimenters' Exchange
nerican Radio Relay League
25 Main Street
ewington, CT USA 06111

Non-Profit Org.
US Postage
PAID
Hartford, CT
Permit No. 2929

QEX (ISSN: 0886-8093) is published monthly by the American Radio Relay League, Newington, CT USA.

David Sumner, K1ZZ
Publisher

Jon Bloom, KE3Z
Editor

Lori Weinberg
Assistant Editor

Mark Forbes, KC9C
Geoffrey H. Krauss, WA2GFP
Harold Price, NK6K
Zack Lau, KH6CP
Contributing Editors

Production Department
Mark J. Wilson, AA2Z
Publications Manager

Michelle Bloom, WB1ENT
Production Supervisor

Sue Fagan
Graphic Design Supervisor

Dianna Roy
Senior Production Assistant

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

Offices
225 Main St, Newington, CT 06111 USA
Telephone: 203-666-1541
Telex: 650215-5052 MCI
FAX: 203-665-7531 (24 hour direct line)
Electronic Mail: MCI MAIL ID: 215-5052
Internet:arrl@mcimail.com

Subscription rate for 12 issues:
In the US by Third Class Mail: ARRL Member $12, nonmember $24;

US, Canada and Mexico by First Class Mail:
ARRL Member $25, nonmember $37;

Elsewhere by Airmail:
ARRL Member $48 nonmember $60.

QEX subscription orders, changes of address, and reports of missing or damaged copies may be marked: *QEX* Circulation.

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

About the Cover:

Another power supply (yawn)? Take a closer look: Dennis Connole's circuit presents some interesting ideas you may find useful for your next power supply design.

TABLE OF CONTENTS

COLUMNS

DECEMBER 1992 QEX ADVERTISING INDEX

THE AMERICAN RADIO RELAY LEAGUE

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

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

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

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

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

Telephone 203-666-1541 Telex: 650215-5052 MCI.
MCI MAIL (electronic mail system) ID: 215-5052
FAX: 203-665-7531 (24-hour direct line)

Canadian membership inquiries and correspondence should be directed to CRRL Headquarters, Box 7009, Station E, London, ON N5Y 4J9, tel 519-660-1200.

Officers

President: GEORGE S. WILSON III, W4OYI
1649 Griffith Ave, Owensboro, KY 42301

Executive Vice President: DAVID SUMNER, K1ZZ

Purpose of *QEX*:

1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters

2) document advanced technical work in the Amateur Radio field

3) support efforts to advance the state of the Amateur Radio art

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

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking...

A Farewell to "Gateway"

The observant reader will have noticed a change on the cover of *QEX* this month. *QEX* no longer includes "Gateway," the packet-radio newsletter. This change comes about as part of the ongoing review of our editorial direction, as promised in earlier editorials.

"Gateway" began life as a biweekly newsletter in August of 1984. It remained in that form until March, 1990, at which time it became a monthly column in *QEX*. During the eight-plus years of its life, "Gateway" served its readers well, providing timely news about amateur packet radio. So why are we dropping it? To answer that question requires a look at the original charter of "Gateway," as described in the premier, August 14, 1984 issue:

> We have called this newsletter *Gateway* because we hope that it will, like a gateway station, facilitate communications between amateurs interested in packet radio.... This will not be a technical newsletter; there are already several fine packet-radio newsletters covering technical issues. This will be a "news" newsletter.... Perhaps when there is a worldwide amateur packet-radio network there will be no need for packet-radio newsletters. Until then, we hope that *Gateway* informs and interests you.

To a large degree, "Gateway" is a victim of its own success. Since 1984, amateur packet radio has grown by leaps and bounds, to the point where it surely must be considered "mainstream" Amateur Radio. Its growth has allowed the dissemination of routine news via the worldwide packet-radio network predicted in that first issue of the *Gateway* newsletter. For those reasons, the need for a column such as "Gateway" in *QEX* has diminished. That's not to say that there is not a place for news about packet radio, merely that *QEX* is not the proper vehicle. The League recognized the "mainstreaming" of packet radio when it instituted the "Packet Perspective" column in *QST*, the ARRL membership journal, and that column will continue to keep you abreast of packet issues.

QEX is supposed to be about Amateur Radio experimentation and development, so we want our coverage of packet radio to concentrate on experimental/developmental issues. To that end, we are pleased to announce a new column, "Digital Communications." As the name implies, the subject matter of this column, which will run every other month, is intended to extend beyond packet radio to the entire field of communications transmitted by digital means. To conduct this column, we are fortunate to have obtained the services of one of the premier contributors to the development of amateur packet radio: Harold Price, NK6K. Harold has been active in packet radio since 1982. He was part of the team that developed the software for the TAPR TNC-1, the first production TNC, and he was (and is) a principal developer of software for amateur spacecraft, including most of the existing packet-radio satellites. A professional computer software consultant, Harold has also served on the ARRL Digital Committee, the TAPR Board of Directors and the AMSAT Board of Directors. We are proud and pleased to have him aboard.

As with all of our columns, "Digital Communications" will be most effective when you, the reader, participate. No monologue about digital communications—or any other technical subject—can be as informative as a dialogue. I urge you to share your thoughts with us.

This Month in QEX

Power supplies, especially those using the ubiquitous 723 regulator chip, have been published *ad infinitum*, but Dennis Connole shows us in "Twin-Bridge Power Supply" that there is still fresh ground to be plowed on the subject.

In "An Audio Peak Filter with Noise Reduction Effect for CW Reception," Yoshiharu Mita, JH1XEO, exploits the common-mode rejection of an op amp to differentiate noise from a signal using bandpass filters.

Gregory Glass, N2MOH, wants us to experiment with video image processing, and in "The Video Experi-menter" he shows us how easy it is to set up a video lab and describes some simple experiments to get us started.

This month, Zack Lau's "RF" column describes a 10-GHz preamp Zack designed and shows us some of the trials and tribulations of building and measuring 10-GHz equipment. Harold Price kicks off his "Digital Communications" column by looking at how HF packet evolved to its present state and how it could be improved using existing techniques.

—KE3Z, email: jbloom@arrl.org (Internet)

Twin-Bridge Power Supply

By Dennis C. Connole
14055 NW Cornell Rd
Portland, OR 97229-5405

Introduction

The power supply described below may be used as a constant voltage source or a constant current source. It also has the ability to shut itself down with a quick-response electronic circuit breaker if overload conditions are detected. It is called the twin-bridge power supply after the two resistance bridges which are important to its operation. Although this power supply is somewhat different, no truly new ideas are presented here and the ideas borrowed from others are gratefully acknowledged. There are defects in this design and no attempt has been made to disguise them. Even so, this power supply offers respectable performance.

A Voltage Regulator with Extended Low Range

The heart of the power supply is U3, the common 723 voltage regulator. Specifications for the 723 indicate that it may be expected to regulate to a lower limit of about 2 V. However, by connecting the 7.2-V reference voltage from pin 6 of the 723 to an ingenious resistance bridge, the differential input terminals of the 723, pins 4 and 5, may be kept above 2 V even when the output from the power supply is near zero volts. (See circuit diagram.) This arrangement was described by F. Perugini and E. Gondolfi of Italy in *Electronic Engineering* in 1978.[1] In their design, the base of a 2N3055 NPN power transistor was driven by the 723 from pin 9, such that the 723's internal 6.8-V Zener diode, located between pins 9 and 10, blocked output from the 723 when the potential at pin 10 was less than the Zener voltage. This was necessary because the minimum potential expected at pin 10 is 2 V.

This circuit was viewed with considerable skepticism but was tested anyway. It was disappointing but not surprising to find that although the output would approach zero, the regulation with changing loads suffered at outputs below 2 V. The test results, however, prompted a question. What would happen if a VMOS transistor with a gate threshold voltage greater than 2 V were driven by pin 10 of the 723 using Perugini and Gandolfi's idea? A test was conducted with an IRF511 VMOS transistor driving a 2N3055 power transistor in a Darlington configuration. The regulation improved at all potentials, but particularly below 2 V.

[1]Notes appear on page 8.

The function of the resistance bridge may be understood by examining the circuit diagram. R2 adjusts the output voltage with R1 acting as a fine adjustment control. R3 and R4 act as a voltage divider to produce an output voltage three times the reference voltage set by R2. R3 and R4 in conjunction with R5 form an effective parallel resistance of about 100 kΩ which balances the bridge nicely.

To further enhance regulation, a small resistor, R6, was added to the resistance bridge. It compensates for the voltage drop in the wires leading to the output terminals as the current increases under heavy loads. The magnitude of R6 is a tiny fraction of one ohm and must be determined experimentally. More information about R6 is included in the section on construction. The idea for this compensation resistor was borrowed from an article in *Hands-on Electronics*, by John T. Bailey, in which he describes a very fine power supply of his design.[2]

The output from the 723 at pin 10 must be about 6 V above the output voltage developed at the emitter of Q2. For this reason a separate transformer, T1, is used to power the 723. Because T1 does not experience a heavy load, it is not subject to a significant voltage drop as is T2. It has been found that the current delivered from pin 10 of the 723 is 2.3 mA under maximum load conditions of 2.5 A. This low demand on the 723 contributes to good regulation. It should be noted that there is no bypass capacitor on the gate of Q1. For some unknown reason the circuit works better without it.

Voltage drift caused by temperature change is about +5 mV/°C. The drift after cold startup has been observed to be about 30 mV at the end of one hour. This is quite noticeable at low output voltages when monitored with a digital meter.

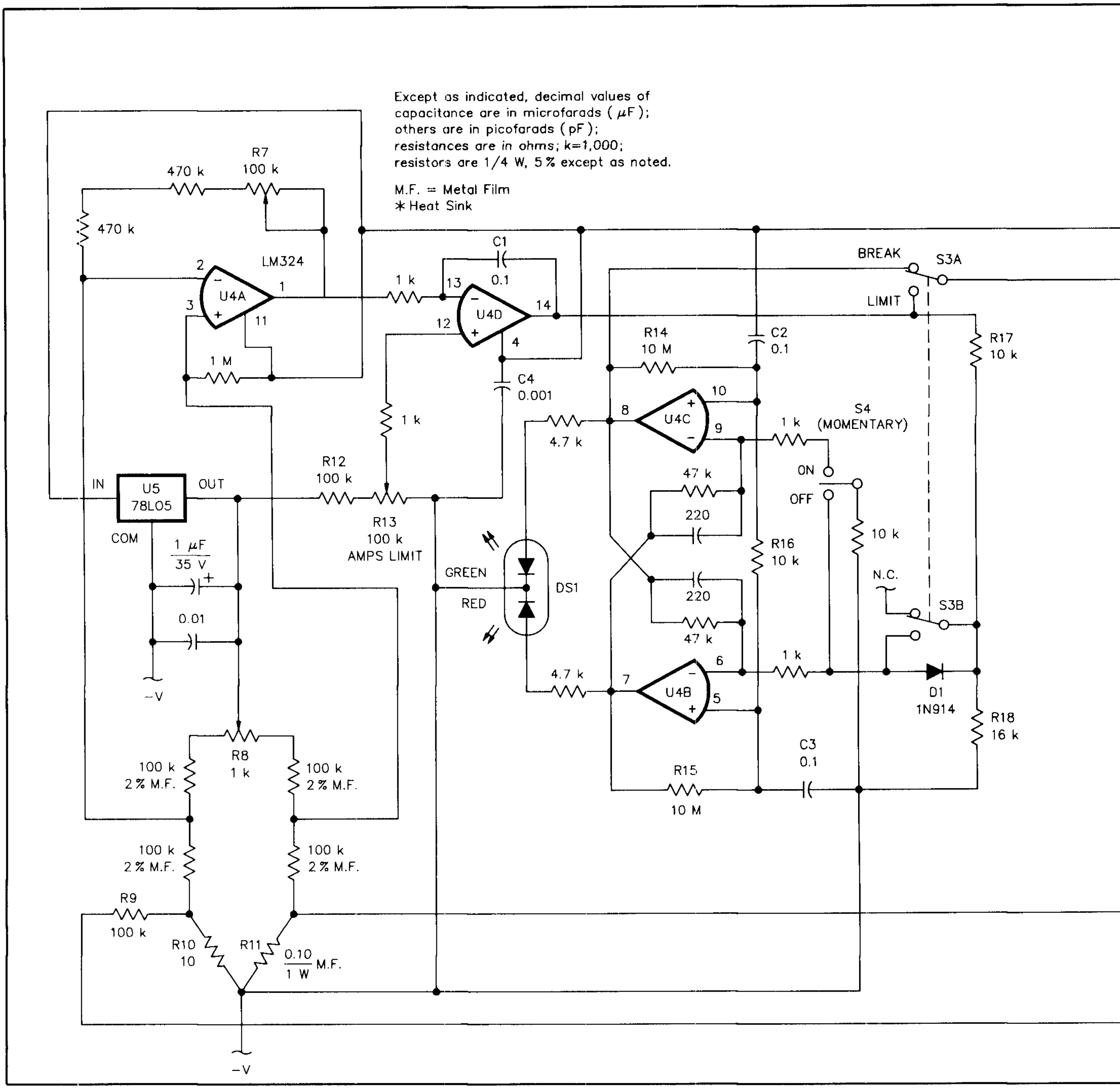
An Adjustable Two-Function Current Limiter

At the center of the current limiter is U4, the common 324 quad operational amplifier. U4A is employed as a bridge amplifier, U4D is used as a voltage comparator, and U4B and U4C are connected together as a set-reset flip-flop. U5 serves as a stable current source for the current bridge resistors and also acts as a voltage reference for the voltage comparator.

When S3 is in the limit position, the current is limited to a preset level determined by R13, an audio taper potentiometer mounted on the front panel. The audio

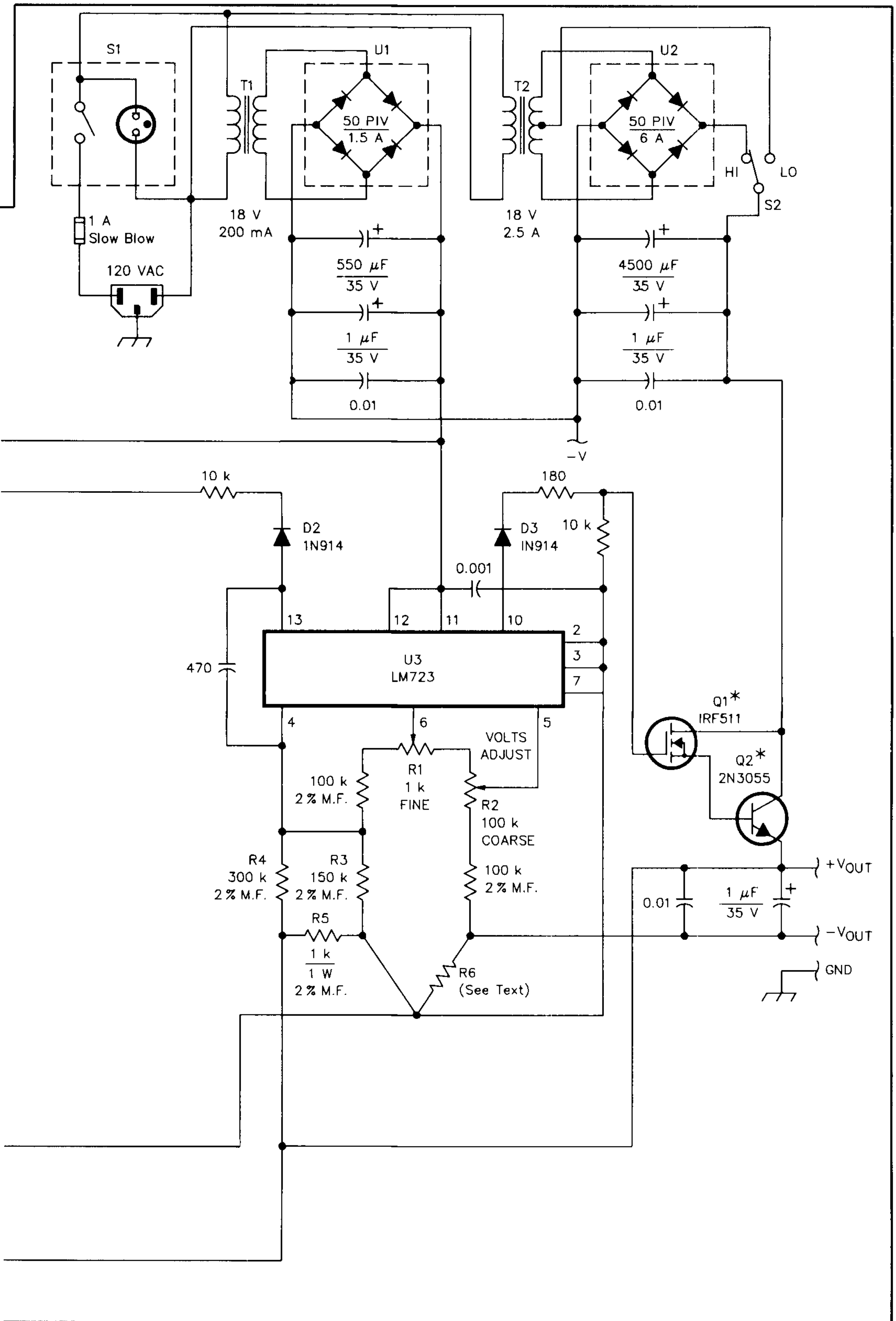
Fig 1—Schematic of the Twin-Bridge power supply.

DS1—Dual red-green LED, RS276-025.
Q1—IRF511 VMOS transistor.
Q2—2N3055 NPN power transistor.
R1—1-kΩ linear-taper, panel-mount potentiometer.
R2—100-kΩ linear-taper, panel-mount potentiometer.
R7—100-kΩ PC-mount potentiometer.
R8—1-kΩ PC-mount linear potentiometer.
R13—100-kΩ audio-taper, panel-mount potentiometer.
U1—50 PIV, 1.5-A bridge rectifier.
U2—50 PIV, 6-A bridge rectifier.
U3—LM723 voltage regulator IC, dual-incline package.
U4—LM324 quad op-amp IC.
U5—78L05 5-V, three-terminal voltage regulator IC.



taper permits more precise adjustment at low currents. In the limit mode the power supply may be used as a constant current source or to charge batteries or form electrolytic capacitors.

If switch S3 is in the break position, the current limiter acts as an electronic circuit breaker which trips at a point set by R13. This mode of operation has been found useful when testing circuits on a solderless experimenter's board where inadvertently crossed wires and other mistakes are a real possibility.



R11 is the current sense resistor for the resistance bridge. R9 and R10 form a voltage divider to compensate for the rise in current through R11 as the voltage increases. This error current is primarily due to R5, which draws 1 mA per volt of output even under no load conditions. R8 adjusts the amplifier offset and R7 adjusts the gain of U4A to 10.0.

U4D receives the amplified output of U4A and compares it to the voltage set by R13. In this power supply, R13 can be set for a maximum reference voltage of 2.5 V which permits a maximum current of 2.5 A, the limit of transformer T2. In the limit position, S3 connects the output of U4D to pin 13 of U3 through a 10-kΩ resistor and D2. When overload occurs, the output of U4D at pin 14 goes low and diverts the output from the operational amplifier of U3 at pin 13. This reduces the drive from U3 pin 10 which also reduces the drive on the Darlington pair, Q1 and Q2. When the output from U4D pin 14 is high, diode D2 blocks the current which would otherwise interfere with the normal operation of U3. The comparator will not behave properly without compensation capacitor C1. This scheme is similar to that used by George Woodward and Mark Wilson in their "R-F Proof 30-Amp Supply" described in *The 1987 ARRL Handbook.*[3]

When S3 is in the break position, the output from the voltage comparator, U4D, is connected through D1 and a 1-kΩ resistor to pin 6 of U4B while pin 13 of U3 is connected through D2 and a 10-kΩ resistor to U4C pin 8. Therefore, when the output of U4D goes low during overload, the flip-flop is triggered. D1 ensures that the flip-flop will remain latched even when overload is no longer sensed. The output of U4C pin 8 remains low, diverting drive from the output transistors, until the flip-flop is reset with S4. A momentary contact SPDT toggle switch with spring return to center was chosen for S4. (Two momentary contact pushbutton switches would work just as well.) When the flip-flop is in the off state, the power supply output voltage is about 12 mV. It is not zero, but it is close enough for practical purposes to provide protection. In the break mode, S4 also serves as a convenient on-off switch for the output of the power supply. S4 is superfluous in the limit mode. DS1 indicates whether the output is on or off or experiencing overload conditions.

The flip-flop is very reliable in the break mode. It seems to be immune to false triggering. However, it is essential that bypass capacitor C4 be present to prevent such difficulties. R14, R15, and R16 are present to give the flip-flop some hysteresis and to promote latching. Increasing R16 will increase the hysteresis. C2 and C3 persuade the flip-flop to assume the off state when the main power switch, S1, is first closed.

Some further comments about the flip-flop are in order. R17 and R18 form a voltage divider which permits the flip-flop to change states as the voltage drops at pin 14 of U4D in the limit mode. It is not perfect. At higher output voltages and currents the flip-flop may not change states to indicate overload. This is because the output of the comparator U4D at pin 14 does not always drop low enough to trigger the flip-flop. Understand that current limiting does occur, but it is not always reflected by DS1 during operation in the limit mode.

The bridge current amplifier shows some of the same thermal drift problems observed in the voltage regulator. Testing indicated that most of the drift was caused by temperature changes in the bridge resistors. The problem was severe enough that the original ¼-watt carbon resistors in the bridge were replaced with ½-watt metal-film resistors. The use of metal-film resistors in both the current and voltage bridges is highly recommended. Experience has shown that metal-film resistors are more temperature stable than carbon resistors. Even with metal-film resistors, the offset drift is about –2 mA/°C which results in a drift of about 10 to 15 mA in one hour.

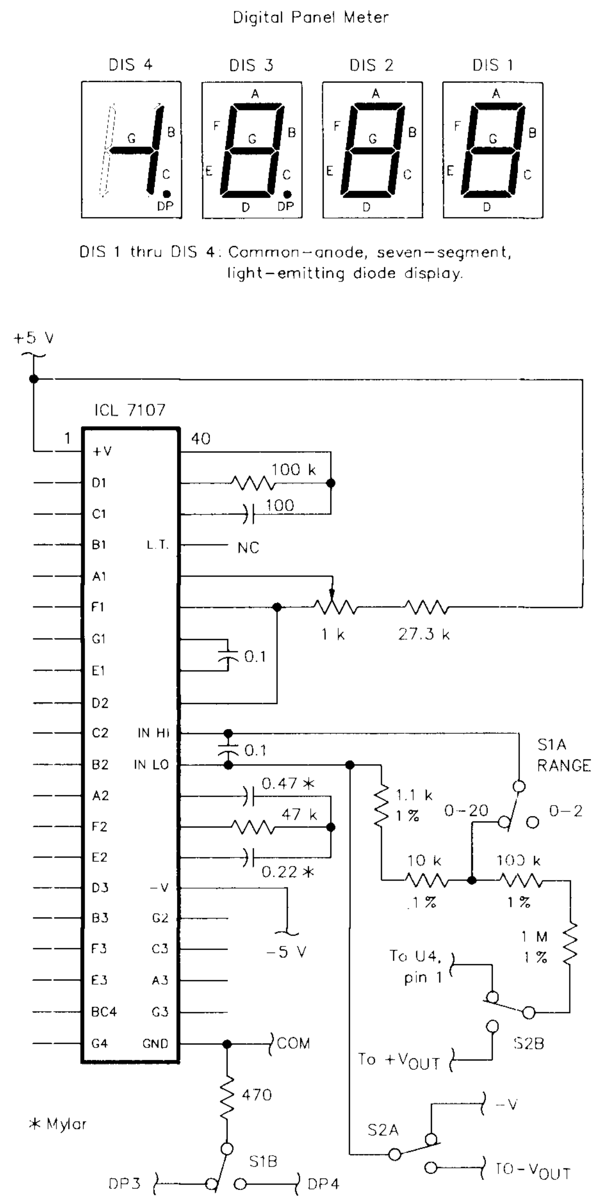


Fig 2—Digital panel meter

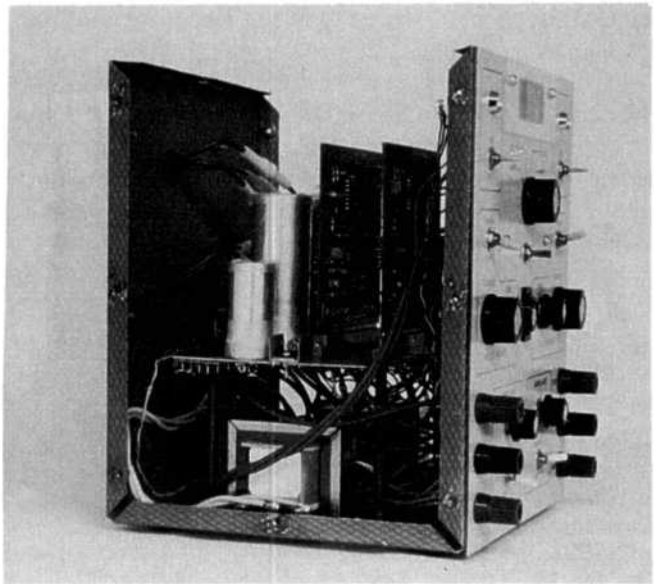
The Digital Panel Meter

Nothing shows the defects of a power supply quite like a digital panel meter. The 3½-digit panel meter used in this project was built around the ICL7107 analog-to-digital converter with outputs for common anode LED displays. The circuit diagram is included in this report but no details are given. The circuit for this panel meter is very similar to that of the digital IC multimeter described by Bob Shriner and George Collins in the Aug 1982 issue of *QST* and *The 1987 ARRL Handbook*.[4] Some builders may prefer to ignore this option or to substitute a D'Arsonval-type meter.

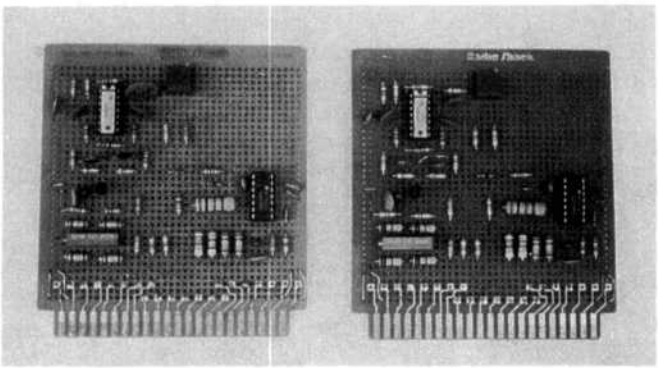
Construction

There are no printed circuit boards available for this power supply. Ordinary perforated board holds the power rectifiers, filter capacitors, R6, R11, Q1, and two sockets for plug-in circuit cards. This project was constructed with two independent, identical power supplies in one enclosure. One plug-in card contains the voltage regulator and current-limiter circuitry for each supply. The card sockets provide convenient tie points for leads to the front panel controls.

There is bound to be a question about R2, the 100-kΩ linear potentiometer used in the voltage bridge. Most potentiometers are rated at plus or minus 20% tolerance although they are usually found to be within about 10%. If R2 is greater than 100 kΩ, so much the better. This ensures that the voltage will be as low as possible with R2 at its minimum because the bridge will be slightly unbalanced. It does not hurt the operation in any way. If



The interior may not be pretty, but it works.



The two plug-in circuit cards.

TEST RESULTS

The Twin-Bridge Power Supply was subjected to some harsh treatment during testing but it performed reliably. It was repeatedly subjected to momentary direct shorts across the output terminals as well as to continuous shorts of as long as five minutes.

In order to test the ability of the power supply to regulate changing loads, a primitive dummy load was made from a length of nickel-chromium resistance wire held between two alligator clips. The magnitude of the load was changed by sliding the clips to different positions along the wire. The results are listed below.

LOAD REGULATION

Volts and amps reported to three decimal places are plus or minus 0.001. Volts and amps reported to two decimal places are plus or minus 0.01.

Load (Amps)	Output (Volts)	Load (Amps)	Output (Volts)	Load (Amps)	Output (Volts)
NONE	0.100	NONE	0.500	NONE	1.000
0.255	0.097	0.130	0.499	0.257	0.999
0.549	0.099	0.418	0.499	0.537	1.000
1.145	0.102	1.288	0.503	0.815	1.001
2.03	0.105	2.11	0.508	1.246	1.002
2.47	0.107	2.49	0.509	2.47	1.008
NONE	2.00	NONE	4.00	NONE	8.00
0.367	2.00	0.400	4.00	0.314	8.00
0.517	2.00	0.800	4.00	0.799	8.00
1.670	2.00	1.514	4.00	1.533	8.00
2.07	2.00	2.12	4.00	1.940	8.00
2.40	2.00	2.47	4.01	2.43	8.00
NONE	12.00	NONE	15.00	NONE	18.00
0.305	12.00	0.379	15.00	0.362	18.00
0.603	12.00	0.753	15.00	0.599	18.00
1.241	12.00	1.448	15.00	1.168	18.00
1.641	12.00	1.798	15.00	1.738	18.00
2.50	12.00	2.46	14.99	1.992	17.98*

*This error in regulation was caused by excessive voltage drop in the power transformer which was in turn caused by the heavy load. Loads of 2 A or greater at 18 V are beyond the capabilities of this power supply.

Test conducted August 1, 1991.

R2 is less than 100 kΩ an adjustment must be made or the minimum output voltage will not be realized. A fixed resistor may be added to the higher voltage end of R2 to raise that branch of the bridge above 100 kΩ, or R2 itself may be modified. The latter approach was used in this project. The resistance may be raised by opening the back of the potentiometer and removing some of the carbon from the resistor by carefully scraping along the edges of the carbon surface. Avoid touching the main surface where the slider makes contact. For this project one of the potentiometers was increased from 90 kΩ to 105 kΩ. The other potentiometer was already 110 kΩ. The load compensation resistor, R6, was made from a short length of 22-gauge wire. Room for R6 was included on the main circuit board and a 14-gauge jumper wire was used temporarily. The value of R6 was estimated by dividing the voltage drop by the current drawn with a known load and further dividing by three. Division by three is necessary because of the voltage divider formed by R4 and R3. In this project R6 was estimated to be about 0.0025 Ω. This required approximately 1.75 inches of 22-gauge wire. A length of wire in excess of 2.50 inches was tried first. The length was progressively reduced until satisfactory correction was achieved. If the length is too long, the output voltage will actually increase under heavy loads.

Unfortunately, adjusting the bridge current amplifier is a tricky operation. It turns out that when the gain is increased with R7, the output measured at U4A pin 1 appears to be reduced. This is because the offset balance is pushed back when the gain is increased. Conversely, the offset balance is moved forward as the gain is reduced. Therefore, R7 and R8 must be coordinated in their adjustments. Also, because of thermal drift considerations, it is important to let the power supply warm up before setting R7 and R8. A power resistor of known value was used as a load to calibrate the current meter. Note that the current as read from U4A pin 1 will not be less than 11 mA. This is a limitation of the 324 when used in a single supply application.

The homemade enclosure fabricated from diamond-embossed sheet aluminum measures 10-inches high × 8-inches wide × 8-inches deep. An external heat sink formed from pieces of sheet aluminum cools both TO-3 power transistors, one from each of the two power supplies. Front-panel labels were applied with pen and ink to a sheet of plastic laminate and protected with polyurethane clear finish.

Conclusion

The need for a power supply with the ability to deliver one-tenth of a volt or less at two and one-half amps is debatable. However, the real advantage of this power supply is the current limiter, and this voltage regulator meshes nicely with it. A three-terminal voltage regulator could probably be adapted to this current regulator, but not without some difficulty.

The thermal drift problems have been mentioned. This may be attributed in part to the decision to package the circuitry in a rather tight enclosure. Also, this power supply has been used in an unheated garage where temperature variations are considerable. Undoubtedly other builders can do better.

Notes

[1] E. Gandolfi and F. Perugini, "Regulation from 0 V with 723 and single supply voltage," Applied Ideas, *Electronic Engineering*, Vol 50, No. 613, Oct 1978, pp 23-25. (This magazine is published by Morgan Grampian Limited in the United Kingdom.)

[2] J. T. Bailey, "Dual Power Supply," *Hands-on Electronics*, Vol 2, No. 1, Summer 1984, pp 21-26 and p 92. (No longer published, this magazine was formerly published by *Radio-Electronics*.)

[3] M. Wilson, ed., *The 1987 ARRL Handbook*, (Newington: ARRL 1986), Chapter 27, Power Supply Projects, "An RF-Proof 30-Amp Supply," pp 27-3 through 27-8.

[4] G. Collins and B. Shriner, "Learning to Work with Integrated Circuits," Beginner's Bench, *QST*, Vol LXVI, No. 8, Aug 1982, pp 29-33. Also M. Wilson, ed., *The 1987 ARRL Handbook*, (Newington: ARRL 1986), Chapter 25, Test Equipment and Measurements, "A Digital IC Multimeter," pp 25-6 through 25-10.

An Audio Peak Filter with Noise Reduction Effect for CW Reception

By Yoshiharu Mita, JHIXEO
4-6-11 Komazawa, Setagaya-Ku
Tokyo 154, Japan

Preface

An audio peak filter (APF) offers a significant improvement in CW reception not only in overall selectivity but also in signal-to-noise (S/N) ratio because of its narrow passband, and the higher its Q, the more improvement of both selectivity and S/N. The higher Q, however, causes a stronger tendency toward "ringing" unless appropriate means to minimize the ringing are used.

The apparatus described here is designed to improve the S/N ratio while minimizing the ringing. It consists of a pair of active APFs and a differential amplifier.

Principal Features

Fig 1 shows a typical differential op amp or common-mode rejection circuit. In this arrangement, when a signal is applied to both of the inverting and noninverting inputs of U, in phase, the output is entirely nulled. Fig 2 shows a differential amplifier like that of Fig 1, but with an APF in each input circuit. If the performance of both APFs is identical, the center frequencies, f_0, being exactly the same, the output is still null because the inputs to U are still common mode. However, if the f_0 of either APF is shifted very slightly (a few hertz), an output appears that favors f_0, discriminating it from the noise. That is to say, the peaked f_0 signal inputs to U are not common mode any more, but a majority of the noise signal remains common mode and is rejected. The greater the difference between f_0 of the two APFs, the more output, but the less the noise reduction and the worse the ringing. In this arrangement the ringing is caused by noise, or the noise is converted to ringing. Therefore, keep the f_0 shift as small as possible to minimize both ringing and noise. The shape of the incoming CW signals (rise and decay times) varies with the Q of the APFs. Higher Q provides better noise reduction but makes the keying delay longer and ringing worse. An acceptable value of Q seems to be about 25, which is based on noise reduction of more than 20 dB and sensible rise and decay times of the keying.

Another approach to reducing the ringing is to incorporate degenerative (negative) feedback into the APF design. In the present experiment, a part of the output from the differential amplifier is returned to one of the APFs, the APF whose f_0 input is 180° out of phase with the feedback signal. The amount of ringing reduction is adjusted by controlling the amount of feedback.

Practical Circuit

The entire circuit is illustrated in Fig 3. A pair of multiple-feedback active BPFs are employed for the APF and arranged so that a perfect balance can be achieved by equalizing input level, f_0 and Q. The input level is adjusted with R2 and R3, f_0 with R5/R7 and R9/R10 and Q with R12 and R14. The filter can be operated at an f_0 of 800 Hz or 500 Hz, selected with switch S1. R4, R6 and R14 are used for fine adjustment and occasional readjustment purposes; R4 for the input level, R6 for f_0 and R14 for Q,

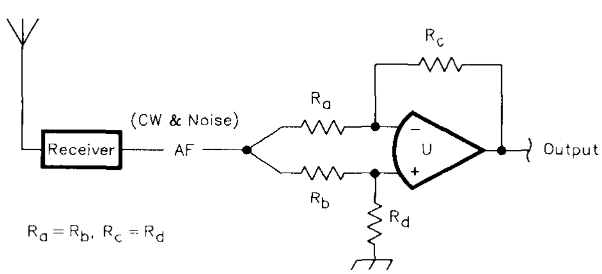


Fig 1

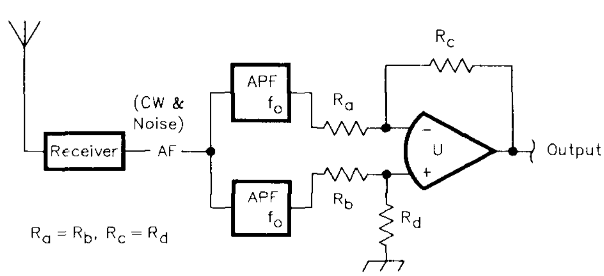


Fig 2

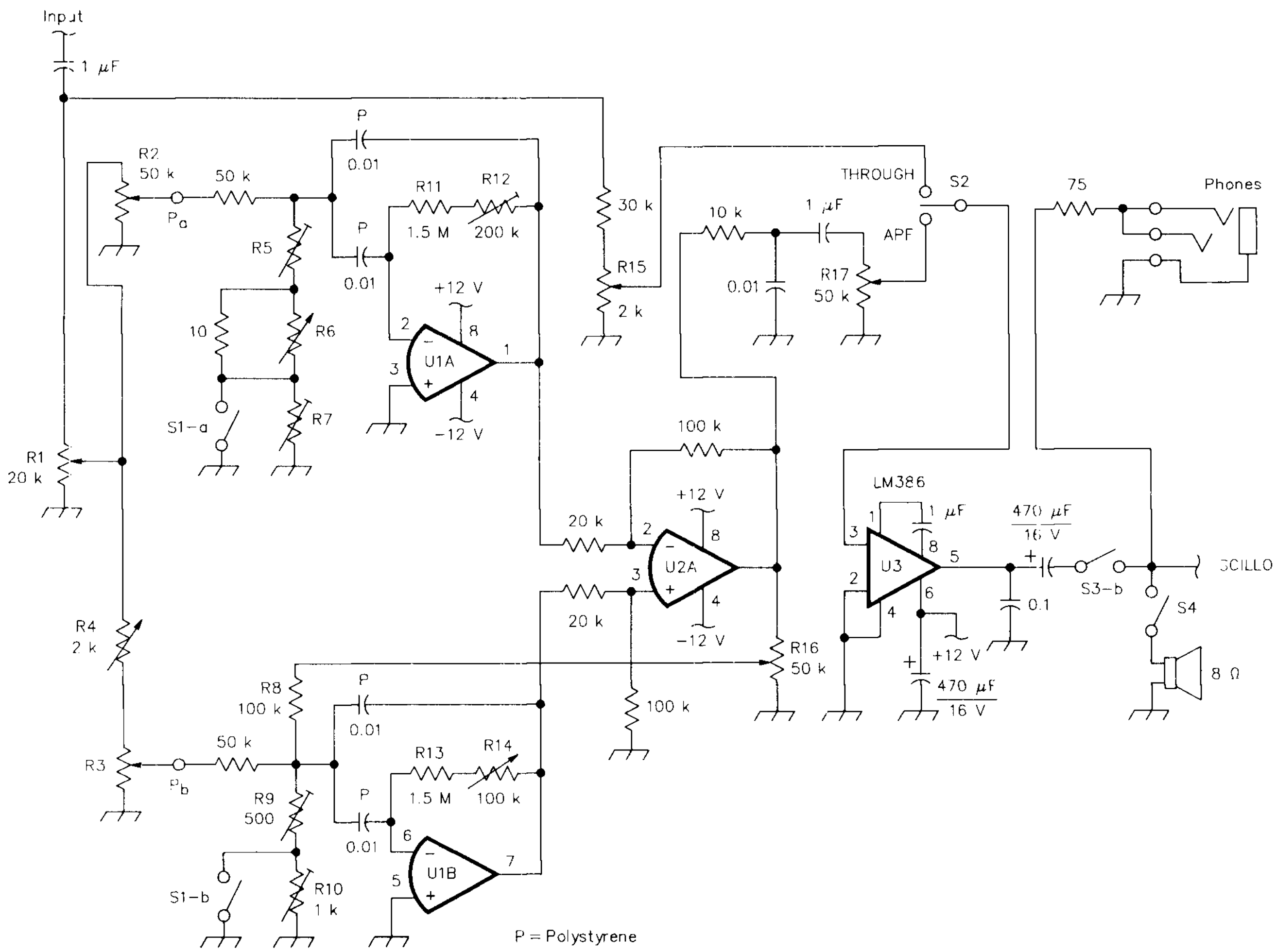


Fig 3—Schematic diagram of the audio peak filter. U1 and U2 are low-noise op amps such as JRC-4562-D.

respectively. S2 selects an operating mode, either APF or THROUGH R1 sets the output level for APF, R15 for THROUGH. The op amps, U1 for the APFs and U2 for the common-mode rejection, should be low-noise devices; I recommend the JRC 4562D or better.[1] R11 and R13 determine the Q of the APFs. A change in the resistance also varies f_0; higher resistance increases Q and lowers f_0.The negative feedback for reducing the ringing is made by inserting a part of the output of U2 into the input of U1-B through R16 and R8. R16 varies the amount of feedback and thus controls ringing.

The value of Q measured through the entire system, consisting of a pair of APFs and a differential amplifier, remains as high as 20-25 even when the ringing is controlled by R16 to be so slight as to be inoffensive.

Alignment

The following paragraphs detail the alignment procedures. An oscilloscope (dual-trace type preferable) and an audio signal generator will be helpful during the work.

Setting of f_0

Set S2 to APF. Adjust R3 for zero input to U1-B. Set R2 and R6 to their midpoints. Connect the audio signal to the circuit input with S1 closed (for an 800-Hz f_0) and apply an 800-Hz signal. Adjust R1 so the input to U1-A is about 10 mV peak at point P_a. Adjust R5 for maximum output from the circuit. Change the frequency of the signal generator to 500 Hz and open S1 (for a 500-Hz f_0). Now adjust R7 for maximum output. Return the frequency to 800 Hz and close S1. Adjust R2 for zero input to U1-A. and set R3 to its midpoint. Check that the input to U1-B is about 10 mV peak at the point P_b. Adjust R9 for maximum output. Change the frequency to 500 Hz and open S1. Adjust R10 for maximum output.

Balancing the APFs

Replace the input source with an audio signal from a receiver. Tune in a noisy CW signal and adjust R1 for

[1]JRC 4562D noise figure: 0.6 μV rms.

an appropriate output level. Adjust R2 or R3, R5 or R9 (for an 800-Hz f_O) and R14 or R12 in concert to achieve a perfect balance to eliminate the noise and ringing completely. Try this alignment also for a 500-Hz f_O using R7 or R10. Use R4, R6 and R14 for fine adjustment.

CW Reception and APF/THROUGH Output Setting

Set S2 to THROUGH. Receive a CW signal and adjust the output level to be the same as the input level using R15. It is preferable to set the output levels of both APF and THROUGH modes equal to the input level. A dual-trace oscilloscope is convenient for such alignments and also for comparing the input and output signals of the APF. Connect output from the terminal OSCILLO to one of the dual-trace inputs and connect the filter input to the other oscilloscope input. Set S2 to APF. Set S1 for an 800-Hz f_O. Set R16 for zero feedback. Destroy the balance of APFs a bit by using R6 to peak the CW signal. This can also be done with R14. Set the APF output level to the same as the input level using R1 and R17. If the APF output is too low, adjust R6 so as to get an acceptable level. When the ringing is offensive, apply negative feedback by adjusting R16. Adjust both R6 and R16 for best reception. As explained earlier, the more shift in either APF f_O, the more output, but the stronger the ringing. Application of negative feedback lowers the output level but reduces the ringing.

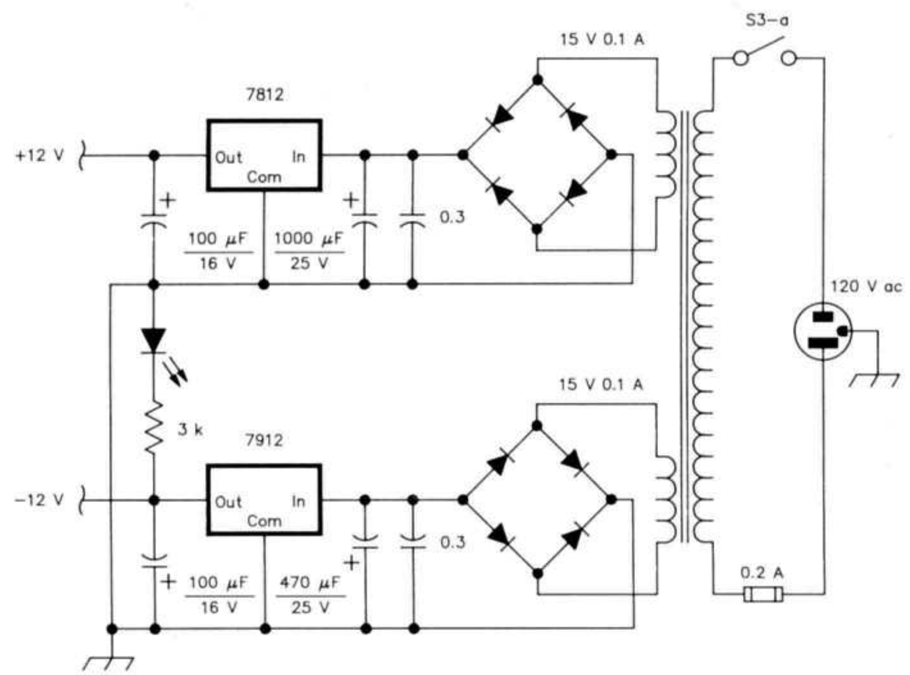


Fig 4—Schematic diagram of the power supply for the filter.

Operation

Performance of the APF system is controlled mainly by the four variable resistors, R4 (input balancing), R6 (f_O balancing), R14 (Q balancing) and R16 (negative feedback control), which are provided on the front panel.

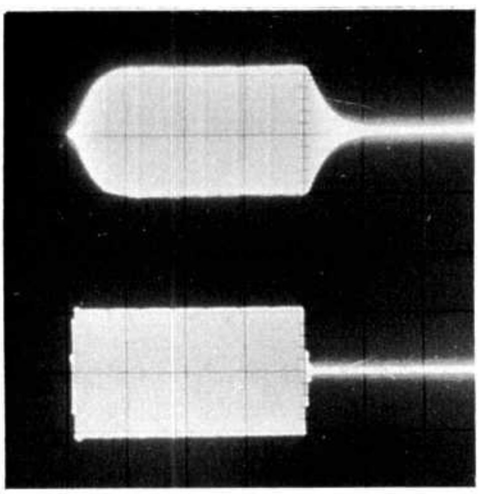


Fig 5—Oscilloscope display of the filter input (bottom trace) and output (top trace) when a keyed CW signal is applied. The risetime of the output signal is about 20 ms.

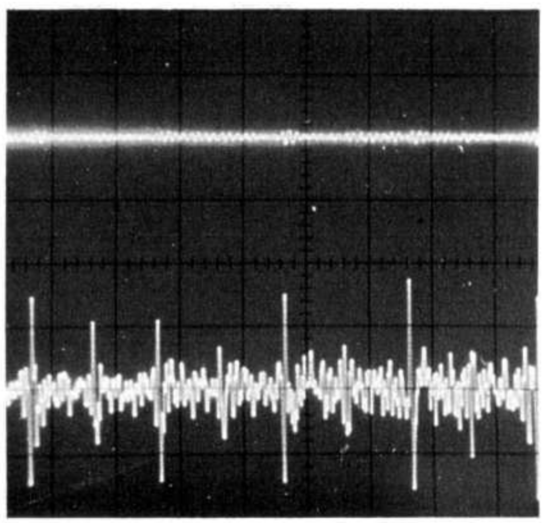


Fig 6—Oscilloscope display of the filter input (bottom trace) and output (top trace) with noise as the input signal. The noise is reduced approximately 20 dB at the output.

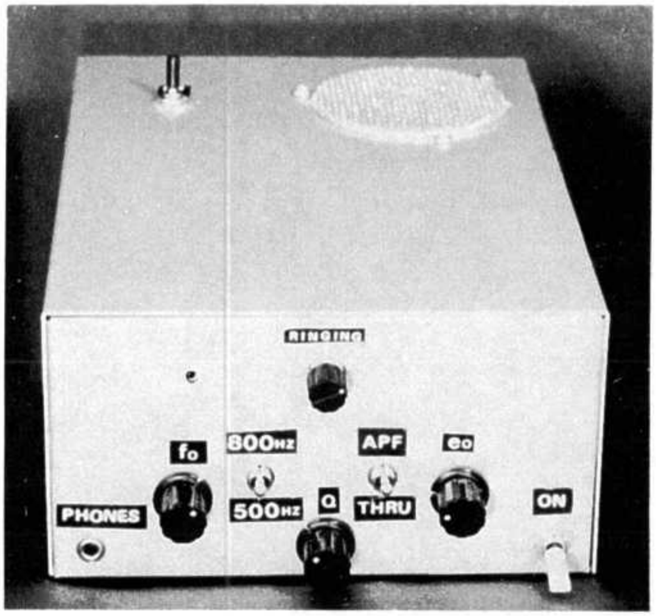


Fig 7

By adjusting these resistors, you can choose the best filter performance for on-air conditions. Reception of a signal at the peaked f_O, with the same level as the input, and noise reduction of about 20 dB, makes comparison of THROUGH and APF modes easy by simply switching S2. For instance, in the case where the noise is not so heavy but the CW signal is weak, a little more destruction of f_O balance using R6 would increase output, although it is accompanied by a slight increase in ringing. On the other hand, unbalancing of the inputs using R4 might effect the same result *without* increase of the ringing but at the sacrifice of selectivity. In cases where the noise is heavy and the CW signal is weak, an increase in the negative feedback by adjustment of R16 and an appropriate destruction of f_O balance with R6 would increase intelligibility of the CW signal even though the output level is lowered.

Input to U1-A and U1-B should be kept below 200 mV peak measured at the points Pa and Pb. Fig 5 shows a display of CW keying shape. The bottom trace is the CW keying input and the upper trace is the filter output. The rise time is about 20 ms and the decay time about 30 ms. Fig 6 shows a display of noise reduction. The bottom trace is the noise input and the upper trace is the APF output. The noise is reduced about 20 dB while the input and output levels of any fo signal are kept the same, as shown in Fig 5.

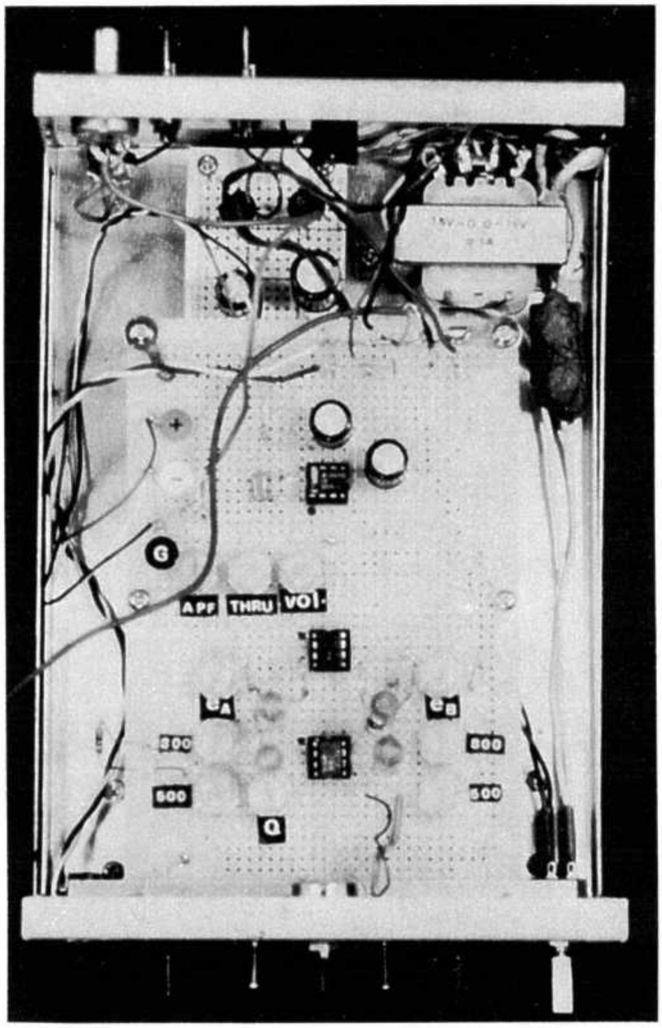


Fig 8

Summary

The conventional noise-blanker erases only impulse noise. The APF system described here reduces a wide variety of noises, but seems to suffer a bit from residual noise in the form of ringing, although it can be reduced about 20 dB in general.

Raising the filter Q as high as 20-25 seems to bring about minor instability in maintaining the balance of the entire circuit. The instablity is mainly due to variation of temperature of the filter capacitors, and the circuit should be en-closed so that they are not exposed to fluctuating temperatures. This is a minor inconvenience in practice, as simple adjustment of R4, R6, R14 and R16 on the front panel compensates for any instability.

The Video Experimenter

By Gregory Glass, N2MOH
71 Park Way
Sea Cliff, NY 11579

Electronics technology is changing. Complex systems rely on the development of sophisticated software to provide functions, so there is less opportunity for the ham to get involved with the development of communication science. Many hams have computer equipment and software development expertise, but the speed required for voice processing puts it out of reach. Yet, if experimentation is limited to still pictures, the facilities of the average ham/computer hacker can produce results comparable to that of the best video research labs in the world.

DSP (Digital Signal Processing) technology is becoming more popular in voice and CW communication systems. There are a number of advantages to digital techniques:

- Cost—DSP can provide functions that are difficult or expensive in hardware,
- Flexibility—performance of a system can be changed by loading new software,
- Speed—new ideas can be tested by writing new software,
- Maintainability—after the initial investment the system can be upgraded with little or no cost,
- Accessibility—writing new software can be done by people who are not hardware "hackers,"
- Low Noise—quantizing in the system makes it resistant to system noise.

These advantages of digital processing in audio systems apply to video systems as well. Moreover, because of the raster organization of video pictures, digital processing is the only practical way to perform some functions like line-to-line low-pass filtering. The problem is, the greater bandwidth of video requires faster processing than audio or CW processing. For example, a simple 2D low-pass filter requires 9 multiplies and 9 adds per sample, plus some housekeeping operations. A picture at video rates (approx 6-MHz bandwidth) requires more than 10 million operations per second. This stresses the performance of the fastest available processors and places great stress on the budget of the amateur experimenter. And more useful work requires even more operations per second.

Fortunately, there is a solution. Most experimentation in the transmission of pictures can be done in non-real-time by slowing down the rate that information is being transmitted. For single pictures, this does not require more hardware than an amateur might already have.

Fig 1 shows the components of an experimental ATV setup. The original signal comes from a frame buffer that is connected to a computer. There are a number of available buffers with an on-board NTSC encoder, or a low-cost external encoder may be used. None of these low-cost systems are broadcast quality, but acceptable results can be obtained by careful selection of frame buffer colors and contrast. The video output of the buffer is fed to the ATV transmitter. If experimentation is limited to the video signal, any ATV transmitter will do.

On the receive side, the NTSC output of the receiver is fed to the frame grabber. The frame grabber takes the analog video, digitizes it, and stores it in a form that a computer can access and process. There are a number of low-cost frame grabbers now available. The scope of experiments will be defined, to some extent, by the capabilities of the frame grabber. Most low-cost grabbers use 6-bit ADCs, many digitize luminances only, and some can't take a whole frame at one time (video digitizers versus frame grabbers). But much work can be done with the simplest hardware if the limitations of the equipment are taken into consideration.

A low-cost system may be constructed using an entry-level PC, a VGA card with an external NTSC

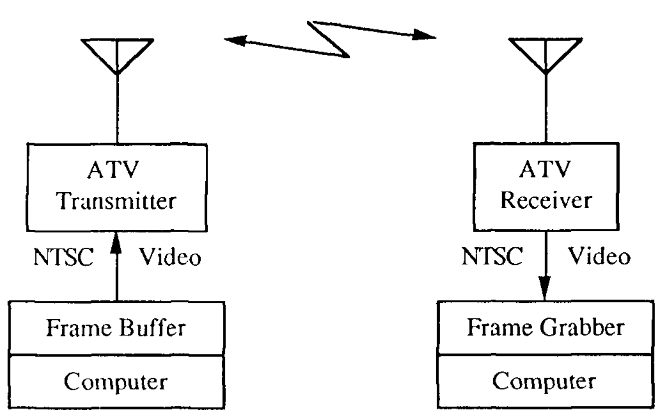


Fig 1

encoder, and a 256 × 256 × 6-bit video field digitizer. The same types of experiments may be carried out with this equipment as at the best video labs in the country. They just must take place at lower resolution, and processing will take longer. Compressing a 256 × 256 pixel image to 128 × 128 pixels then expanding it back to 256 × 256 resolution demonstrates the same compression ratio as taking a 1K × 1K image and reducing it to NTSC video.

A good, but more costly, amateur experimenter's system would be based on a faster computer, such as a 486 PC (the floating point processor will make software development easier), with an integrated display card/frame grabber and NTSC encoder/decoder. The *Truevision Targa+* card will work well. (The *ATVista* card permits reprogramming of the video parameters that may lead to other areas of experimentation, but off-the-shelf ATV equipment expects an NTSC signal.) Note that the frame grabber or digitizer requires a baseband NTSC video signal, while the output of most ATV downconverters is RF meant to feed the receiver in a TV set. A TV or VCR that has NTSC (composite video) output may be used in this case.

Low-power ATV equipment is perfect for across-the-basement testing. Because frame buffers like the *Targa+* cannot be a video source at the same time they are digitizing pictures, you need two computers and cards for "closed" experiments. A VCR can be used as a time multiplexer for sending and receiving tests with one computer. First the output signal is recorded on the VCR, then that signal is transmitted into the digitizer. Just remember that the resolution of a VCR is not very high. This is not a major limitation; your encoder is probably limited to less than this.

Any spatial transformation of a picture can be carried out and tested with this setup, although temporal processing cannot be tested without single-frame recording equipment. While such equipment is expensive now, low-cost versions will soon be available. Software-only solutions, like Apple's *QuickTime*, are sometimes adequate.

Now that an experimental framework is outlined, some tests can be defined. These may cover:

- noise reduction
- ghost elimination
- alternative color processing
- digital video transmission
- resolution enhancement (HDTV)

These are just a few experimental areas that can be supported by this environment.

Some Experiments

For now, I will outline a few experiments that can take place using the video lab that I have outlined. To carry out these tests requires some understanding of software development and the ability to read and write data to the frame buffer, but that's all.

Digital Video

Digital transmission of information is often done using quadrature modulation. This is just a fancy term for limiting the received signal to 16 levels. I think that 16 levels are used because they can be represented using 4 bits, and that is half of 8 bits. To transmit a picture using quadrature modulation, first take half of the picture, split each 8-bit pixel into two 4-bit samples. Then send these two pixels. In the receiver, digitize the frame, quantize the pixels to 4 bits and combine to get the original 8 bits.

This digital encoding will eliminate noise in the low bits. The number of levels can be reduced to remove more noise in the picture with a further reduction of the spatial resolution. Digital compression of the picture can recover some of the area that is lost.

Digital Video with Compression

A major limitation of the above video encoding system is that it takes at least twice as much bandwidth as the original analog video. This requirement can be reduced by compressing the picture before it is transmitted. If the picture is reduced to requiring only 4 bits per pixel, there is no spatial reduction. A simple way to convert the picture to 4 bits is to use delta encoding. Each pixel represents the change in value from the previous pixel. Because the change from one pixel to the next is limited to –7..+8 this has the effect of a horizontal low-pass filter. If the asymmetry of –7 to +8 bothers you, just eliminate the zero change and go from –8 to +8. Pictures look better with a little low-order noise in flat areas anyway. Delta encoding can be expressed as:

$$P_{new[i,j]} = \text{Quantize\_to\_4\_Bits}(P_{old[i,j]} - P_{old[i-1,j]})$$

To keep the ±8 level changes from flattening out the picture, an error term may be calculated and added to the next pixel. This *dither* method more quickly brings values to what they should be at the expense of having to store a few more values.

$$\text{Error} = P_{new[i,j]} - P_{old[i,j]}$$

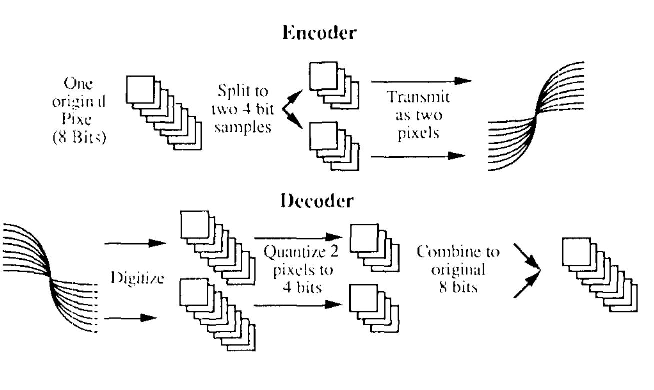


Fig 2

$$P_{old[i+1,j]}=P_{old[i+1,j]}-Error$$

Using delta encoding with error propagation can make very realistic pictures.

Ghost Reduction

This is a fun one. Because ghosts in video pictures are correlated with the picture content, they are very apparent. If the picture elements in the image are randomly shuffled before transmission and then unshuffled in the receiver, the ghosts will become uncorrelated and turn into noise that does not degrade the picture as much. For motion video, it is best if a few shuffling tables are used and randomly selected during transmission because temporal correlation of noise is detectable by the eye.

Component Video

NTSC encoding has less resolution for color information in pictures than intensity information. There are also interactions between the chromaticity and luminance that limit resolution. The super VHS and Hi8 recording standards record chroma and luminance separately to increase resolution. You can do the same with transmission. The first thing to try is sending just the red, then the green, then the blue.

The MAC system that is sometimes used in Europe places the luminance (Y) at the beginning of the scan line, then follows with red minus luminance (R-Y) and blue minus luminance (B-Y) at lower resolution. Luminance of a picture may be calculated as:

$$Y = Red \times 30\% + Green \times 59\% + Blue \times 11\%$$

These are just a few of the experiments that the amateur can perform. Experiments can also take place with high-resolution (HDTV) transmission. With the addition of a *Sound Blaster* audio card, the video frame buffer can serve as the basis of experiments with digital sound transmission over a video link. Think of the frame store as being a buffer that can hold any type of data and then spit it out at a high rate. The possibilities are endless.

Feedback

In September's "Synchronous Detection of AM Signals," we indicated that Sony has information available about their CX-857 integrated circuit. It turns out that the information is not provided in the US by Sony, nor can they provide technical assistance. But they have agreed to allow us to provide the data sheets by copying them for our readers. If you want a copy of the data, write to: Technical Depart-ment, ARRL, 225 Main St, Newington, CT 06111. Ask for the CX-857 data and include a self-addressed, stamped envelope.

RF

By: Zack Lau, KH6CP
225 Main St.
Newington, CT 06111
email: zlau@arrl.org (Internet)

The Quest for 1 dB NF on 10 GHz

Like everyone else, I find it quite satisfying to know that my receiver is as hot as possible, although it's sometimes debatable whether it's really necessary to go through all the effort of getting it there for terrestrial use. For instance, I doubt whether I could have made any more contacts last year in the 10-GHz contest, despite having a noise figure of approximately 5 dB! From what I could tell, I heard signals as well as anyone else. This might be considered surprising, since many people claim noise figures of 2 dB or less.

Perhaps that 5 dB figure is a little misleading—it refers to the *total* estimated system noise figure. This is usually a bit higher than the noise figure of just the preamp, and it can be much higher. For instance, the typical SMA relay adds about 0.5 dB of loss, which translates directly into increased NF, though I've seen relays with anywhere between 0.3 and 1 dB of loss. This of course is a fuzzy number, since there is also the effect of mismatch tossed in there. And it's usually not even the mismatch loss that is the big problem, but the mismatching of the preamp itself. In other words, you may have only 0.1 dB of mismatch loss, but the different impedance seen by the preamp may cause your 1-dB NF preamp to have a 1.5-dB noise figure. And if you aren't careful, you might have an oscillating preamp, in which case you probably don't want to know what the noise figure is! Finally, the gain of your preamp may not be enough to override the noise of the following stages. While a well-designed image-reject filter, mixer, and post-amplifier may have a noise figure as low as 10 dB, something just thrown together could have a noise figure as high as 20, or even 30 dB.

As you might guess, measuring noise figure isn't easy at 10 GHz. You probably will want to use a coaxial adapter for routine measurements, which adds to the system noise figure. Ideally, you would use precision 3.5-mm connectors on your noise figure source and preamp, but there is usually an affordability problem with these. Even if you do have a 3.5-mm connector on your noise figure source, there is a wear problem to deal with in mating them with SMA connectors, at least according to the Hewlett Packard literature I've read. And I'd have to agree, having inspected well-used connectors and seen the amount of wear on them. The ARRL Lab currently uses an HP 346A noise source with the usual type-N male connector. Using a low ENR head (approximately 5 dB) seems to yield good results, but a high ENR head, such as an HP346B, and a "calibrated" attenuator, often seems to yield results that are several tenths high or low.

But the noise source is only part of the test setup. Just as important is the equipment following the preamp under test. At lower frequencies, the preamp output is typically connected directly to the noise-figure meter, providing highly accurate results. But few NF meters receive 10-GHz signals. Rather, you must mix the preamp output down to a frequency within the range of the NF meter. For example, the HP8970A NF meter we use here in the ARRL Lab can receive only up to 1.6 GHz. (The 8970 is a fancy receiver with internal number crunching.) In this arrangement (see Fig 1), the characteristics of the system you use to mix the preamp output into the range of the NF meter can have a profound effect on the NF measurement. Despite its high price tag, the 8970 does not like spurious signals. The IF should be well filtered, even if the local oscillator is in the microwave range, to keep unwanted mixing products from reaching the NF meter. You also have to remember that the NF meter merely detects noise pulses; it has no idea whether the noise pulses are images or the real thing. Thus it is easily fooled by poorly designed equipment. One way to test whether this is happening is to insert a low-loss, narrow bandpass filter between the noise source and the converter. If the noise figure goes up much more than the filter loss, your measurement system needs work, unless the filter is introducing some sort of mismatch. To measure a 1-dB NF preamp, I think the image rejection of the test setup should be at least 17 dB. It can be shown that image rejection is less important as sensitivity gets worse.

For a good sanity check on your measurement setup, you might consider listening to the IF output on an SSB/CW receiver, although an AM receiver does a better job. Induced instabilities or performance oddities can often show up in the received noise pulses.

Once we know we can measure the NF, the challenge is to find the appropriate low-noise device to use, which seems to be a PHEMT. Engineering samples are often the hottest devices available, so I got a pair of NEC 32684As from the local NEC rep. Using Touchstone (an RF design program by eesof), I came up with a "hot design"—a 0.52-dB noise figure and 13 db of gain. Unfortunately, it's not unconditionally stable in the passband, meaning that it needs to see 50-ohm source and load impedances to ensure stability. I'm not too worried about the source impedance since dish feeds are often pretty wideband. But I did make sure that it doesn't oscillate and is unconditionally stable at much higher and lower frequencies, where the antenna impedance may depart from 50 ohms. It just misses at 6.5 GHz (K=0.99), but I haven't found that to be a problem.

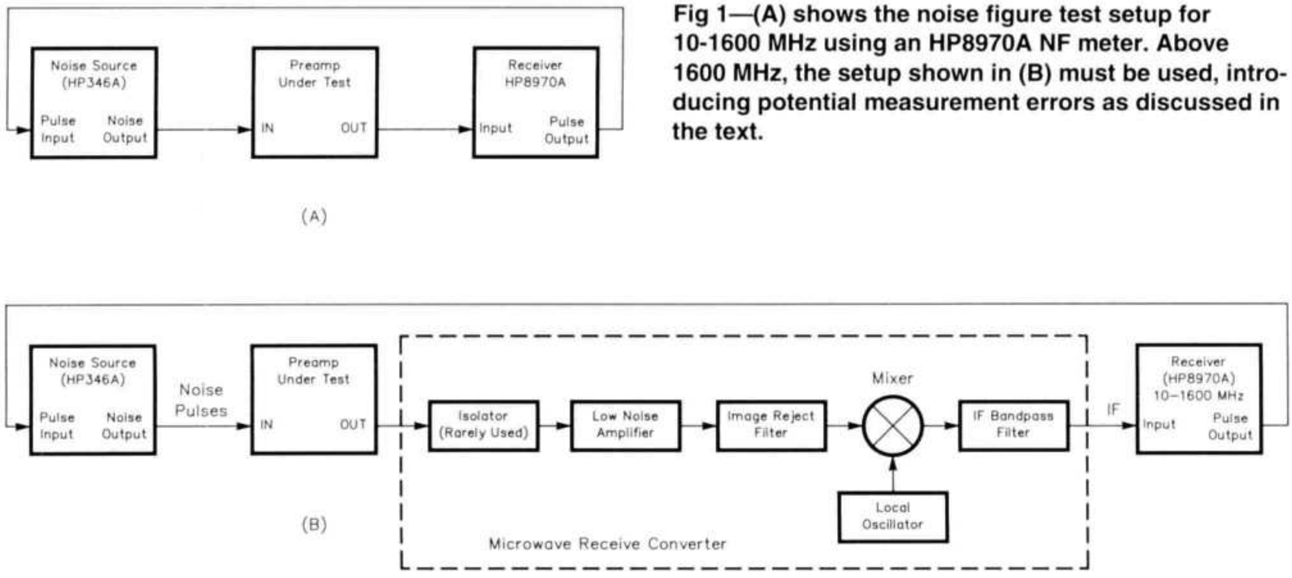


Fig 1—(A) shows the noise figure test setup for 10-1600 MHz using an HP8970A NF meter. Above 1600 MHz, the setup shown in (B) must be used, introducing potential measurement errors as discussed in the text.

I chose to do the board on 15-mil 5880 RT/Duroid™, which is available from Microwave Components of Michigan.[1] I believe Norm stocked it for people building the TNT—a 10-GHz transverter design that appeared in the 1988 Microwave Update (sorry, no LO, and definitely not no-tune). The thin board avoids the radiation-loss problem, which seems to be considerable with 30-mil board at 10 GHz. It's tough to build a low-noise preamp if much of your signal is radiated before it gets to the FET! For example, my guess is that the Al Ward ATF-13135 design described in recent editions of *The ARRL Handbook* loses about half a dB in NF due to radiation loss.

My design makes use of source inductance to make construction easier. 15-mil board is just thin enough to allow through-board source grounding with Avantek -036 case devices. Unfortunately, I initially laid out the board wrong, and I got an unacceptably high noise figure. Fortunately, the device survives the unsoldering test! Putting the preamp in front of a 1.3-dB NF transverter with lots of gain, I measured a 1.1-dB noise figure. But I also noticed that holding the preamp just right reduced the noise figure another tenth of a dB! This problem resulted from the way I mounted the SMA connectors. The ground plane connection had a gap due to the connector dielectric. While it was a pretty small gap, it apparently caused a tenth of a dB of loss at 10 GHz. Covering the gap with solder cured the problem.

Having five GaAs FETs cascaded, which was the case in my system with the preamp connected to the transverter, really isn't stable, even if you have three separate boxes, all shielded. So I decided to build yet another 10-GHz transverter to mate the preamp with. The second stage is a pair of ATF-13036s on 15-mil board (1.2 dB NF), and the mixer is a carefully tuned rat-race mixer fed with LO from a splitter that also feeds the transmit mixer. The image-reject filter between the mixer and

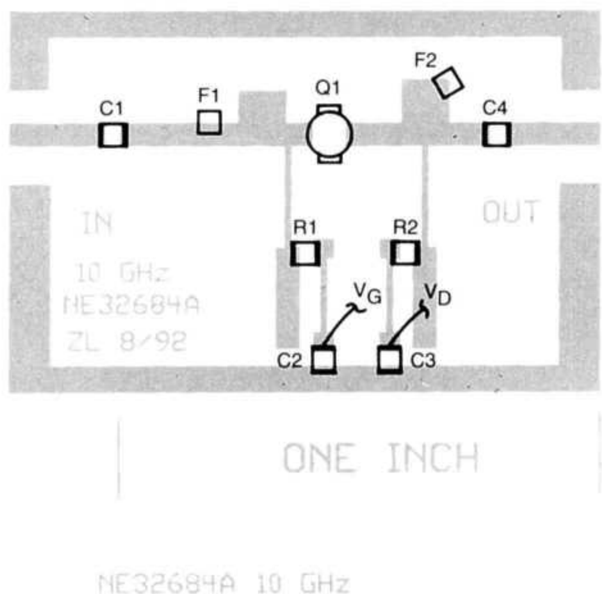


Fig 2—Layout of the NEC32684A preamp board. Use 15-mil 5880 Duroid. The board I used had a dielectric constant of 2.2 and a dissipation factor of 0.0011. A negative of the board layout is available from ARRL. Send an SASE to: Technical Department, ARRL, 225 Main St, Newington, CT 06111. Ask for the template from the December 1992 *QEX*, "RF" column.

[1]Microwave Components of Michigan, PO Box 1697, Taylor, MI 48180, tel: 313 753-4581 evenings.

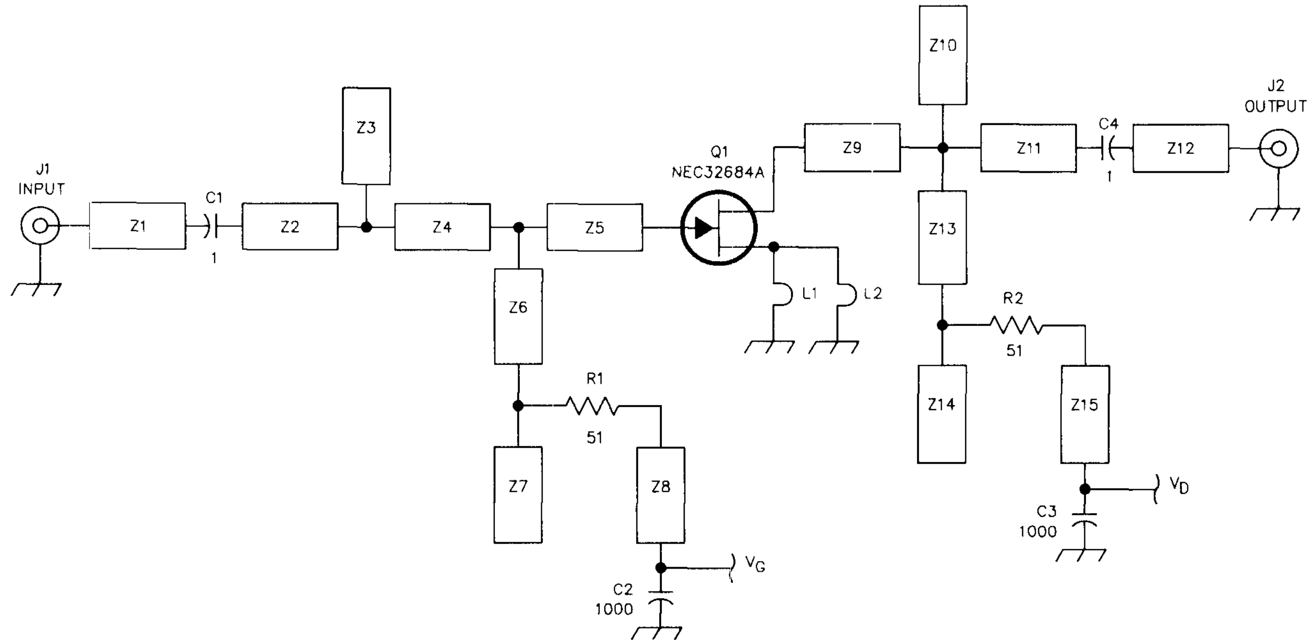
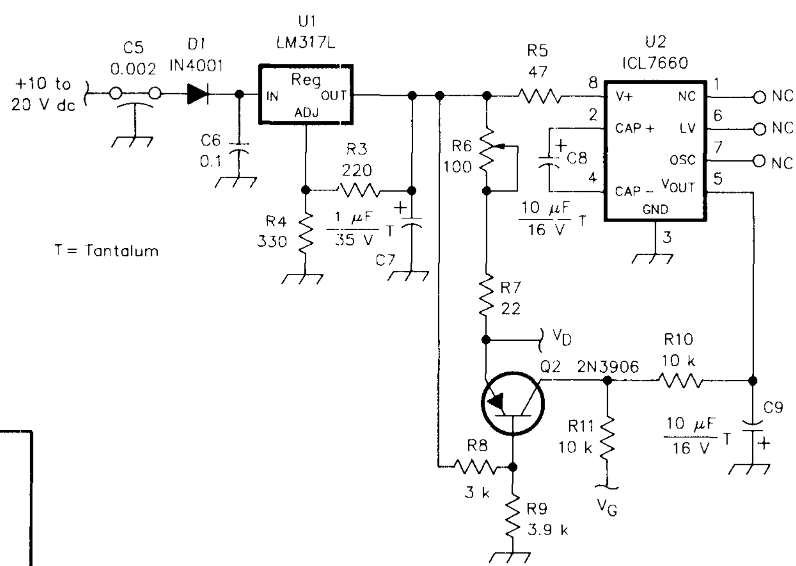


Fig 3—RF schematic of the 10-GHz preamp.

C1, C4—1-pF ATC 100A chip capacitors. C1 must be very low loss.
C2, C3—1000-pF chip capacitors. *(Not critical.) The ones from Mini-Circuits work fine.*
F1, F2—Pieces of copper foil used to tune the preamp.
J1, J2—SMA jacks. *Ideally these should be microstrip launchers. The pin should be flush against the board.*
L1, L2—The 15-mil lead length going through the board to the ground plane.
R1, R2—51-Ω chip resistors.
Z1-Z15—Microstriplines etched on the printed circuit board.

Fig 4—Schematic of a dc bias supply for the PHEMT.
C5—0.002 µF. (Not critical.) I bought a bag of these cheaply, which explains why it isn't 0.1 µF or 1000 pF.
Resistors—Except for the miniature trimmer, R6, all are ⅛-watt resistors to allow building the entire bias circuit on the sheet of copper used for the RF ground plane. The entire preamp, including connectors, has a volume under 1 cubic inch. Tantalum capacitors are similarly used for their small size. Electrolytic capacitors should probably work just fine.



the preamps is a half-inch plumbing-cap filter (Kent Britain's design on page 162 of the *Proceedings of Microwave Update 1988).* With 0.1-inch probes made from 0.085-inch semi-rigid coax center conductor, I usually get 17 or 18 dB of image rejection at 2 meters.

After hooking everything up, including the T/R circuitry on the IF, I measured the receiver and got a 2-dB NF and 19 dB of gain. Something isn't right—should be about half that noise figure. Turns out the T/R switch was generating noise at 2 meters and injecting it into the transmit mixer. Since there was only about 4 dB of isolation between the IF ports at 2 meters, noise on the transmit port was a big problem. The solution, once I figured out what was going on, was pretty simple. I merely added quarter wavelength (at 10 GHz) pieces of wire across the RF and LO ports of the mixer. This kept the 2-meter signals where they belonged while not affecting 10-GHz operation. Actually, they may have improved 10-GHz operation. I have noticed unusual variations in conversion loss when diode pairs are floating with respect to ground, but these variations seem to have disappeared after adding these DC shunts/RF chokes.

So, how well did I do? Well, without the RF relay and connecting cables, my transverter has a 1.0-dB NF and 22 dB of gain, according to our NF measurement setup. During the 10-GHz contest this year, I noticed I could tell the difference between sky and ground noise with my 2-foot dish. At the 1992 Microwave Update, my transverter measured 0.9-dB NF. So, all I have to do is find that 0.1-dB-loss SMA SPDT relay with cables and I'll be all set! A second 32684A preamp easily beat another PHEMT preamp in terms of NF, though you will have to check the upcoming North Texas Microwave Society *Feedpoint* for the exact numbers.

Feedpoint, by the way, is well worth reading if your interests include microwave experimentation. This newsletter is published every two months. Dues are $12 per year. *Feedpoint* is available from:

North Texas Microwave Society Feedpoint
Membership—Wes Atchinson WA5TKU
Rt 4 Box 565
Sanger TX 76266

So you want to build one of these preamps? Here are the circuit details. Keep in mind that single-stage preamps at 10 GHz are very sensitive to variations in load impedance. Ideally, you should terminate the preamp with an isolator. Alternately, you can tune up the preamp into your second stage with small pieces of foil on or near the etched transmission lines, for best performance. Adjust R6 for best noise figure. According to the transistor data sheet, a drain bias current of 10 mA is a good starting point.

Note
Duroid is a trademark of Rogers Corp.

Digital Communications

By Harold Price, NK6K
5949 Pudding Stone Lane
Bethel Park, PA 15102
email: nk6k@amsat.org (Internet)
or 71635,1174 (CompuServe)

New Column—Old Columnist

Well, not all that old, really. Not old enough to avoid getting caught in one of the most ancient of traps. You know, the one where someone asks your advice on how something should be done, and you end up doing it? That pretty much explains my current presence on these pages. Jon Bloom, the editor of *QEX*, wants to have a recurring technical digital column that will help push the current state of amateur digital technology. "It can't be done," said I. "No one person will be able to provide true leading edge information and the range of subjects you want. The people who are doing these things seldom write them up. The best you could do would be to get someone who knows the people doing the good work. Such a person could force, cajole, or trick the right people into writing up what they are doing and how they are doing it. Said person could also provide background, and continuity."

In a way that is not entirely clear to me now, I got volunteered to be that person. I've been given the charter to get the interesting technical ideas and algorithms into print. A column in *QEX* is an excellent forum. The column format is less formal, and *QEX* readers are expected to already have a basic understanding of terms and methods. Those of you who are doing interesting work, and who don't take the time to write it up for others to see what you're doing, and you know who you are, can expect to hear from me, some place, some time when you least expect it.

To set a good example, I've provided the subject material for this month's column. This month's topic: "Applications of Low Earth Orbit Store-and-Forward Protocols in the Terrestrial Environment," or "Geeze, Are We Still doing *that* on HF?" [Disclaimer: This column is not meant to cast aspersions on the current HF network, which grew from objectives that were valid when it was first designed. My desire here is to suggest alternatives and start a dialogue. You know, columnist stuff.]

BBS Forwarding on HF

In a series of meetings, letters, conference calls, protests, and follow-up meetings, some of which are best described as "intense,"[1] the issue of BBS forwarding on HF has recently been reviewed. Although the review was mainly from a political/administrative, regulatory/band-planning point of view, it got me to thinking about the pure technical merits of what we're doing on HF. Since I haven't been directly involved in terrestrial BBS forwarding since I ran the RLI-based 145.36- to 145.01-MHz forwarding bridge in Los Angles in the mid to late 1980s, I checked with Dave Toth, VE3GYQ, who is very active in the HF packet game. He assures me that the basic technology remains the same.

First, some history. Packet BBS systems saw their first widespread use in late 1983 and early 1984 as TNCs became readily available. The BBSs were geared toward the normal operating mode of the TNC. Although it seems silly now, the first TNCs were not designed for computer-to-computer communication. The interface was meant for use by a human typing at the keyboard of a "dumb terminal." Most hams did not have access to sophisticated computers at home, or compilers, or real operating systems. Imagine the KA9Q NET or NOS software running under CP/M on a 2-MHz Z80 with 64 k of RAM and two 180-kbyte floppy drives. TNCs were designed to run with a swap-meet ASR-33 or glass TTY, and the BBS systems used what was available.

TNC-based BBS systems were very popular. Since the TNC allowed eight-bit codes rather than the five-bit Baudot code used in RTTY, the look and feel of a packet BBS was closer to the more familiar telephone BBS systems than were the RTTY MBOs. When IBM PC clones brought the price of better computers in range of the average ham, BBSs spread quickly. Even though more computing power was available, the interfaces remained the same, that is, keyboard-to-BBS. For a time, BBS systems were isolated, users left messages on the BBS, where they remained. Hank Oredson, WØRLI was one of the first hams who addressed the problem of message forwarding, a way to get messages from one packet BBS to another. Since the virtual keyboard was the most common and best documented access method, Hank used that model to build his BBS-to-BBS forwarding scheme. One BBS would pretend to be a keyboard user and connect to another BBS much as a live user would. The idea was fairly easy to implement, and very easy for nonnetwork gurus to understand. The idea spread, became popular, standard, ubiquitous, and finally cast in stone.

Packet BBS systems were first used on VHF and UHF frequencies. These are characterized by a fairly low bit-error rate (BER), a high bit-transfer rate, and long access times. The systems were next used, with little or

[1]Notes appear on page 22.

no modification, on HF. HF frequencies are characterized by a higher error rate, a low bit-transfer rate, and access times limited by propagation. Almost by design, therefore, these systems perform poorly, from a theoretical point of view. From a practical point of view, HF packet BBS systems far outperformed existing HF RTTY and AMTOR systems in areas that packet users deemed important. Packet systems:

1. Allowed use of (new) standard forwarding software.
2. Allowed use of readily available, low-cost, standard hardware (several years ago, before I stopped tracking manufacturers, more than 100,000 TNCs had been sold).
3. Allowed use of an extended character set (lower case, special characters).
4. Allowed high transfer rates when the signal path was good, allowing longer messages to be transferred in a shorter time.
5. Had a better error detection system, resulting in less data errors.

These advantages encouraged many people to use HF packet and to begin forwarding messages through various ad hoc networks that became more formalized as time passed. Thousands of messages are now passed world-wide each month. The system has worked well enough that it has unfortunately permitted us to overlook the following problems with HF packet BBS forwarding:

1. The existing standard software does not take the realities of HF propagation into account. For example, if a message transfer is stopped due to the band closing, the transfer must be restarted in its entirety.
2. The existing standard hardware uses a modem standard that was designed for telephone lines, not the noisy, multi-path HF environment.
3. Most forwarding programs don't take advantage of the full eight bits provided by AX.25.[2] Compressed files are not common.
4. The HDLC frame format causes long frames to be rejected even if only a single bit is in error. Signal paths with a poor BER result in very low throughput, placing an artificial limit on message sizes due to item 1 above.
5. Lack of a true end-to-end transport protocol results in occasional damage to messages, such as "*** connected" appearing in the middle of a message.

Some of the problems with forwarding are not limited to HF, but they do more harm there. The root problem is that a system designed for VHF/UHF keyboard-to-keyboard communication was forced into computer-to-computer service, and then forced into HF service. While this may have been a good use of the tools that were available at the time, new tools are now available. It may be time for a change.

Some changes have already begun. AMTOR MBO/BBS SysOps have addressed part of the Baudot versus ASCII problem.[3] AMTOR, while not a panacea, was at least designed with HF in mind. The new CLOVER system addresses the modem issues. I would like to address the other issue, more efficient high-level data transfer.

Starting from Scratch

In 1988, Jeff Ward, GØ/K8KA and I took on the task of designing the high-level data transfer protocols to be used for the AMSAT and UoSAT Low Earth Orbit (LEO) store-and-forward spacecraft. The initial design discussions for the store-and-forward missions, begun in 1983, had always been based on the WØRLI forwarding scheme.[4] The spacecraft would run a keyboard-access BBS, perhaps even a direct clone of an RLI BBS. Several events changed our minds.

The first was that the UoSAT spacecraft was going to run 9600 baud instead of the 1200 baud the original designs called for. We could not imagine a keyboard user being able to make effective use of this speed without a great deal of computer assistance. Writing software to emulate a keyboard user is always a painful task, so a more computer-friendly transfer mechanism would be desirable.

Next, delays in the PACSAT program put the Japanese FO-12 spacecraft in orbit first. Their BBS design was based on the keyboard entry method that had been discussed for PACSAT. While power problems limited the access time to the FO-12 BBS, we could see that the limitations of the access method kept the number of messages small, the size of messages low, and the content mostly text. No serious attempts at automatic message forwarding were made for FO-12, or for the follow-on FO-20 flight.

Finally, we had learned some lessons from the 1984 UO-11 Digital Communications Experiment. We quickly saw that the 14-minute access times offered by a low orbit meant that we could not live with the keyboard-entry standard that required a message to be completely uploaded (or downloaded) in one connection. The spacecraft seemed to always go over the horizon in the middle of a transfer. If we had to restart the transaction from the beginning of the next pass, some of the previous pass would be wasted.

We decided, therefore, to start from as clean a slate as possible and design a message transfer mechanism that would take the real-world limitations (and advantages) of the LEO environment into account. We still felt we wanted to make use of the existing installed base of TNCs, which meant HDLC and AX.25. We defined the following design drivers:

1. No real-time data entry by humans would be required. To take maximum advantage of the limited pass time, all upload and download activities could be pre-scheduled.
2. Uploads and downloads could be continued from any point in the process, so that any data that had been transferred in a previous orbit would not have to be

resent in a subsequent orbit, even if that occurred days later.

3. To make the best use of the available bandwidth, compression would be possible and encouraged. To make the best use of the resource, binary files would be possible (sound files, images, etc).

To make the implementation as simple as possible, while making no restrictions on the data, we came up with this idea: All data would be transferred as files. There would be no message transfer, as such. While it sounds obvious, this was radically different from message transfer as it then existed on the amateur bands. While some files might contain messages (email), others might not. In any case, the spacecraft "BBS" would not get involved. There would be no parser onboard that looked for "s nk6k@wb6ymh" at the start and "/ex" at the end, or any of countless other variations. The "BBS" portion of the spacecraft code consists of a file system emulator and file transfer protocol, nothing more.

To allow a file (be it message, .GIF, .JPG, .VOC, .ZIP, or what have you) to pass transparently through the file server, we designed a file encapsulation scheme. A standard header is preappended to each file. The header provides fields for spacecraft specific information, and for information about the file. The file's information includes its creation and modification dates, the name of the file on its native system, titles (if a message), keywords, forwarding addresses, user definable fields, and more. The spacecraft fields include file size, upload and download station addresses, and checksums. The header is added by the uploading system, and stripped by the downloading system.

With this information, the file transfer system can function as an end-to-end transport mechanism. The file's checksum can be verified on the spacecraft when received, and again on the ground when delivered. The file transfer protocol allows for continuation of a file upload or download. The header allows for further automation of file downloading. By requesting that only the header of the files be downloaded, the user can use off-line time to select files for complete download. More sophisticated software can immediately examine the file header for destination, title, keyword, and other criteria, and automatically start a complete download.

The complete set of protocols that define the PACSAT file transfer mechanism are in the *ARRL Computer Networking Conference Proceedings*[5] and are available on the CompuServe HAMNET data library 5 as file pacdoc.zip. These protocols were designed with the idea that they could also be used in terrestrial applications. Several implementations of the client portion of the software exist; the server portion is similar.

To be sure, we've had some comments on breaking with the keyboard-access tradition. The file sizes (more than 300 kbytes in some cases), and the varied contents, even on the 1200-bps spacecraft, have shown that we took the right path. These types of files are not regularly transferred on any terrestrial amateur packet BBS system. We've also had the recommendation to use TCP/IP protocols, and use long time-outs to handle the file continuation requirements. We didn't want to maintain that much state information for many users over many days, however.

What has this to do with HF?

On the basis of the number of users and the amount of traffic handled by the PACSAT spacecraft, we feel the break with the past was justified. As a general observation, a new look at solving the unique problems of HF forwarding, eight years after it started, should result in some improvements. Specifically, I believe that techniques similar to those used on the satellites, in particular the file transfer model, can be applied to the HF network as well. Now may be the best time for a new look. New HF modem technology is almost available. New digital subbands in the band plan and other emerging technologies require us to make the most effective use of the available bandwidth. Let's take another look.

Space here does not permit a discussion of another of the satellite protocols, the "Broadcast" protocol. This protocol, used almost as is, could reduce some of the VHF gridlock experienced in large metropolitan areas. Overnight, your local BBS can transmit all the new messages it received in the last 24 hours such that your station can select and keep the messages you wanted to read. All the other users on the frequency can make use of the same transmission, requiring the data to be sent only once. If 40 users wanted to read the same file, this results in a 40:1 reduction of required bandwidth. Transmission is done in a way that allows you to request a retransmission of only those parts of a file that you missed due to random noise. This is the way most of the hundreds of PACSAT satellite users get files from the spacecraft. See the *PACSAT Broadcast Protocol* description in pacdoc.zip or see note 5.

I intend to continue working with some active BBS gateway SysOps and make some firm proposals for use of the PACSAT protocols in terrestrial applications. See you in two months.

Notes

1 Joe Subich and Tom Clark, quoted in "Gateway: Automatic HF Digital Forwarding Recommended," *QEX* 129, November 1992, pp 22.

2 The subject of eight-bit data on HF has Part 97 ramifications and will be the subject of a future column.

3 Victor Poor, "ASCII over AMTOR," *QEX* 129, November 1992, pp 18.

4 John Markoff, "Bulletin Boards in Space," *BYTE*, May 1984, pp 88-92. This is an interview with Harold Price, NK6K, and is a good look at the PACSAT designs of that time.

5 Harold Price, Jeff Ward, "PACSAT Protocol Suite—An Overview" and four companion papers, *Proceedings of the ARRL/CRRL 9th Computer Networking Conference*, ARRL 1990, pp 203-252.