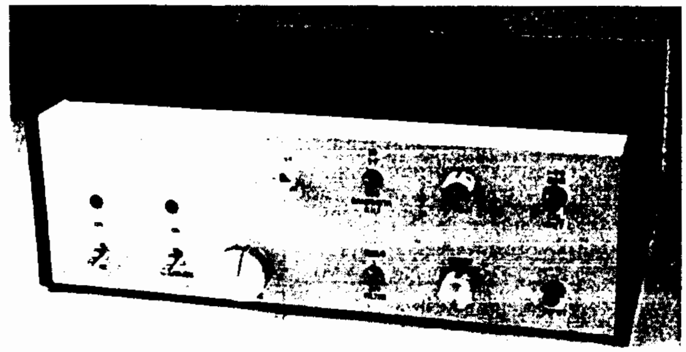
By Wes Hayward, W7ZOI, and Terry White, K7TAU

A Spectrum Analyzer for the Radio Amateur

Good tools are priceless when you need them. Here's a piece of test equipment you've always wanted for your workbench. Now you can have it—without spending a fortune.



Among the many measurement tools sought by the amateur experimenter, the most desired—but generally considered the least accessible—is the *radio-frequency spectrum analyzer* or SA. This need not be. Simple and easily duplicated, this home-built analyzer is capable of useful measurements in the 50 kHz to 70 MHz region. The design can be extended easily into the VHF and UHF region with methods outlined later. The instrument is configured to be self-calibrating, or capable of calibration with simple home-built test gear.[1]

[1]Notes appear on page 43.

We often read and hear about "simple designs." Simplicity implies that something is eliminated to make the equipment easier to build, use or afford. Unlike designs that sacrifice performance for cost and simplicity, this one sacrifices only convenience, while retaining the capabilities needed for accurate measurements.

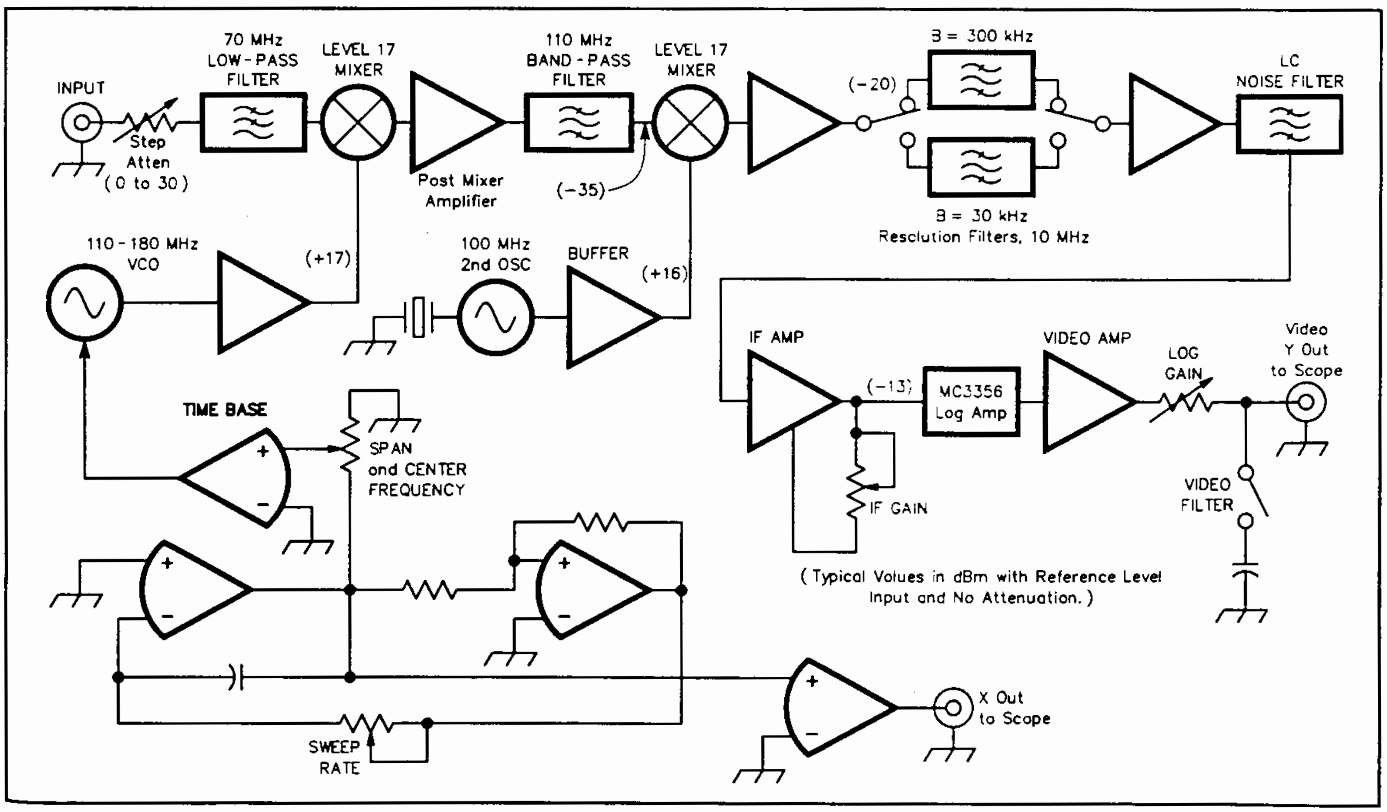


Figure 1—Block diagram of the spectrum analyzer. The circuit is a double-conversion superheterodyne design with intermediate frequencies of 110 and 10 MHz.

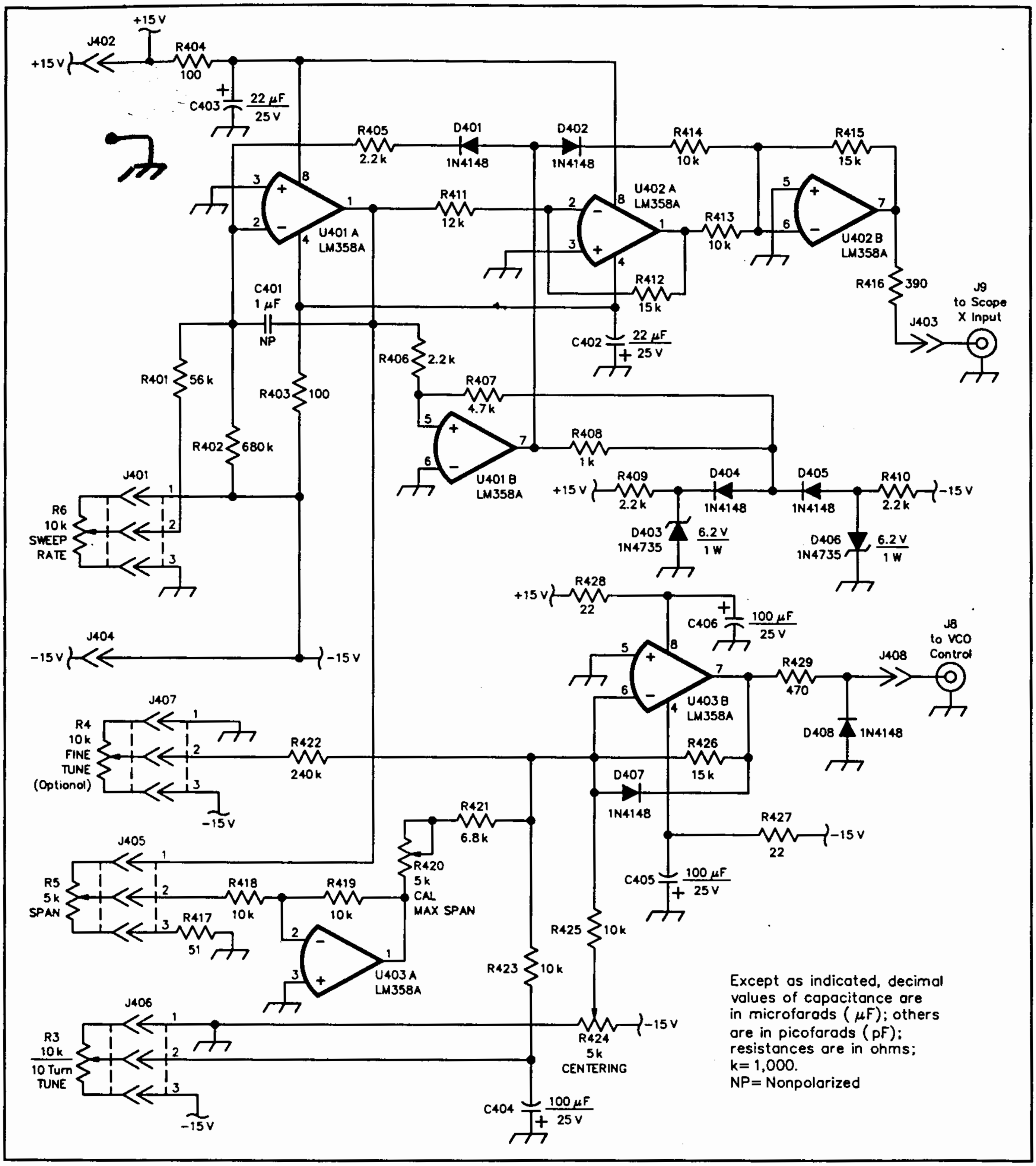


Figure 2—Time base for the spectrum analyzer. Refer to the text for a discussion of the various circuit functions. Front-panel controls include **SWEEP RATE, SPAN** and **TUNE**. Unless otherwise specified, resistors are 1/4 W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted.

U401, U402, U403—LM358 op amp
D403, D406—6.2 V Zener diodes, 1 W
C401—Metal film or Mylar, 1.0 µF capacitor
R420, R423—PC-mount trim pots, 5 kΩ or 10 kΩ suitable
R3, R4, R5, R6—Panel-mounted linear control, 5 kΩ or 10 kΩ suitable. If a 10 turn pot is used for R3, R4 is not needed.

Modern technology eases the construction of this spectrum analyzer. The logarithmic amplifier uses an IF amplifier IC found in cellular telephones and includes a *received signal strength indicator* (RSSI) function. Hybrid and monolithic IC building blocks are employed extensively. These include mixers, amplifiers and VCOs—all vital elements in an analyzer. Finally, it is a rare devoted experimenter today who does not own an oscilloscope. With good basic 'scopes available for about the price of a hand-held FM transceiver, every experimenter should have one. Our spectrum analyzer uses a 'scope as the display. There are no special requirements for 'scope performance other than an X-Y mode with dc coupling in the X and Y axes.

Some Spectrum-Analysis Basics

The RF spectrum analyzer is essentially a swept receiver with a visual display. The display shows the strength of all signals within a user-defined frequency span. Each signal is represented by a line or blip that rises out of a background noise, much like the action of an S meter. Commercial analyzers are calibrated for signal power, with all signals referred to a reference level at the top of the screen. Our analyzer is designed for a basic reference level of –30 dBm, a common value in commercial analyzers.[2]

Signal levels are read from the display by noting that power drops by 10 dB for each major division on the 'scope. You can change the reference level. Adding gain to the analyzer moves the reference to lower levels; introducing attenuation ahead of the instrument moves the reference to higher power levels.

Circuit Overview

Figure 1 is a block diagram of our spectrum analyzer. A double-conversion superheterodyne, it begins with a step attenuator, followed by a low-pass filter and the first mixer, where incoming signals are upconverted to a 110 MHz first IF. After some gain and band-pass filtering, a second conversion moves the signals to a 10 MHz IF. The resolution bandwidths available are 30 kHz and 300 kHz. A video filter smooths or averages noise. The available frequency spans range from a per-division maximum of 7 MHz to about 50 kHz. The center frequency can be adjusted over the entire 70 MHz range. An uncalibrated **SPAN** control allows expansion of the display about the screen center. An uncalibrated **SWEEP RATE** control allows the sweep to be controlled and matched to a given span while avoiding excessively fast scans that could introduce errors.

Ideally, a receiver's first IF should be greater than twice the highest input frequency, a design rule that we bend in this application. The input tuning range includes all HF amateur bands and 6 meters. (We'll discuss higher tuning ranges later.) We picked the 10 MHz second IF because surplus-crystal filters and LC filters for this frequency are easily built. You can easily adapt the design's IF to 10.7 MHz, or other close, convenient values.

The swept LO tunes from 110 to 180 MHz with a commercial VCO module. The VCO output is amplified to drive a high-level-input mixer. The commercial VCO is a recent modification to a design that started with a homebrew oscillator.[3]

Amplifiers are included at the 10 and 110 MHz IFs. These establish signal levels that properly match the log-amplifier window, while preserving system dynamic range. The proper distribution of gain, selectivity and signal-handling capability (intercepts) of the amplifiers and mixers is vital to achieving good performance in a spectrum analyzer, and indeed, any receiver. A proper design will have the same number of stages as a poor one, but will probably use different components and consume more current.

The analyzer uses a ±15 V power supply. The positive supply delivers about 0.5 A. The negative supply current drain is under 50 mA.

Following sections present the circuit blocks in greater detail, in the order that they should be built. The partial but growing system can then be used to test the other sections as they are built, turned on and integrated. We *strongly discourage* building the entire analyzer before testing specific sections. Such an approach may work for casual kits, but is *not suitable* when careful control of signal levels is required. That approach also robs you (the builder) of the excitement of the process: the learning that comes from detailed examination.

Before jumping into the circuit details, we reemphasize that this analyzer—although simple—is intended for *serious* measurements. This means that a normal maximum span display contains no spurious signals. When clean (well-filtered, harmonic-free) signals are applied to the analyzer, there should be *no* extra products as long as the signal level is kept on screen. This performance goal applies for a single tone, or for two equal signals at the top of the screen.

Time Base

Figure 2 shows the analyzer time base, designed for basic functionality without frills; the result is a circuit using only a handful of op amps.[4] U401A and U401B form a free-running sawtooth generator, a circuit commonly found in function generators. U401A operates as an integrator; current is pulled from the inverting input through a 56-kΩ resistor connected to the **SWEEP RATE** control. This current must flow through the capacitor (C401), creating a linearly changing op-amp output voltage. This ramp is applied to U401B, a regenerative comparator, which provides a reset signal to the integrator. The sawtooth waveform (pin 1 of U401A) is asymmetrical: The positive-going ramp grows with a slope determined by the front-panel-mounted **SWEEP RATE** pot, while the negative-going, faster reset ramp is determined by fixed-value components.

The U401 ramp is used twice. U402A and B process the ramp to generate a signal that drives the 'scope's X axis. The signal has a 0 V-centered range with just over a 10 V total swing. Some of the "square wave" from the basic time base (U401B, pin 7) is added to the input of U402B to cause the sweep to reset quickly, even though the return sweep for the VCO occurs in a more stable, smooth way. A slight overscan is generated for the X axis, serving to hide an aberration occurring near the sweep beginning.

The sweep also generates the signal that controls the VCO. The sweep signal (U401A pin 1) is applied to a **SPAN** control. When the analyzer is set for maximum span, the VCO voltage (about 2 to 10 V) generates a sweep from 110 to 180 MHz. The VCO uses only positive sweep voltages, so the output of U403B is diode-clamped to prevent negative output. The center frequency **TUNE**, **FINE TUNE** and a **MAX SPAN** calibration pot set up the proper sweep for maximum span. As the span is reduced with the **SPAN** control, the sweep expands on (or zeroes in on) whatever appears at the center of the screen, determined by the tuning. The center frequency must be set for 35 MHz at maximum span, which coincides with having the zero signal, or

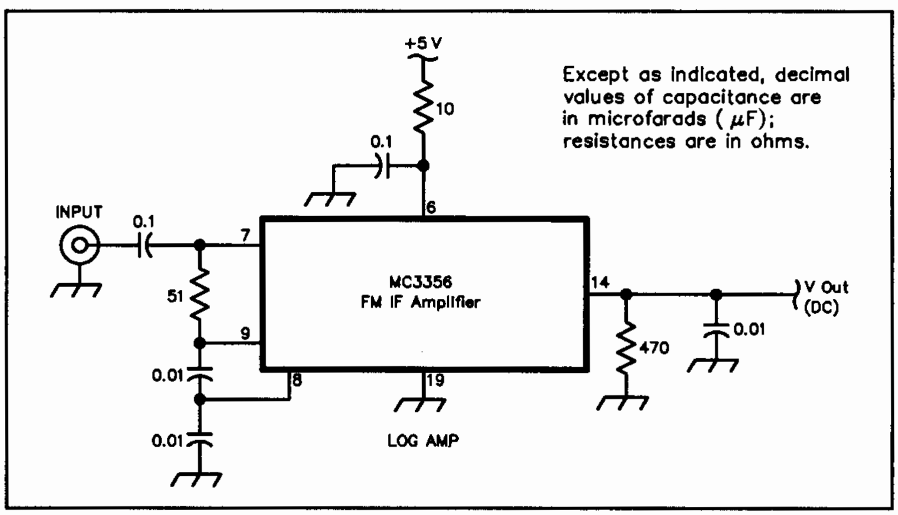


Figure 3—An experimental logarithmic amplifier breadboarded to evaluate performance prior to analyzer construction. You may want to duplicate this circuit and analyze its performance if you decide to use other log-amp ICs.

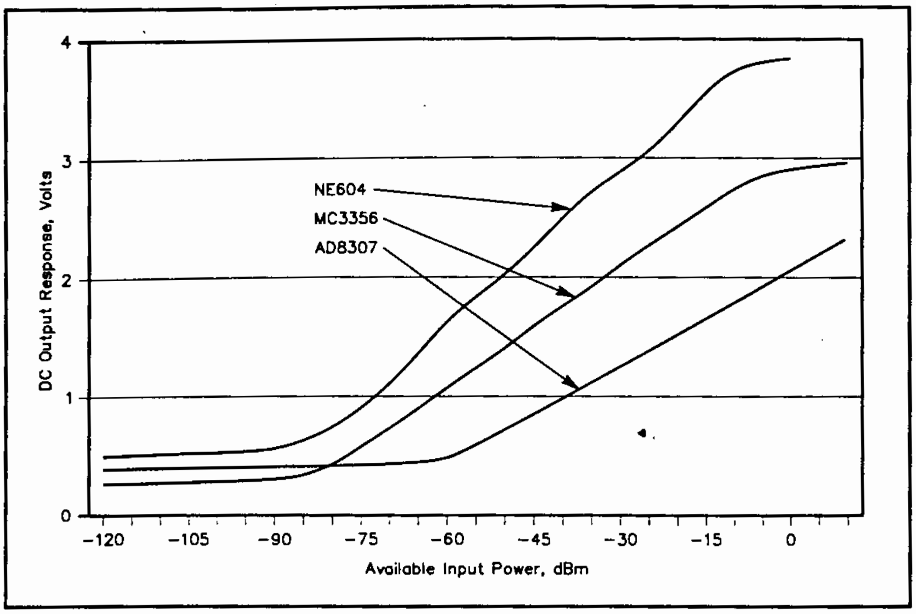


Figure 4—Transfer characteristics for three different logarithmic amplifier ICs. Although the MC3356 is used in our analyzers, use of the AD8307, shown in the lower curve, is recommended. Some curves have been linearly scaled to ease comparison.

"zero spur" at the left edge of the screen.

Setting up the time-base function is generally straightforward. The 'scope can be used to debug, check and study the circuits. The X-axis signal is a ramp ranging from −6 to +6 V with a reset to −15 V during the retrace. A similar ramp appears (without a reset pulse) at the VCO output, but with an amplitude dependent on the **SPAN** control setting.

Although the op amps are carefully bypassed, and the signal that tunes the VCO is shielded, most circuits are noncritical. Normal op-amp circuit precautions are taken with resistors injecting signals into inverting inputs positioned close to the op-amps.[5]

A 10-turn front-panel-mounted pot is used for the **TUNE** control (any value from 5 kΩ to 50 kΩ is suitable). A single-turn pot can be substituted if a 10-turn pot is not available. A fine-tuning function is included in this design, but may be omitted if a 10-turn pot is used for the main tuning.

Log Amplifier and Detector

Central to any spectrum analyzer is a *logarithmic amplifier*. The need for logarithmic processing becomes clear if we consider the range of signals we want to measure: At the low end, we may want to look at submicrovolt levels: under −107 dBm in a 50-Ω system. At the other extreme, we may want to measure the output of small transmitters, perhaps up to a power of 1 W, or +30 dBm. The difference between the two levels is 137 dB. The human ear is capable of handling linear ranges well over 60 dB.[6] This is a wide dynamic range world and linear displays, such as our screen, are inadequate unless some form of data compression or logarithmic processing is used.

The circuit element we use for this processing is the log amplifier.[7] The term is a misnomer, for the usual log-amp IC is both a logarithmic processor (amplifier) and a detector. The chips provide a dc output voltage that increases in proportion to the logarithm of the input amplitude. The central sensitivity specification for a log amp is a voltage slope that is equal to the voltage change (per decade or per decibel) of input-voltage-amplitude change.

An experimental log amp is shown in Figure 3. We breadboarded and tested this circuit to evaluate the log IC. To produce the MC3356 curve shown in Figure 4, the 10 MHz output of an HP-8654 signal generator was applied through HP-355 step attenuators. Exact dc output levels are insignificant, for they can be adjusted with dc voltage gain in a following amplifier. The salient detail that we observe is the dynamic input window. The MC3356, with a 50-Ω input termination, produces a nearly straight-line output voltage versus input power for inputs in the −80 to −10 dBm range. Hence, the analyzer log amp should operate with an input signal of −10 dBm for signals at the top of the screen.

We evaluated two other ICs. One, the commonly available NE/SA604, shows considerable ripple. The best performance offered came from a recently introduced chip from Analog Devices: the AD8307. This IC is designed specifically for measurement applications and offers outstanding logarithmic accuracy, a dynamic range exceeding 90 dB and better temperature stability than found with the usual cellular-receiver chips. The AD8307 requires a high drive level, so it must be preceded with higher-power amplifiers or impedance-transforming networks. The bandwidth of the AD8307 is about 500 MHz, so care is required in its use.

Our analyzer uses the inexpensive and readily available MC3356 log amp shown in Figure 5.[8] An op amp, U303, used to increase the signal output to 0.5 V per division, follows the log chip, U301. The 0 V level corresponds to the bottom of the screen; a signal of 4 V brings the response to the top of the screen. The op-amp output is slightly higher than this, but is then attenuated with a **LOG AMP CAL** control, R2. This pot should be accessible from the outside of the instrument.

The log amp is preceded by an IF amplifier, Q301 through Q303. These stages are biased for relatively high-current operation to preserve linearity. Gain is controlled through variable emitter degeneration in the form of a PIN diode, D301. Most common 1N4000-series power rectifiers work well for gain control. The **IF GAIN ADJ** control (R1) should be available from the exterior of the RF-tight amplifier box. We have placed it on the front panel of our analyzers.

Calibration of the IF and log amplifier is straightforward. First, set the 'scope's Y axis to 0.5 V/division and short it. Set the now-working time base to drive the X axis and adjust the 'scope's vertical position control to place the horizontal line at the bottom of the screen. Inject a −10 dBm signal from a signal generator into the log amplifier input, remove the short circuit and adjust R2, **LOG AMP CAL**, for a full-screen (reference level) response. The input level is next reduced in 10 dB steps. The horizontal sweep line should drop down one major division for each 10 dB reduction over a 60 dB range. If this does not happen, repeat the procedure with a slightly different drive level. In our analyzers, a typical drive level of −13 dBm produced good accuracy.

Now, attach the IF amplifier to the log amplifier and drive them with an input level of −23 dBm. Peak the IF output filter for maximum response and set R1, **IF GAIN ADJ**, for a full-screen response. A true filter peak can be confirmed by varying the generator frequency. There is considerable extra range in the **IF GAIN ADJ**, providing extra flexibility during use.

Resolution Filters

Continuing the backward progression through the system, we encounter the resolution-bandwidth-determining filters. Our analyzer uses bandwidths of 30 and 300 kHz, provided by crystal and LC filters, respectively. The 300 kHz LC filter, the crystal filter and the relay circuitry for bandwidth switching are shown in Figure 6. Although shown as individual modules, they can be incorporated into one. The PC board for the filter includes the LC filter and switching relays with room for a user-selected crystal filter. Builders may want to implement their own scheme here. We

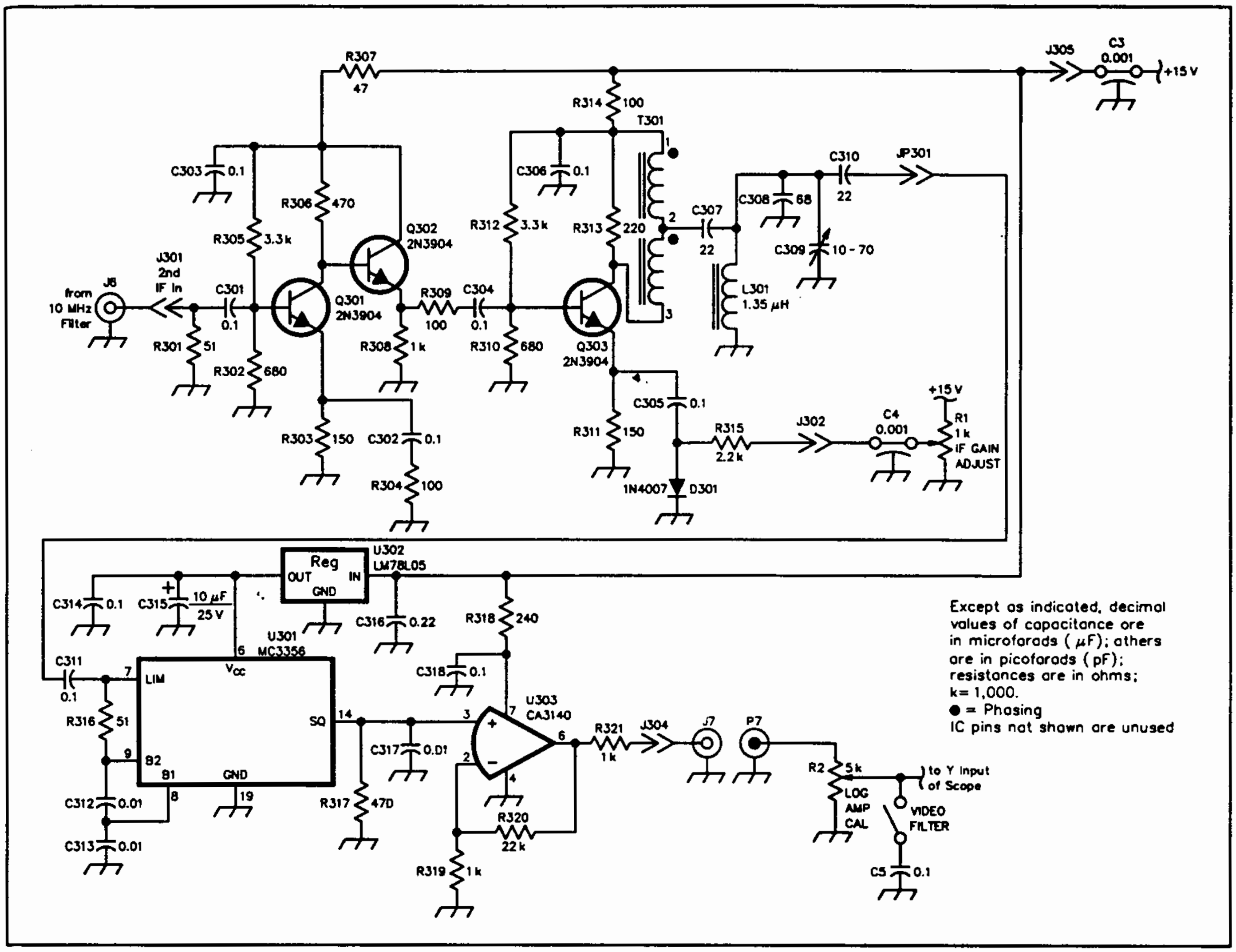


Figure 5—The 10 MHz IF amplifier and log amplifier used in the analyzer. Refer to the text for adjustment details. Unless otherwise specified, resistors are ¼ W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted.

C309—Plastic dielectric trim cap (Sprague-Goodman GYD65000)
C307, C308, C310—Silver mica or NPO ceramic capacitors, 10% tolerance
C316—0.22 µF ceramic
D301—PIN diode; 1N4007 used
L301—1.35 uH, 18 turns #24 enameled wire on T-44-6 core, Q >150
Q301, Q302, Q303—2N3904
R1—Panel-mount, 1 kΩ linear
R2—Panel-mount, 5 kΩ linear
U301—Motorola MC3356
U302—78L05 +5 V regulator
U303—CA3140 op amp

reasoned that builders would want to implement their own ideas. Maintain reasonable shielding for this part of the system. Additional attenuator pads can be inserted in line with one filter or the other to approximately equalize filter loss in the two paths.

You may want to build crystal filters for your analyzer.[9] The VCO stability in this analyzer will support resolution bandwidths as narrow as 3 to 5 kHz. For a simplified beginning, a very practical analyzer can be built with only one resolution bandwidth of 300 kHz.

Second Mixer and Second Local Oscillator

Figure 7 shows the second mixer and related LO. The heart of this module—and to some extent that of the entire analyzer—is U202, a high-level second mixer. This mixer is bombarded by large signals that are as strong or stronger than those at the front end. Accordingly, the second mixer should have an intercept similar to that of the first mixer. This is the usual weak point in all too many homebrew spectrum analyzers—as well as more than a few receivers! The second mixer, U202, uses a +17 dBm level Mini-Circuits TUF-1H. This is *not* the place for a current-starved telephone component! The second mixer is terminated in a high-pass/low-pass diplexer followed by an IF amplifier (Q202) biased at 50 mA. This is a critical stage for dynamic range: Don't replace it with a monolithic substitute of reduced gain or intercept.

The second LO begins with a 100 MHz, fifth-overtone crystal oscillator (Q201), followed by a pad and a power amplifier. The oscillator inductor, L201, in Q201's collector is made of five turns of #22 wire wound on a 6-32 machine screw. (Remove the screw before installing the coil.) Here's an excellent way to align the oscillator: Temporarily replace the crystal, Y201, with a 51 Ω resistor. Adjust the tuned circuit until oscillation occurs at the desired 100 MHz frequency. Then, replace the 51 Ω resistor with the 100 MHz crystal; no further tuning is required. Measure the oscillator's output with a power meter before applying it to U202. Adjust the pad attenuation (R205, R206, R207) to realize the specified LO drive level.

After the second LO is operating, attach it to the second mixer and the rest of the analyzer. With a second mixer input of –35 dBm at 110 MHz, you should obtain a reference-level response.

Voltage-Controlled Local Oscillator and First Mixer

Figure 8 shows the analyzer's swept LO. The foundation for this module is a Mini-Circuits POS-200 VCO module, U101. Similar VCOs are available from many

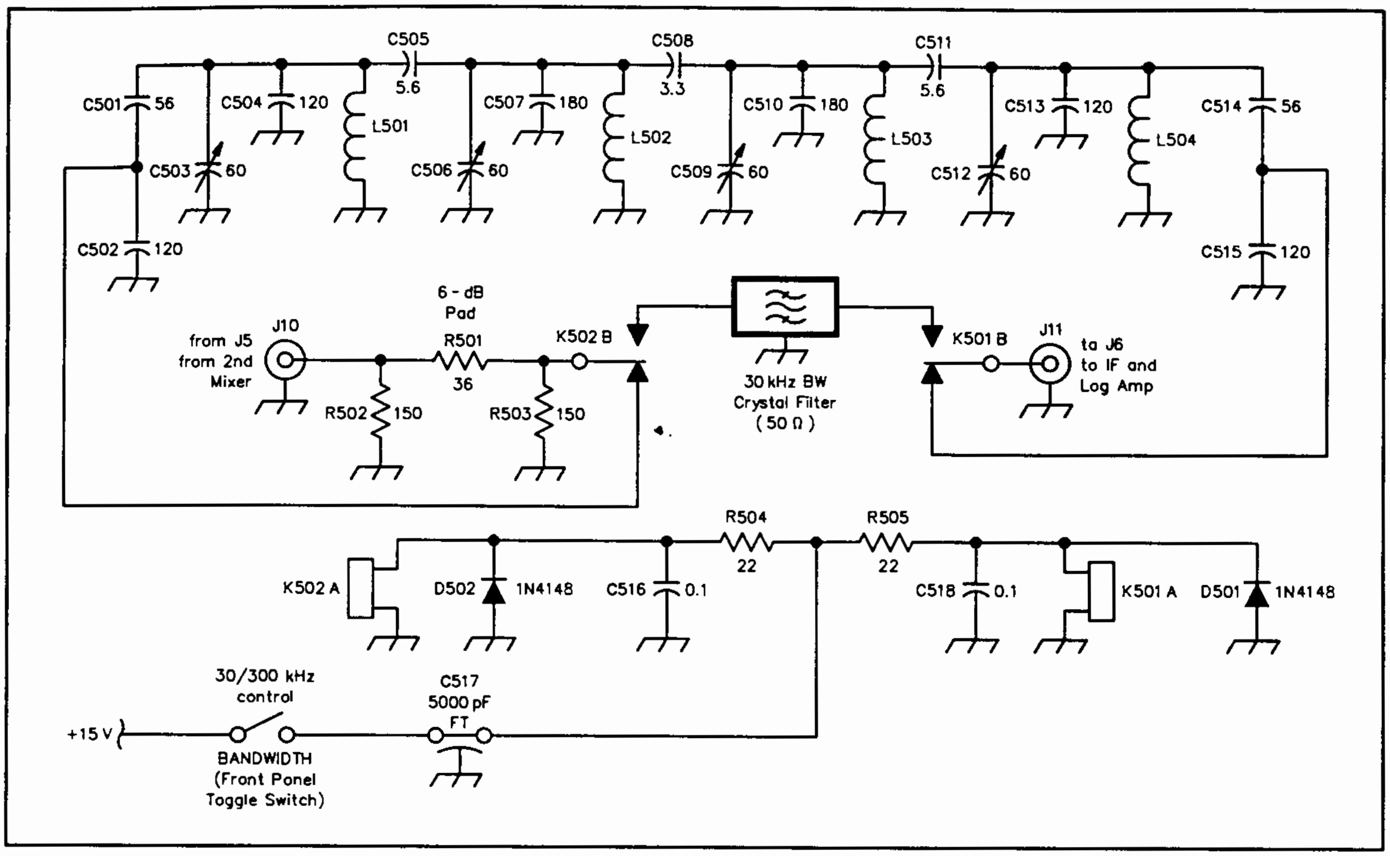


Figure 6—Resolution filters: The upper schematic shows the 300 kHz bandwidth 10 MHz LC filter. If desired, that circuit can be realigned at 10.7 MHz without other design changes. The LC filter is shown as a separate unit connected to the rest of the analyzer with coaxial cable. However, the filter can be constructed on the board with the crystal filter and relays. Unless otherwise specified, resistors are ¼ W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted.

C501, C502, C504, C505, C507, C508, C510, C511, C513, C514, C515—Silver mica or NP0 ceramic, 5%
C503, C506, C509, C512—65 pF plastic dielectric trim cap (Sprague-Goodman GYD65000)
K501, K502—SPDT relay; Aromat TF2-12V used here (one contact set unused); values of associated dropping resistors may need adjustment.
L501-L504—17 turns of #22 enameled wire on a T-50-6 toroid (1.15 µH), Q > 250

vendors.[10] The VCO output is about +10 dBm, too low a level for the high-level mixer. A MAV-11 amplifier, U102, preceded by a pad to provide level adjustment, increases the signal level. Confirm the output power level before applying it to the mixer, U103.

Once the VCO output level is adjusted and confirmed, calibrate its frequency against the VCO control voltage. If a VHF counter is not available, you can obtain a few points by tuning the VCO to hit local FM broadcast signals of known frequency. Calibrating the VCO is useful if the module is used later as a signal source for alignment of the 110 MHz band-pass filter.

Figure 8 also shows the input mixer, U103, another Mini-Circuits TUF-1H, terminated in a 6 dB pad. Although the pad degrades the noise figure, it presents a solid output termination for the mixer. This termination is reflected, helping to provide a good mixer-input impedance match, important in a measurement instrument. The pad is followed by a Mini-Circuits MAV-11 IF amplifier (U104) that restores the gain lost in the mixer and pad.

The mixer application differs from a normal diode ring: The RF input is now attached to the dc-coupled port. This allows input frequencies as low as 50 kHz to be converted to the first IF. The low-frequency end is limited by mixer LO to RF isolation, which determines the LO energy that reaches the first IF. The related on-screen response is often termed the *zero spur*, a familiar "feature" in most RF spectrum analyzers.

This module (VCO and first mixer) is contained in a shielded enclosure with coaxial inputs and outputs, including coaxial routing of the VCO control voltage. The front end is susceptible to any VHF and UHF signals reaching it, making shielding and decoupling especially important.

The 110-MHz IF Band-Pass Filter

One of the more-critical blocks in the analyzer is the filter that establishes the bandwidth of the VHF IF. The bandwidth must be at least as wide as the widest 10 MHz filter, but must be narrow enough to reject the 90 MHz second-conversion images by 80 dB or more. This performance is only available with a three-pole or higher-order filter. The best double-tuned circuits we built (or computer simulated) came close, but just didn't cut it.

We described double-tuned circuits in detail in a 1991 *QST* tutorial paper.[11] Those methods have recently been extended to three-resonator filters.[12] One of the methods presented in the later paper is a sequential approach that begins with a double-tuned circuit (DTC). First, a DTC is built for the desired 3 dB bandwidth and has its performance confirmed with a wideband sweep (a vital requirement!) Then, a third resonator is inserted between the original two. Coupling elements similar to the one that produced the required DTC bandwidth are repeated in the triple-tuned circuit, but end-section loading is not changed. The center frequency of the three resonators is aligned to complete the filter.

The schematic for the 110 MHz triple-tuned circuit is shown in Figure 9. The inductors, 100 nH, are made by winding 5 turns of #18 wire on the shank of a ¼-inch drill bit. These inductors typically have an unloaded Q of just over 200 at 110 MHz. Larger-diameter inductors would have produced higher unloaded Q with the attendant lower insertion loss. However, the

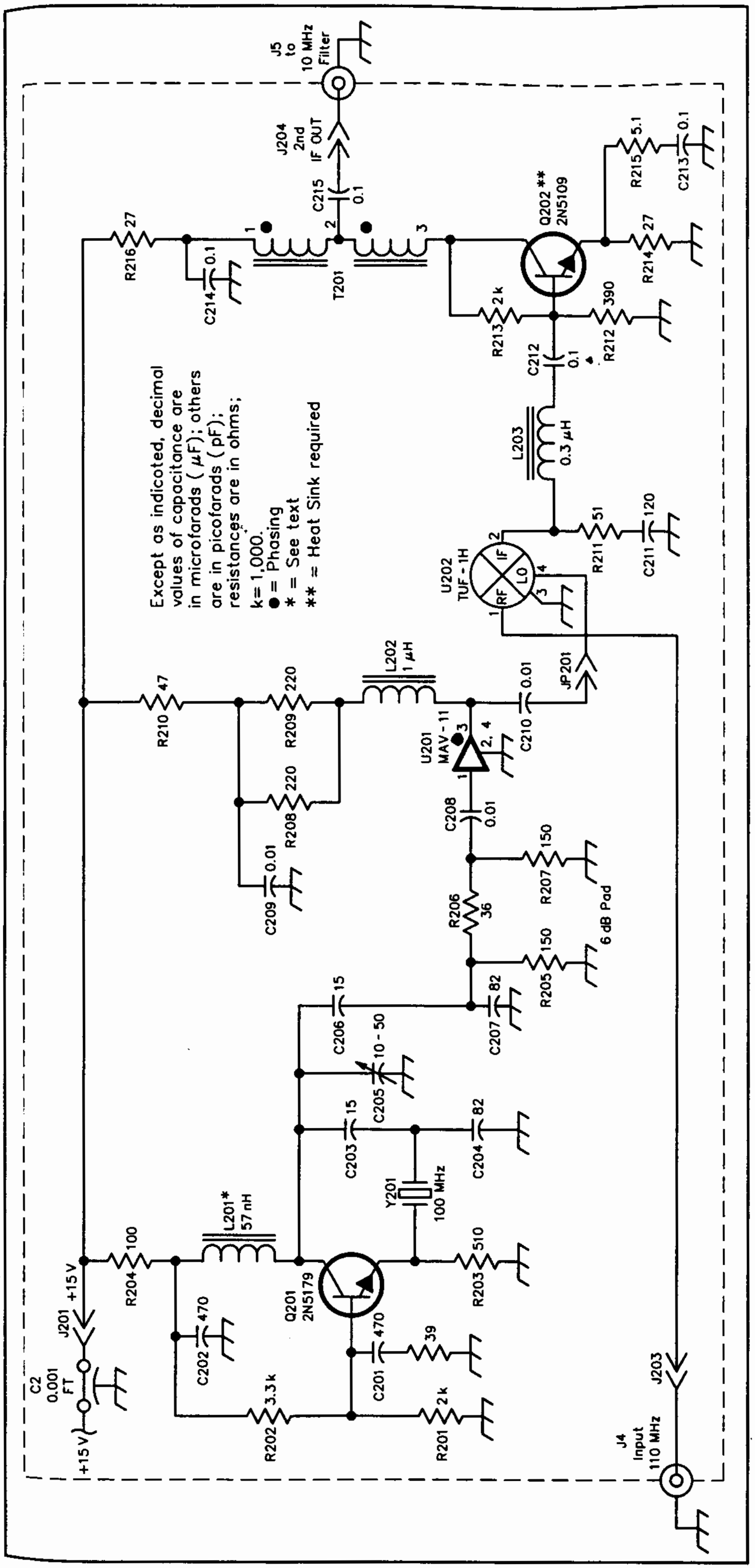


Figure 7—Second mixer and second LO. L203 in the output of U202 consists of 17 turns of #28 enameled wire on a T-30-6 toroid. The actual value is not critical and a molded RF choke can be used in place of the toroid. Unless otherwise specified, resistors are 1/4 W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted.

C201, C202—470 pF ceramic
C203, C206—15 pF NPO ceramic or silver mica
C204, C207—82 pF NPO ceramic or silver mica
C205—65 pF plastic dielectric trim cap (Sprague-Goodman GYD65000)
C211—120 pF, silver mica or NPO ceramic
L201—57 nH; 5 turns of #22 wire wound on a 6-32 machine screw. Remove the screw before installing the coil.
L202—1 μH molded RFC; any value from 100 nH to 2.7 μH is okay
L203—0.3 μH; 9 turns #24 enameled wire on a T-30-6 core
Q201—2N5179
Q202—2N5109, 2N3866, 2SC1252, etc
U201—Mini-Circuits MAV-11
U202—Mini-Circuits TUF-1H mixer
T201—10 bifilar turns #28 on FT-37-43 ferrite toroid

stray coupling between coils would have increased, which would have necessitated shields between filter sections. The smaller (10 nH) end-matching inductors are one-inch lengths of #18 wire. The triple-tuned filter, and its parent DTC, have bandwidths of 2 to 3 MHz.

The filter alignment and experimentation is usually done with a sensitive power meter.[13] a step attenuator and a signal source. As mentioned earlier, the VCO can serve the role of signal generator, if one is not available.

The second-conversion image rejection is easily measured with a finished analyzer. Apply a 40 MHz signal to the analyzer and adjust it for a reference-level response. Don't touch the analyzer tuning, but move the signal generator to 60 MHz. An image signal may appear at the same point on screen as the original 40 MHz signal. The rejection was only 66 dB with a DTC used for some experiments. The triple-tuned filter produced 90-dB rejection. The slight extra effort of the triple-tuned circuit is easily justified. No PC board is available for the IF filter.

Input Low-Pass Filter

A 70 MHz low-pass filter is shown in Figure 10. This circuit and a step attenuator are housed in separate shielded enclosures in one of our analyzers. In the other, the filter and the attenuator remain outboard elements. Integral components are more convenient for routine analyzer applications, but incorporation removes them from the equipment pool available for other experiments. Also, operation without an outboard low-pass filter allows the instrument to be

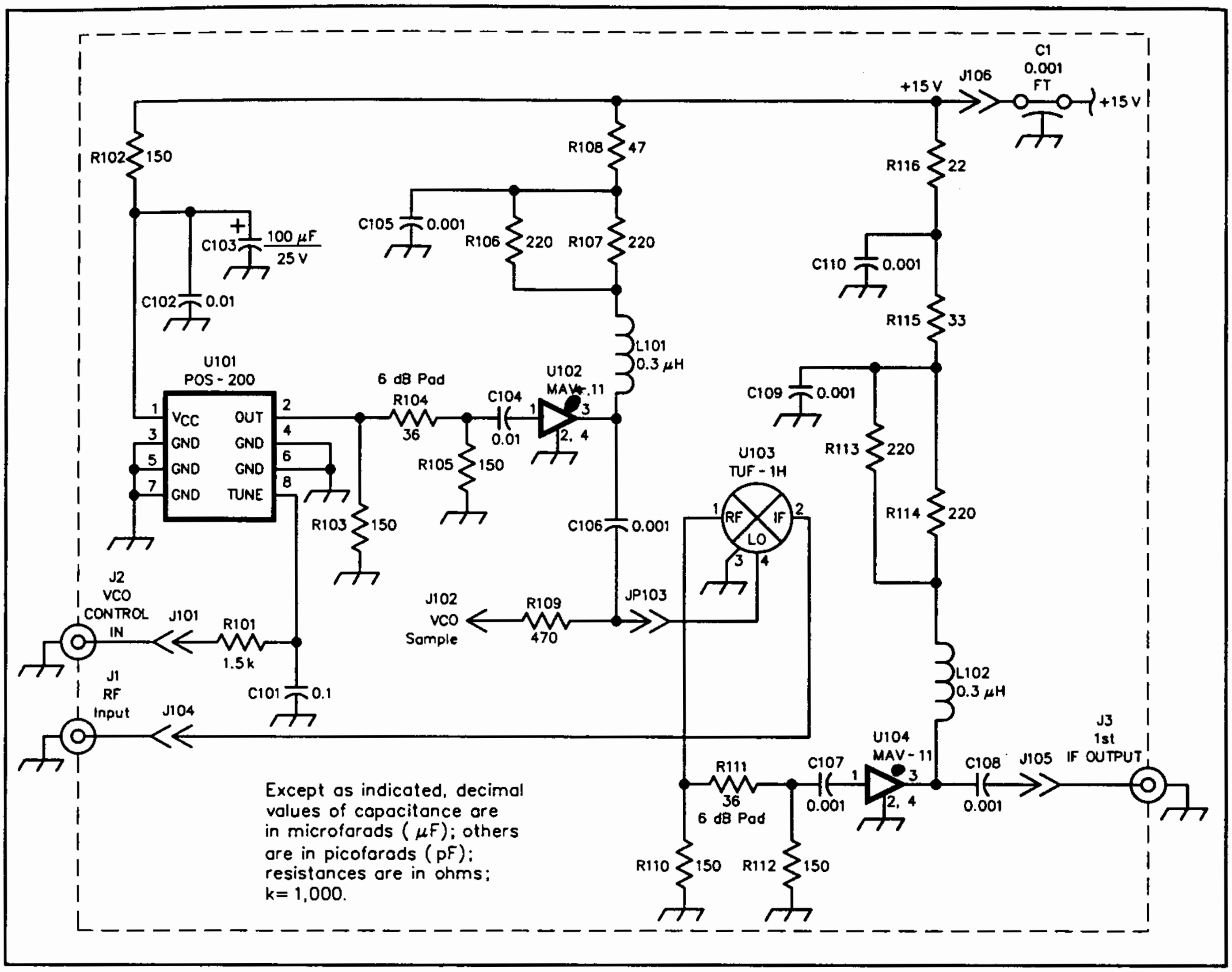


Figure 8—Front-end for the spectrum analyzer. The LO drive level is set to between +16 and +18 dBm by trimming the pad attenuation. The 1 μH inductors used in the MAV-11 outputs can be molded RF chokes or made of 17 turns of #28 enameled wire wound on T-30-6 toroids. Unless otherwise specified, resistors are 1/4 W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted.

L101, L102—1 μH molded RFC; any value between 100 nH to 2.7 μH suitable

U101—Mini-Circuits POS-200 VCO

U102, U104—Mini-Circuits MAV-11

U103—Mini-Circuits TUF-1H

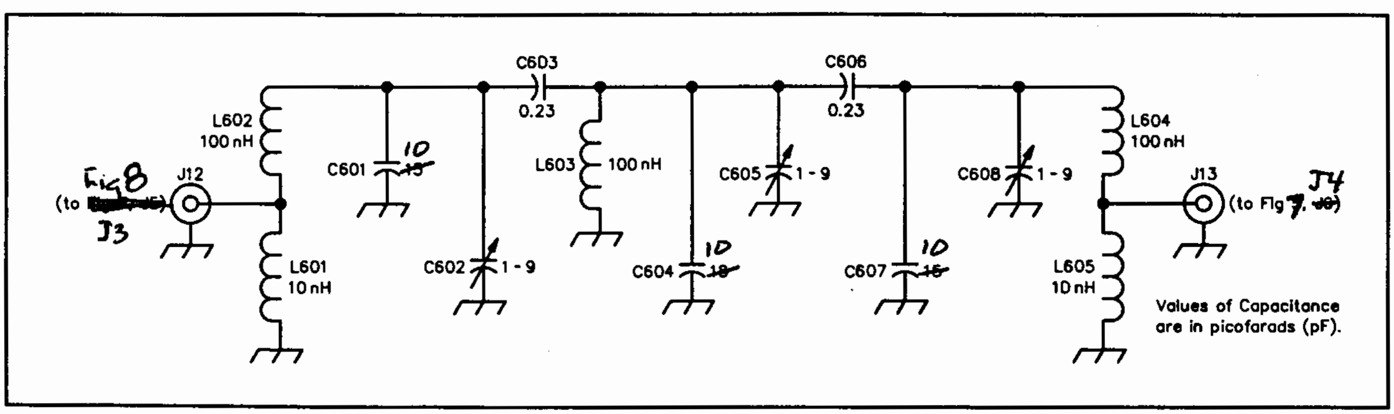


Figure 9—VHF band-pass filter used in the 110 MHz IF.

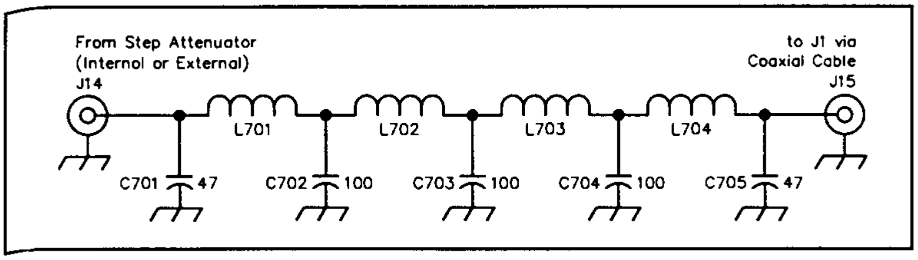


Figure 10—Input 70 MHz low-pass filter. The filter started as a ninth-order Chebyshev design, but was modified through computer manipulation to use equal-value inductors and standard-value capacitors. The inductors each consist of 8 turns of #22 wire. The wire is wound on the threads a 1/4-20 bolt as a form; remove the bolt after winding the inductor. See text.

C701-C705—NP0 ceramic, 5% tolerance.

L701-L704—8 turns #22 bare wire wound in 1/4-20 bolt threads; see text.

used well into the low UHF area by operating the mixer with VCO harmonics. (No circuit boards are available for this filter.)

Construction and Adjustment Hints

The spectrum analyzer can be built using a number of RF techniques. Our analyzers are collections of small boxes using coaxial-cable interconnections. Power supplies reach the interior of RF modules through feedthrough capacitors. The only open boards in the system are the time base and (in one instrument) the log amp, both constrained to low frequencies. The 110 MHz IF filter and all RF circuit boards are designed to fit in Hammond 1590B diecast aluminum alloy boxes.

The sensitivity of such RF measurement equipment justifies the extensive shielding. Spurious responses are readily seen on the display. While a few can be tolerated, they seriously detract from a measurement when, for example, spurious transmitter products are being examined.

Most of the adjustment has already been discussed. Indeed, by the time the analyzer is finished, there is little alignment left! The block diagram (Figure 1) has typical signal levels shown in parentheses.

The finished analyzer is set up with a −30 dBm signal from a stable-impedance source. Adjust the IF gain to establish the reference level. If the 10 dB per division tracking is not quite correct, change the log amp gain, followed by a readjustment of the IF gain until tracking is correct. IF gain is adjusted to the reference level whenever the analyzer is used.

Next month, we'll present methods to extend the analyzer to higher frequencies. We'll discuss simple test equipment that can be used in alignment and present some typical examples of spectrum-analyzer use.

Notes

[1] Homebrew analyzers described in amateur literature in recent years have often used TV tuners as the front end. See, for example, Al Helfrick, K2BLA, "An Inexpensive Spectrum Analyzer for the Radio Amateur," *QST*, Nov 1985, pp 23-29. A more recent example is Fred Brown, W6HPH, "Build a 5- to 850-MHz Spectrum Analyzer," *Communications Quarterly*, Winter, 1997, pp 91-96.

[2] The usual unit for power measurement with a spectrum analyzer is dBm. This is an absolute, impedance invariant unit, signifying power compared with 1 mW. Hence, −30 dBm is 30 dB below a milliwatt, or a microwatt. The dBm unit is especially useful in spectrum analysis where two levels are often compared. The difference between two powers in dBm is a power ratio in dB. (See Jay Craswell, WB0VNE, "Converting Between dBm, Milliwatts and Watts," Technical Correspondence, *QST*, Jul 1998, p 70.—*Ed.*)

[3] Excellent homebrew VCOs can be built using the work of Allan Victor, WA4MGX, "Wideband VCO Design," *Ham Radio Magazine*, Jul, 1984, pp 49-58. See also the recent work of Colin Horrabin, G3SBI, as reprinted in "Tech Notes," *Communications Quarterly*, Winter, 1996, pp 94-104.

[4] Many amateurs are uncomfortable with op-amp circuits, a situation easily remedied by a study of the excellent text by Horowitz and Hill, "The Art of Electronics," 2nd Edition, Cambridge University Press, 1989. See the golden rules presented on page 177. An op-amp integrator is presented on page 222. The information on sawtooth oscillators beginning on page 288 is also useful.

[5] PC boards for several of the analyzer circuits are available from Kanga USA. Details are available at the Kanga Web site, **http://www.bright.net/~kanga/kanga/**, or from Kanga USA, 3521 Spring Lake Dr, Findlay, OH 45840.

[6] The best digital audio recording and playback systems have a dynamic range around 90 dB.

[7] An outstanding fundamental treatment of log amps is given in applications data from Analog Devices. See, for example, the data sheet for the AD8307, Analog Devices, Norwood, MA, 1997, **http://www.analog.com/product/Product_Center.html**.

[8] The log/IF board available from Kanga (see Note 5) is configured for the MC3356. Most of the 20 pins allotted to the IC are not used and their locations can be employed to breadboard and retrofit the AD8307 (8 pin DIP) into the system, if desired.

[9] If crystal filters are built especially for this project, we recommend a peaked response shape (such as Gaussian-to-6 dB). We have investigated wider bandwidth crystal filters as presented in Wes Hayward, W7ZOI, "Refinements in Crystal Ladder Filter Design," *QEX*, Jun, 1995, pp 16-21. See also the careful work of Bill Carver, "High Performance Crystal Filter Design," *Communications Quarterly*, Winter, 1993, pp 11-18, and that of Jacob Makhinson, N6NWP, "Designing and Building High Performance Crystal Ladder Filters," *QEX*, Jan 1995, pp 3-17.

[10] See the Mini-Circuits catalogs and applications manuals. A family of viable VCOs are also available from Synergy Microwave. These units are more expensive, but offer lower phase noise, which would be significant for more-stringent applications.

[11] Wes Hayward, W7ZOI, "The Double Tuned Circuit: An Experimenter's Tutorial," *QST*, Dec 91, pp 29-34.

[12] Wes Hayward, W7ZOI, "Extending the Double-Tuned Circuit to Three Resonators," *QEX*, March-April, 1998, pp 41-46.

[13] Denton Bramwell, K7OWJ, "The Microwatter," *QST*, Jun 1997, pp 33-35. Also, see the sidebar in the referent of Note 11.

Wes Hayward, W7ZOI, and Terry White, K7TAU, are not strangers to QST. *This is also not the first* QST *project on which they've worked together. The first, "The Mountaineer—An Ultraportable CW Station," described a QRP transceiver and appeared in the August 1972 issue. At the time this spectrum-analyzer project evolved, the authors worked together in the Advanced Circuits section of the Receiver Group at TriQuint Semiconductor in Hillsboro, Oregon. Over the years, Wes has provided readers of* QST, The Handbook *and other ARRL publications with a wealth of projects and technological know-how. You can contact Wes at 7700 SW Danielle Ave, Beaverton, OR 97008; e-mail* **w7zoi@teleport.com**. *Terry can be reached at 9480 S Gribble Rd, Canby, OR 97013; e-mail* **twhite@tqs.com**. QST

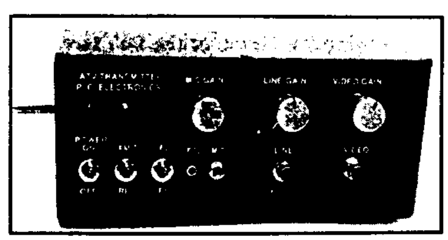
New Products

ATV TRANSMITTER FROM PC ELECTRONICS

◊ The TX70-10 ATV transmitter from PC electronics is aimed at operators who use cable-ready TVs (tuned to channels 57-60) to receive amateur television signals in the 70-cm ham band. The continuous-duty transmitter puts out 10 W AM (color or black-and-white, just like broadcast TV stations) and features a built-in T/R relay, variable power output, front-panel RCA jacks for video and audio (with line and mic inputs), a video monitor circuit and a PTL jack (push-to-look) in parallel with the unit's T/R switch.

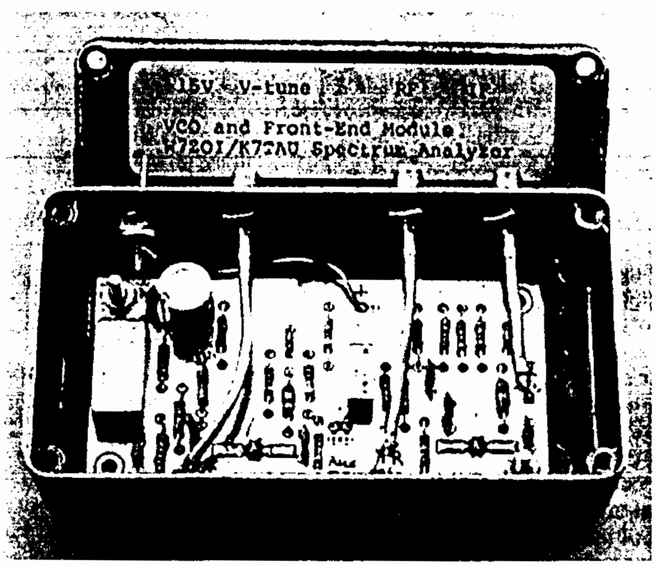
The TX70-10 is housed in a die cast aluminum box measuring about 7×5×3 inches and is powered by 13.8 V dc at 3 A. The unit comes with one user-specified crystal in the 440-MHz range. An optional second front-panel-selected crystal is available for $20. A repeater version, the RTX70-10, is also available.

Price: $439 shipped to the lower 48 states. For more information, contact PC Electronics, 2522 Paxson Ln, Arcadia, CA 91007; tel 626-447-4565; **tomsmb@aol.com**. QST



By Wes Hayward, W7ZOI, and Terry White, K7TAU

A Spectrum Analyzer for the Radio Amateur—*Part 2*



Here are some examples of procedures you can use to become familiar with your new spectrum analyzer.

In Part 1,[14] we described the design and construction of a simple, yet useful spectrum analyzer. This installment presents some applications and methods that extend the underlying concepts.

Amplifier Gain Evaluation

One use for a spectrum analyzer is amplifier evaluation. We can illustrate this with a small amplifier from the test-equipment drawer—an old module that has been pressed into service for a variety of experiments. This circuit, shown in Figure 11, is used for illustration only and is not presented as an optimum design. It's a project that grew from available parts and may be familiar to some readers. The circuit uses four identical 2N5179 amplifier stages. A combination of emitter degeneration and parallel feedback provides the negative feedback needed to stabilize gain and impedance. (Ideally, construction and measurement of a *single* stage should precede construction of the complete amplifier.)

We began the experiment by setting the signal generator at a known power level, −20 dBm. (If a good signal generator is unavailable, you can easily build a suitable substitute.[15] We used a surplus HP8654A for most of these experiments.) With the generator and spectrum analyzer connected to each other through 50-Ω coax cable, we set the generator to 14 MHz with 10-dB attenuation ahead of the analyzer. The amplifier was not yet connected. We adjusted the analyzer's **IF GAIN** control for a response at the top of the screen. Using a resolution bandwidth of 300 kHz allows for a fast sweep without distortion. In our case, the second harmonic was only 26 dB below the peak response. However, when we added a 15 MHz low-pass filter, the second harmonic dropped to −57 dBc.[16] The third harmonic is well into the noise. (It is not unusual for a signal generator to be moderately rich in harmonics.)

Next, we inserted the amplifier between the signal generator and the analyzer, keeping the low-pass filter in the generator output. The on-screen signal went well above the top as soon as the amplifier was turned on. Decreasing the signal generator output to −51 dBm produced the same −20 dBm analyzer signal that we saw before the amplifier was inserted. The gain measured 31 dB.[17] Increasing the analyzer attenuation by 10 dB (for a reference level of −10 dBm) and increasing the generator output to −41 dBm produced the same gain, but a growing harmonic output. The second-harmonic response was now at −43 dBc and a third harmonic appeared out of the noise at −60 dBc.

We continued the process—moving both step attenuators produced an amplifier output of 0 dBm with second and third harmonics at −28 and −36 dBc, respectively. The next 10-dB step, however, didn't work as well, producing gain compression. With a drive of −21 dBm, the output was only +4 dBm, a gain of only 25 dB instead of the small-signal value of 31 dB.

Amplifier Intermodulation Distortion

Next, we measured intermodulation distortion (IMD). The setup for these experiments is shown in Figure 12. Two crystal-controlled sources[18] at 14.04 and 14.32 MHz are combined in a 6 dB hybrid combiner (return-loss bridge),[19] applied to the 15 MHz low-pass filter and a step attenuator having 1 dB steps. This composite signal drove the amplifier, with its output routed to the spectrum analyzer. For this measurement, we dropped the resolution bandwidth to 30 kHz. (The video filter was turned on and the sweep rate reduced until the signal amplitude was stable.) The analyzer's attenuator, set for a reference level of −10 dBm at the top of the screen, was confirmed with a calibration signal from the signal generator. We adjusted the **IF GAIN** to compensate for changes in analyzer bandwidth and for log-amplifier drift.

The output of the two-tone generator system was adjusted to produce a spectrum analyzer response of −10 dBm per tone. The IMD responses were readily seen, now 47 dB below the desired output tones. The output intercept is given by

$$IP3_{out} = P_{out} + \frac{IMDR}{2}$$

where P_{out} is the output power of each desired tone (−10 dBm) and IMDR is the intermodulation distortion ratio, here 47 dB. The output intercept for this amplifier measured +13.5 dBm. This is well in line with expectations for such a design.

When performing IMD measurements, it's a good idea to change the signal level while noting the resultant performance. Dropping the drive power by 2 dB should cause a −2 dB response in the desired tones, accompanied by a 6 dB drop in distortion-product tones. The output intercept

[14]Notes appear on page 40.

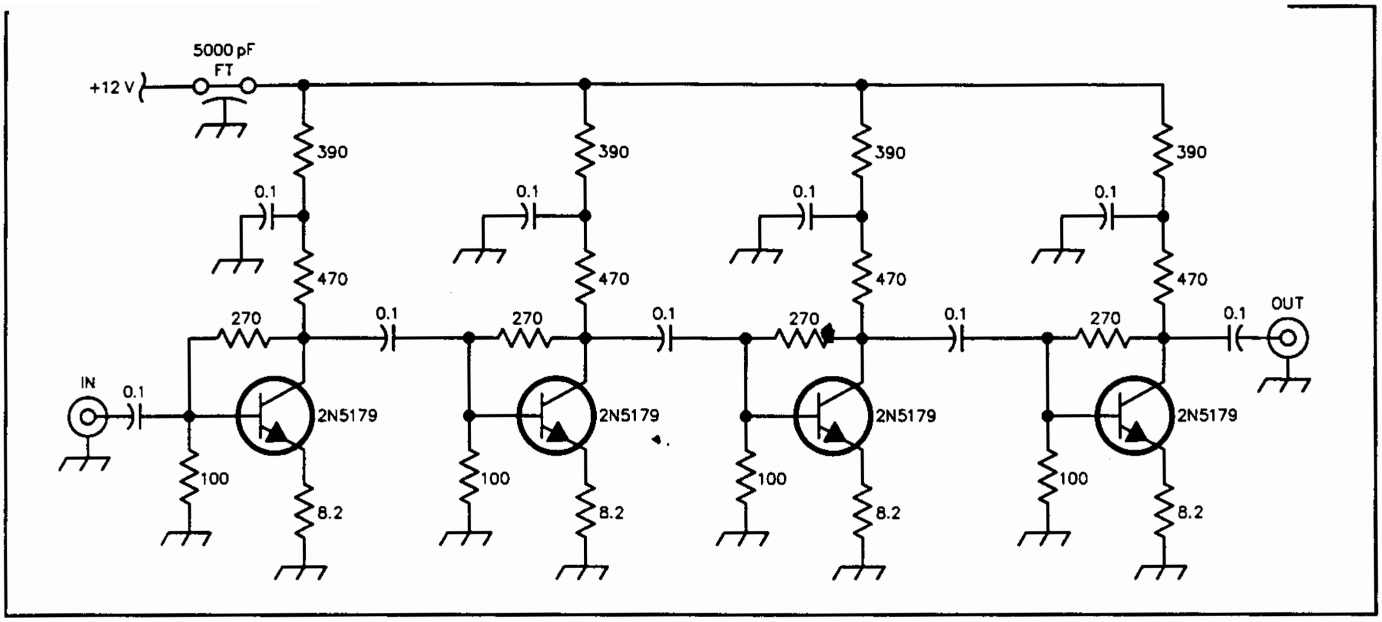


Figure 11—Sample wideband amplifier used to illustrate amplifier measurements.

should remain unchanged.

The IMD in the preceding example was 47 dB below the desired output tones, a value that we obtained by simply reading it from the face of the 'scope, possible because we use a log amplifier that has moderate log fidelity. If the log amplifier was not as accurate as it is, we could still get good measurements. In this example, you would note the location of the distortion products on the display. Then, using the step attenuator, decrease the desired tones until they are at the noted level. The result would be −47 dBc for the distortion level, a measurement that depends solely on the accuracy of the attenuator. This illustrates the profound utility of a *good* step attenuator, an instrument that can be the cornerstone of an excellent basement RF laboratory.

During the third-order output-intercept determination just described, we assumed that the distortion was a characteristic of the amplifier under test. This may not be true. It is important to determine the IMD characteristics of the spectrum analyzer used for the measurements before the amplifier measurements are fully validated. Specifically, for results to be valid, the input intercept of the analyzer should be much greater than the output intercept of the amplifier under test.

The spectrum analyzer input intercept is easily measured with the same equipment used to evaluate the amplifier. The two tones are applied to the analyzer input with no attenuation present at the analyzer front end. Then, the input tones are adjusted for a full-screen response. In this condition, there should be no trace of distortion. Although this is generally an adequate test, it *does not* establish a value for the input intercept. To do that, we must overdrive the analyzer, using signals that exceed the top of the screen.

The following steps were used to measure the analyzer input intercept:

• We calibrated the analyzer for a reference level of −30 dBm with a 30-kHz resolution.

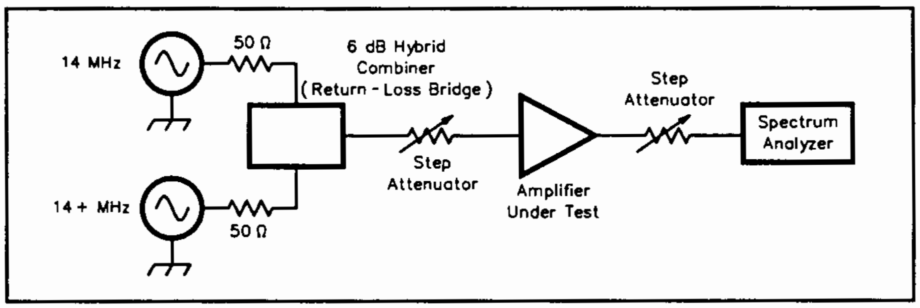


Figure 12—Equipment setup for evaluation of amplifier intermodulation distortion.

• Confirmed the lack of on-screen distortion with two tones at the reference level.

• Increased the drive of each tone by 10 dB to provide a pair of −20 dBm tones to the analyzer. This higher-than-reference-level input produced distortion products 66 dB below the reference level, or −96 dBm. The input signals producing this were each −20 dBm, so the IMD ratio is (−20) − (−96) = 76 dB. Following the earlier equation, the input intercept was +18 dBm.

• A 2-dB drive increase produced the expected 6-dB distortion increase. If this had not occurred, distortion measurements under overdrive would be suspect. The +18-dBm value seems to be a good number. This analyzer generally seems happy with signals 20 dB above the top of the screen, but not much more.

The intercept for the analyzer with attenuation in place is the measured value with no pad plus the attenuation. Hence, with 20 dB of attenuation, the input intercept will be +38 dBm, and so forth.

Return-Loss Measurements

The next amplifier characteristic that we measured was the input impedance match, or return loss, performed with the setup shown in Figure 13. With the signal generator set for a 14-MHz output of about −30 dBm, we set

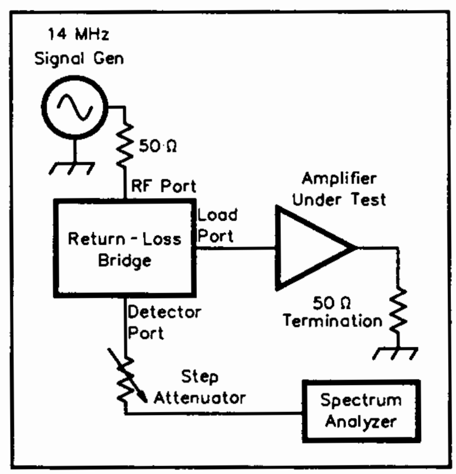


Figure 13—Test setup for impedance-match measurements with return-loss bridge.

the analyzer for full-scale response with the **LOAD** port of the return-loss bridge open circuited. Placing a 50-Ω termination momentarily on the **LOAD** port, produced a 38-dB signal drop. This is a measure of the bridge directivity. A 38-dB directivity is more than adequate for casual measurements.

Then, we removed the termination from

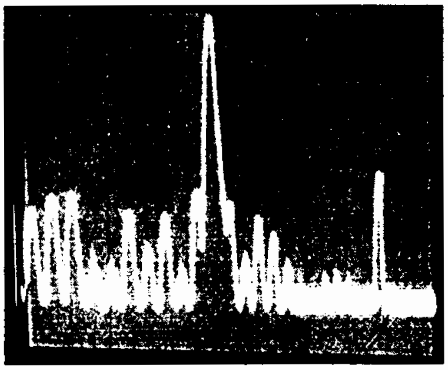


Figure 14—Output of a typical QRP transceiver kit. The 1-W plus output at 7 MHz is the dominant signal; all others are spurious outputs more than 40 dB down, currently meeting FCC specifications.

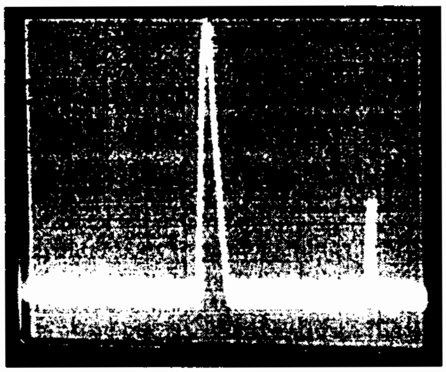


Figure 15—Output of a simple homemade QRP rig. The desired signal is the large pip at the center of the trace. Two measurable spurious responses exist, one to the left and one to the right of the main signal. The response to the left is 3.5 MHz feedthrough from the VFO at –64 dBc; the response to the right of the desired signal is a second harmonic at –44 dBc.

the bridge and placed it on the amplifier output. A short length of cable connected the bridge load port to the amplifier input with power applied to the amplifier. The result was a response 20 dB below the top of the display; 20 dB is the return loss for the amplifier input, an excellent match for a general-purpose amplifier.[20] The match improved slightly when the output load was removed, an unusual situation.

The next variation measured the output match for the amplifier. The load was transferred to the amplifier input and the cable from the bridge **LOAD** port was moved to the amplifier output, producing a reading 15 dB below the screen top. This match was virtually unchanged when the input termination was removed. The weak dependence of both amplifier return losses is the result of a four-stage design. A single-stage feedback amplifier will have port impedances that depend strongly on the termination at the opposite port.

The –30 dBm drive from the generator provides an available power of –36 dBm from the bridge **LOAD** port. This is low enough that the amplifier is not over-driven. The match measurements should be done at a level low enough that the amplifier remains linear. In this case, we saw no difference in the results with drive that was 10 dB higher.

When performing return-loss measurements—and indeed, most spectrum-analyzer measurements—it is wise to place at least 10 dB of attenuation ahead of the spectrum-analyzer mixer. When this attenuation is switched in, the reference level changes from the top of the screen to a point down screen a bit. Return loss is then measured as a decibel difference with regard to the new reference.

Antenna Measurements

It is interesting to look at some other impedance values while the return-loss bridge is attached to the signal generator and spectrum analyzer. The obvious choice is the station antenna system, especially if it is connected through a Transmatch. Playing with the tuning will readily demonstrate that the return-loss bridge and sensitive detection system will allow adjustments to accuracy unheard of with traditional diode detector systems. Although such tuning accuracy is not needed in a normal antenna installation, it is interesting to see what *can* be measured when the need does arise.

Transmitter Evaluation

Another obvious application for a spectrum analyzer is in transmitter evaluation. Figure 14 shows the output of a typical QRP transceiver kit. The 1-W plus output at 7 MHz is the dominant signal, with all others being transmitter spurious outputs. All spurs are more than 40 dB down, which meets current FCC specifications. On the other hand, significantly better performance is easily obtained, especially if the builder has the facilities to measure them. Figure 15 is a photograph of a simpler QRP rig with two measurable spurious responses. One is the 3.5-MHz feedthrough from the VFO at –64 dBc; the second is a harmonic at –44 dBc.

The output available from a typical QRP rig (and certainly higher power rigs) is enough to damage the spectrum-analyzer input circuits. Attenuators that we generally build are capable of handling 0.5 to 1 W input without damage, while commercial attenuators are rated at from 0.5 to 2 W input. The mixers used in this analyzer can be damaged with as little as 50 to 100 mW signals. Two methods can be employed to view the output of a high-power transmitter without causing damage to the spectrum analyzer. In one, the transmitter output is run through a directional coupler with weak coupling to the sampling port—perhaps –20 to –30 dB. The majority of the output is dissipated in a dummy load. The second method uses a fixed, high-power attenuator. Figure 16 shows an attenuator that will handle about 20 W while providing 20-dB attenuation. The design is not symmetrical.

Spectrum Analysis at Higher Frequencies

Although the 70-MHz spectrum analyzer is extremely useful, we constantly wish that it covered higher frequencies. Not only do we want to experiment on the VHF and UHF bands, but we need to examine higher-order harmonics of HF gear. One method we can use with a regular receiver is a converter, usually crystal controlled. The same can be done with a spectrum analyzer, although crystal control is not needed. We can build a simple block converter, consisting of nothing more than a 100 to 200 MHz VCO (just like that used in the analyzer) and a diode ring mixer. A Mini-Circuits POS-200 VCO with a 3 dB pad will directly drive a Mini-Circuits SBL-1 or TUF-1 mixer to produce a block converter with a nominal loss of 10 dB. (One of the spectrum analyzer front-end boards could be used, with slight modification, for the block converter.)

This block converter allows analysis of much of the VHF spectrum. With the converter VCO set at 100 MHz, frequencies from 100 to 170 MHz are easily studied. The 70 to 100-MHz image is also available—it can also lead to confusion, as a few minutes with a signal generator will demonstrate. With the converter VCO up at 200 MHz, the 200 to 270 MHz spectrum is also available. Clearly, there is nothing special about the particular VCO used in the converter. All that is required to convert other portions of the low UHF spectrum to the analyzer range is a different VCO, and perhaps, a higher-frequency mixer. We will soon build similar block converters to allow analysis of the 432-MHz area. One of the popular little UHF frequency counters is quite useful with these converters. The block converters should be well shielded and decoupled from the power supply.

Even without the analyzer, the block converter is a useful tool. For example, it can be used with a 10 MHz LC band-pass filter and an amplifier to directly drive a 50-Ω-terminated oscilloscope. This can serve as a sensitive detector for the alignment of a 110-MHz filter.

As useful as the block downconverter is, it has image-response problems that can greatly confuse the results. The preferred way to get to the higher frequencies is with a spectrum analyzer—just like the one we have described—but having higher-frequency oscillators and front-end filters. If you're careful, you can build helical resonator filters for the 500-MHz region, or higher, with sufficiently narrow selectivity to allow a second conversion to 10 MHz. A more practical route uses a third IF. Such a triple-conversion analyzer is shown in Figure 17, where a 1 to 1.8 GHz VCO moves signals in the 0 to 800-MHz spectrum to a 1 GHz IF. This signal is still easily amplified with available monolithic amplifiers. Building a 1 GHz filter should not be too difficult, for a narrow bandwidth is not required. A bandwidth of 20 or 30 MHz with a three-resonator filter would be adequate. The resulting signal is then heterodyned to a VHF IF (such as 110 MHz) where the remaining circuitry is now familiar.

Clearly, there are several ways to attack the project. Recent technology offers some help in the way of interesting band-pass filter structures, as well as high-performance, low-noise VCOs. Time spent with some cata-

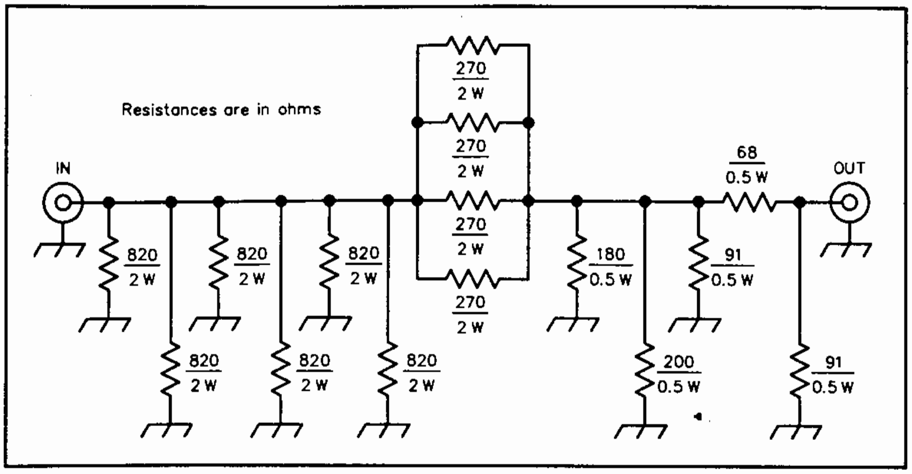
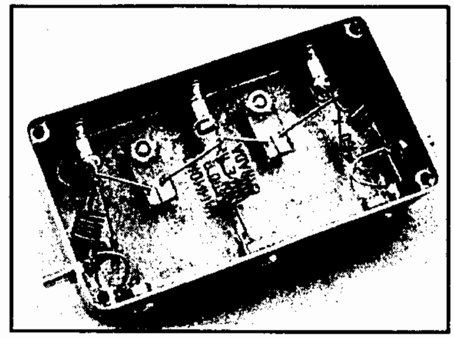


Figure 16—A 20-W, 20-dB attenuator for transmitter measurements.



An inside view of a prototype VHF band-pass filter. See Figure 7.

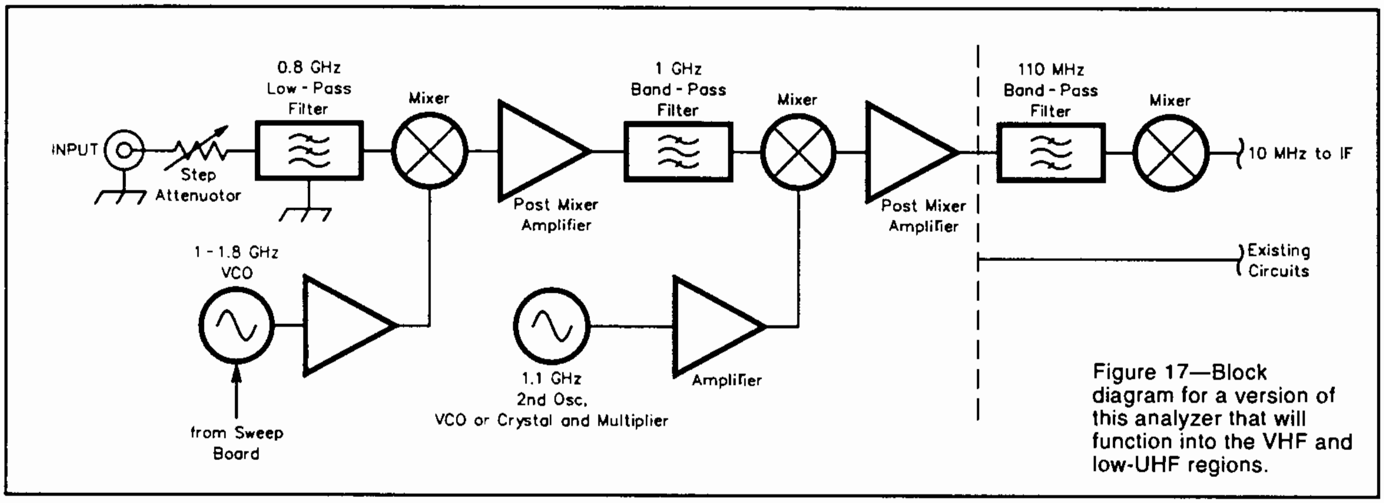


Figure 17—Block diagram for a version of this analyzer that will function into the VHF and low-UHF regions.

logs and a Web browser could be very productive in this regard. Just as important is the experience that the lower-frequency analyzer will provide. Not only is it the tool (perhaps supplemented with block converters) that is needed to build a higher-frequency spectrum analyzer, but it is the vehicle to provide *the confidence* needed to tackle such a chore.

Summary

This analyzer has been a useful tool for over 10 years now. It was a wonderful experience and fun to build HF and VHF CW and SSB gear with the "right" test equipment available. We have built special, narrow-tuning-range analyzers to examine transmitter sideband suppression and distortion. The equipment uses the same concepts presented.

There are many ways that this instrument can grow. One builder has already breadboarded a tracking generator. (Let's see that in *QST*!—*Ed.*) Many builders will want to interface the analyzer with a computer instead of an oscilloscope. A recent *QST* network analyzer paper suggests circuits that may provide such a solution.[21]

A recent *QST* summary of WRC97[22] outlines new specifications regarding spurious emissions from amateur transmitters. Generally, the casual specifications that we have enjoyed for many years are being replaced by new ones that are more stringent—and more realistic in safeguarding the spectral environment and reflecting the sound designs that we all strive to achieve. Equipment such as the spectrum analyzer described here can provide the basic tool needed to meet this new challenge.

Acknowledgments

Many experimenters had a hand in this project and we owe them our gratitude. Jeff Damm, WA7MLH, and Kurt Knoblock, WK7Q, built versions of the analyzer and have garnered several years of use with them. Their experiences have been of great value in our efforts. Barrie Gilbert of Analog Devices Northwest Labs suggested the AD8307 log amplifier and provided samples and early data needed for evaluation measurements. Many of our colleagues within the Wireless Communications Division at TriQuint Semiconductor have helped us with filter measurements: Thanks go to George Steen and to Don Knotts, W7HJS. Finally, special thanks go to our colleague in the receiver group at TriQuint, Rick Campbell, KK7B, who provided numerous enlightening discussions and suggestions regarding the preparation of the paper and the role of measurements in amateur experiments.

Notes

14 Wes Hayward, W7ZOI, and Terry White, K7TAU, "A Spectrum Analyzer for the Radio Amateur," *Part 1*, *QST*, Aug 1998, pp 35-43.

15 The wide-range oscillator presented in Fig 68, Chapter 7 of *Solid-State Design for the Radio Amateur* (Newington: ARRL, 1997) is still intact and still often used.

16 The term dBc refers to dB attenuation with respect to a specific *carrier*.

17 Formally, this is the transducer gain, or 50-Ω insertion gain. There are many different parameters that are called "gain."

18 The signal sources used are updated versions of the circuits shown in Fig 66, p 168, in the Note 15 referent.

19 See page 154, *Solid-State Design for the Radio Amateur*.

20 A 20-dB return loss corresponds to a voltage reflection coefficient of 0.1, or an SWR of 1.222. See Wes Hayward, W7ZOI, *Introduction to RF Design* (Newington: ARRL, 1994), p 120.

21 See, for example, Steven Hageman, "Build Your Own Network Analyzer," *QST*, Jan 1998, pp 39-45; *Part 2*, Feb 1998, pp 35-39.

22 Larry Price, W4RA, and Paul Rinaldo, W4RI, "WRC97, An Amateur Radio Perspective," *QST*, Feb 1998, pp 31-34. See also Rick Campbell, KK7B, "Unwanted Emissions Comments," Technical Correspondence, *QST*, Jun 1998, pp 61-62.

Extending the Double-Tuned Circuit to Three Resonators

Here's an empirical design method for triple-tuned filters.

By Wes Hayward, W7ZOI

Although the most common LC band-pass filters used by radio amateurs are single and double-tuned circuits, there are situations where a higher-order filter is desired. The triple-tuned circuit, a band-pass filter with three resonators, is an especially useful example. It is easily designed and offers improved stop-band attenuation with little increase in insertion loss. This paper presents two simple methods that allow the VHF triple-tuned circuit to be adjusted and tuned with simple instruments and little more complication that a double-tuned filter.

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The Double-Tuned Circuit as a Step toward the Triple-Resonator Filter

Two filter schematics are shown in Fig 1. Both were designed for a Butterworth response and a center frequency of 110 MHz with a 3-dB bandwidth of 2 MHz. The N = 3 (three pole) filter has slightly higher insertion loss than the double-tuned circuit, but is more selective at high attenuation. The reader can calculate the responses with any of dozens of available computer programs.[1]

It is interesting to compare the two schematics. The coupling capacitors are identical for the same bandwidth. However, the end-section matching is slightly different, with the end Q being higher for the double-tuned circuit (DTC) than for the triple-tuned circuit (TTC). This leads us to ask what the performance would be for a DTC that has the same end designs as a TTC. This is shown schematically in Fig 2 with responses shown in Fig 3.

[1]Notes appear on page 46.

Experimental Methods

The introductory example presented above used computer generated and analyzed designs. This is all the preparation that is needed for lower-frequency filters where discrete reactances are large. This works well for filters up through 30 to 50 MHz with capacitor values of 1 pF or more and inductors of a couple hundred nanohenries or more. As we move into the VHF area, and higher, the components are not described well by "printed" values. Stray reactances begin to dominate, encouraging us to use experimental methods.

I described the DTC in detail in a previous *QST* paper.[2] Experimental methods were emphasized in that tutorial, which described a method for building a double-tuned circuit without ever going through the numbers. The method entails adjusting the end loading on both resonators while also adjusting coupling. The resonators are tuned for maximum response after K and Q adjustment; the insertion loss is then measured. The reader is urged to study that paper if he or she is not very familiar with the procedure.

Perhaps the most important detail presented in the DTC tutorial was the need to perform a wide-band sweep to locate any double-humped response that might be present. A common and potentially disastrous error that the experimenter can make with the DTC is severe over coupling that produces two widely separated response peaks. It is easy to miss one of the two peaks if a wide sweep is not done.

With this background in mind, we can use the earlier observations to implement a triple-tuned circuit. The three-element filter is built with the middle resonator eliminated. The loading is kept identical at the two ends and is adjusted for a desired filter bandwidth. Coupling is set up between the first and *third* resonator and the combination is adjusted as a DTC. A wide-band sweep is done to find the double humped response, if present. It is then eliminated through further adjustment, if found. This is

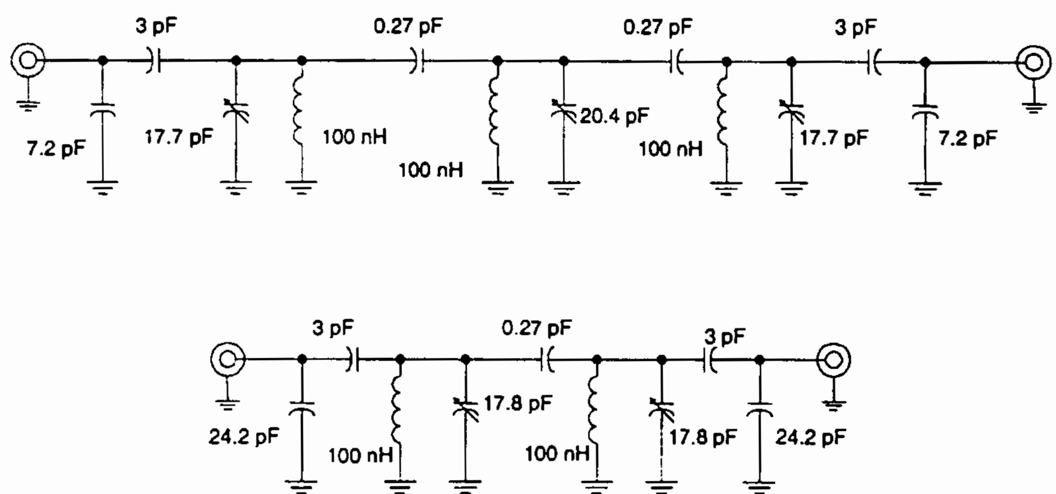


Fig 1—Butterworth LC Band-pass Filters with a 2 MHz bandwidth, centered at 110 MHz. The upper schematic is a triple-tuned circuit (N = 3), while the lower schematic is a double-tuned circuit.

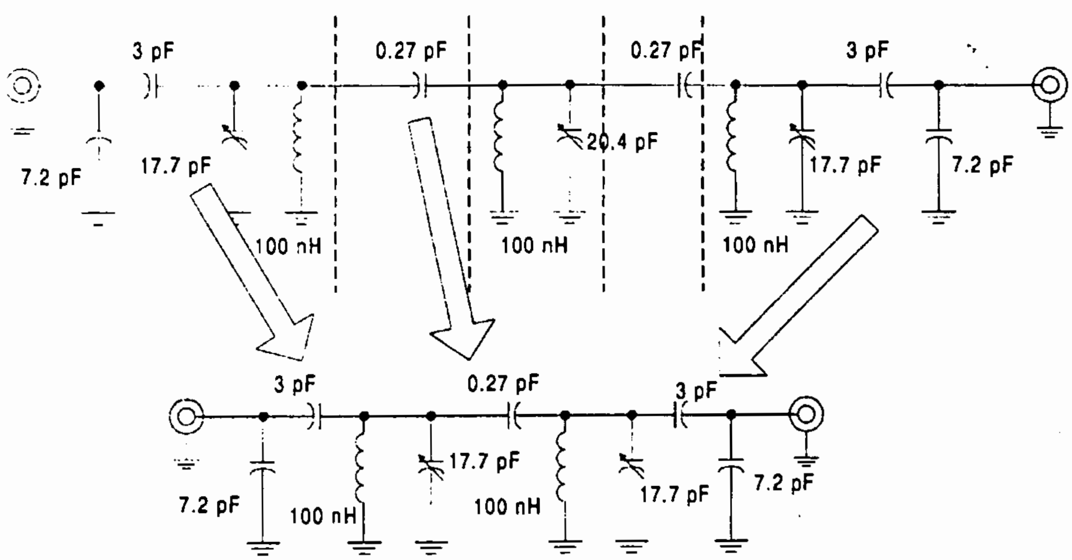


Fig 2—A TTC is designed "on paper" and is mentally segmented into end resonators, a middle resonator, and coupling elements. A DTC is then fabricated consisting of the two end resonators and a coupling element. The response of the DTC, and the parent TTC are shown in Fig 3.

repeated until the desired bandwidth is achieved in a DTC.

With a working DTC now in hand, the coupling is examined and duplicated as the third, middle resonator is added. The coupling from resonator 1 to 2 should be the same as that from 2 to 3. The TTC is then tuned and measured, performing a wide-band sweep to confirm the absence of extra peaks. The final bandwidth should be close to that of the intermediate DTC with a slightly higher insertion loss.

The filter shown in Photos A and B was built for use as the first IF of a spectrum analyzer. Filter design began with selection of an inductor. Values around 100 nH are practical at 110 MHz. The inductor was built with a 6.0 inch piece of #18 enameled wire. The ends were stripped and five turns of the wire were wound on the shank of a 1/4 inch drill bit.[3] The unloaded Q was assumed to be around 200, a value later confirmed with measurements. This inductor resonates at 110 MHz with a capacitor of about 21 pF, realized in the filter with combinations of fixed and glass trimmers. The predicted coupling capacitor value is 0.27 pF for each of the two components. A computer design is useful to provide guidance during construction, even if some components are less than practical.

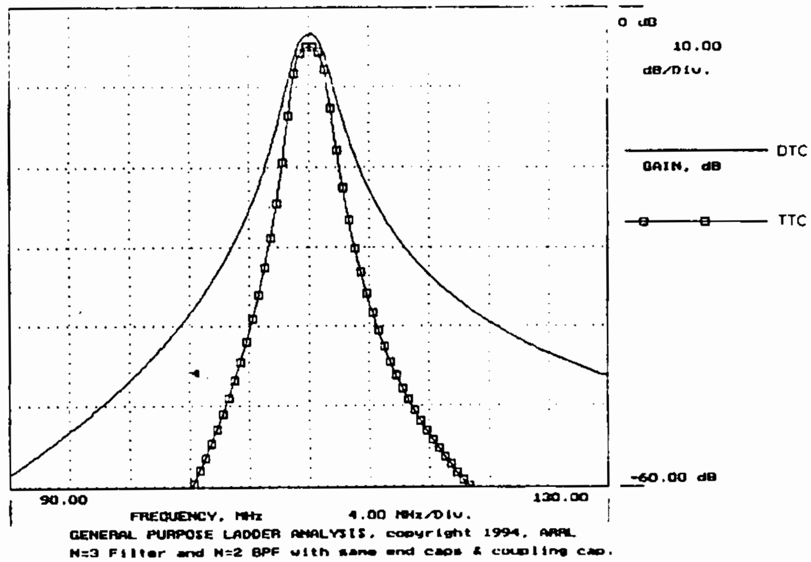


Fig 3—Response of a TTC and a DTC derived from it. Both have approximately the same bandwidth.

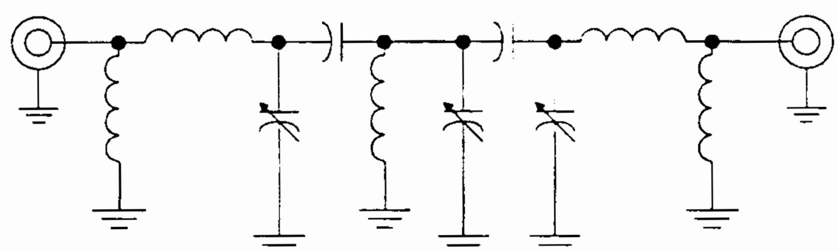


Fig 4—Filter form used with a band-pass circuit shown in photograph A.

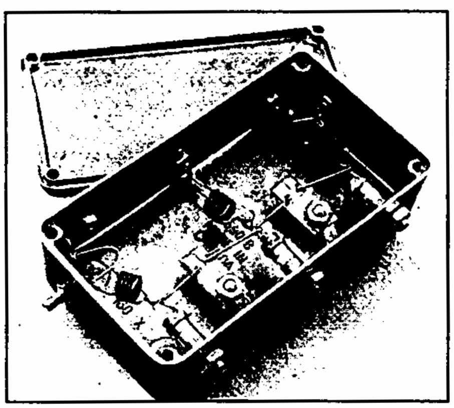


Photo A—A TTC filter for 110 MHz.

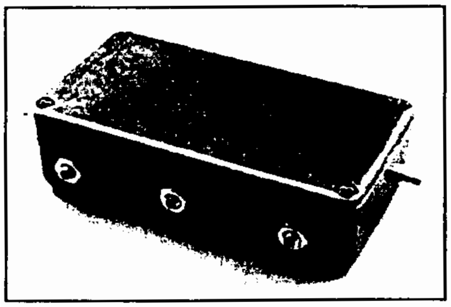


Photo B—An outside view of the TTC in its cast box.

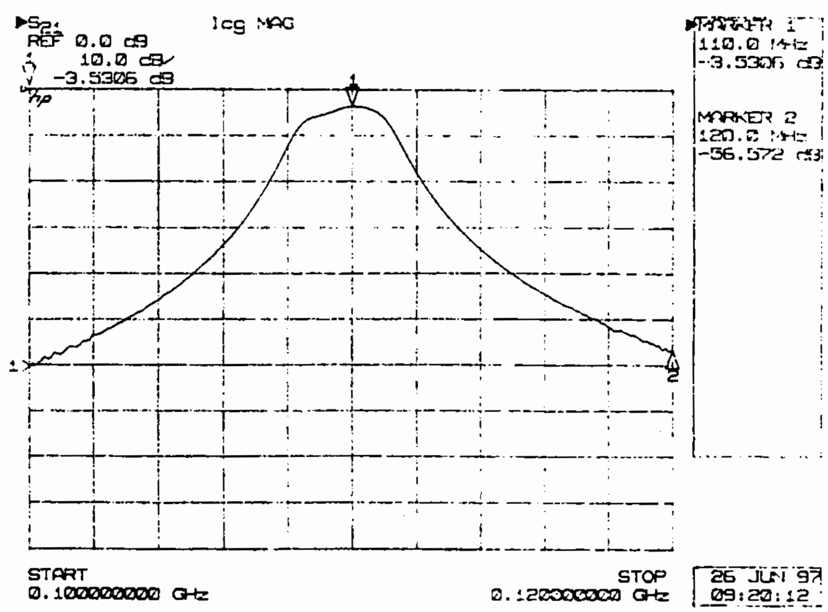


Fig 5—Experimental results with an example triple-tuned filter.

The generality of these methods allows great flexibility. Any of several coupling methods can be used for end-section loading. Even though the schematic shows capacitive taps for attachment between end resonators and the 50 Ω loads, other methods could just as well be applied. For example, the filter shown in the photos used end loading realized with the circuit of Fig 4, where the grounded end of the inductor is lifted from ground and attached to a coaxial connector. Then, a small inductor is attached from the connector to ground. The size of this inductor is varied to establish the end Q. Even though the physical details are different, the measurement schemes and results are not.

Figure 5 shows experimental results, a plot for the experimental filter. The evaluation measurements were performed with an HP-8510B network analyzer. However, the sophisticated instrument was not used for any of the adjustments.

Filter Adjustment with a Return-Loss Bridge

A second useful method is available to the experimenter with a return-loss bridge in his or her laboratory. This instrument, which has often been described in the literature[4], is an impedance bridge where the error, or unbalance signal serves as a measure of reflection coefficient (or SWR) looking into a load. Fig 6 is the schematic for a filter where the input reflection coefficient is to be measured. This analysis was performed with the evaluation version of the popular *PSpice*.[5] The use of a pair of voltage sources at the input generates a voltage, shown as "gam," which *is* the voltage reflection coefficient.[6] Return loss relates to Gamma through

$$Return\ Loss = -20 \cdot Log(\Gamma) \qquad \text{(Eq 1)}$$

Figure 7 shows the gain and magnitude of S11 versus frequency for the complete filter. The match plot dips down to a return loss of 20 dB at the filter center frequency.

Figure 8 shows a modification where the match is calculated for an end resonator that is no longer coupled to the rest of the filter. The result is shown in Fig 9. This match, now only 6.5 dB at resonance, is not nearly as good as the complete filter. If the unloaded resonator Q was much higher, the return loss would be even less. Hence, it is important that the unloaded Q value be reasonably accurate during this simulation.

We eventually wish to use this as an aid to measurements. This was to be a measure of impedance at resonance, but we have not obtained enough information to establish an impedance. Two different resistive loads can provide a 6 dB return loss, just as two different pure resistances can provide a 3:1 SWR. This dilemma is solved with the computer simulation shown in Fig 10, where the single resonator is swept several times with different values of matching inductor with each sweep. A return loss of 6 dB occurs with matching inductor values of 4 nH and 10 nH. Yet only the higher, 10 nH, value produces the desired filter response.

The information from the simulations can now be used to tune our circuit. The TTC filter is built, but the center resonator is either short circuited, or removed. One end is driven at 110 MHz by a return-loss bridge and the resonator is tuned for best return loss, occurring with the largest dip. The matching (grounded) inductor is then varied and retuned until a return loss of about 6 dB is measured. The matching inductor should have a total wire length of about an inch, corresponding to an inductance of about 10 nH using the rule-of-thumb that a

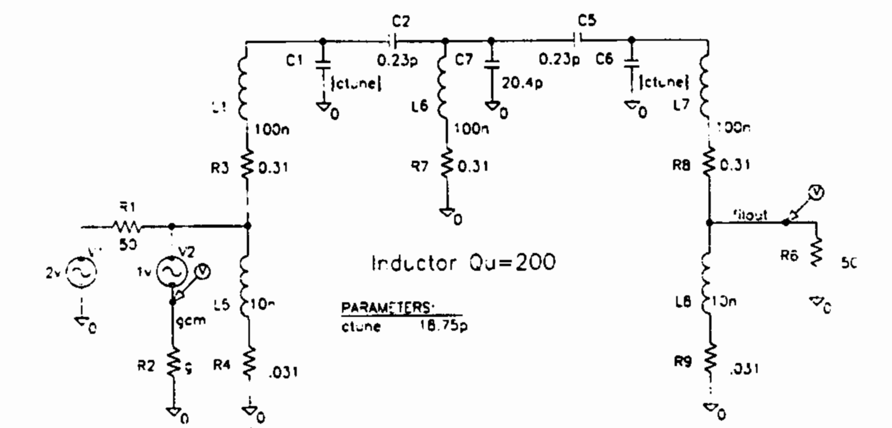


Fig 6—Schematic for practical version of the filter, set up for analysis in *PSpice*.

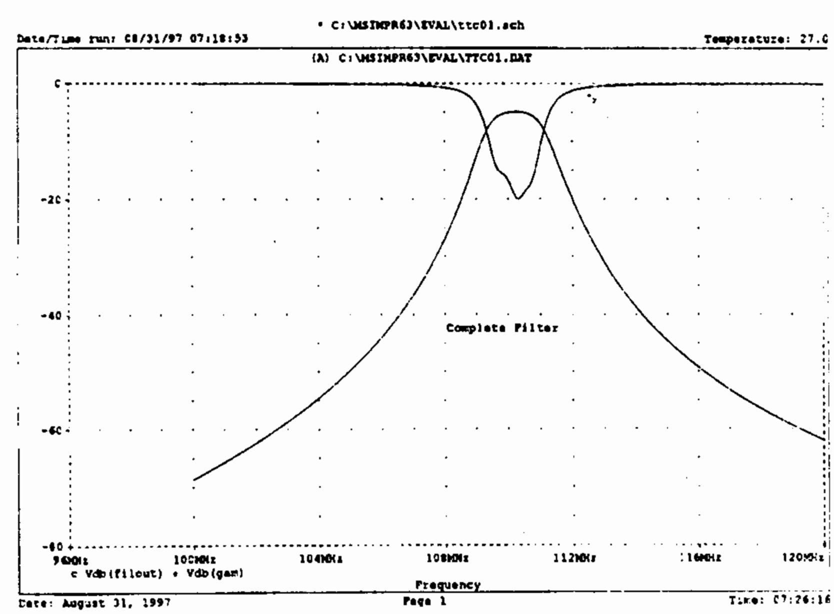


Fig 7—Transfer function and impedance match for complete filter. The return loss looking into the filter is about 20 dB at band center.

wire has L = 0.5 nH per millimeter of wire length. This procedure is repeated for the other end of the filter.

Having established the end Q, the center resonator is added to the circuit. Equal valued coupling capacitors are added between resonators, always keeping the values as small as "seems" reasonable. The three resonators are tuned at 110 MHz as the coupling caps are adjusted. A wide-band sweep is performed during the process to guarantee that overcoupling is not producing extra response peaks. If extra peaks appear, the coupling capacitors are reduced until the desired band-pass shape and bandwidth are obtained.

Some Practical Considerations

Band-pass filters are critical elements in most RF systems and should be built with care. Shielding is often needed; not only is it important to shield the resonators from the outside

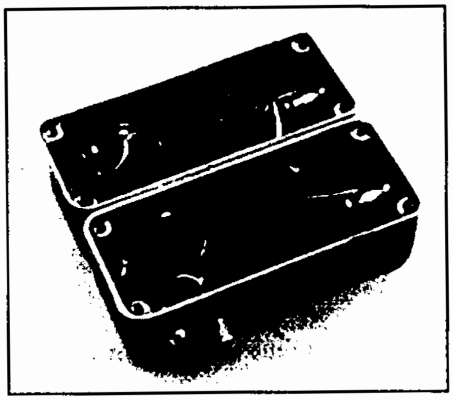


Photo C—This DTC is mounted in *two* boxes to improve isolation.

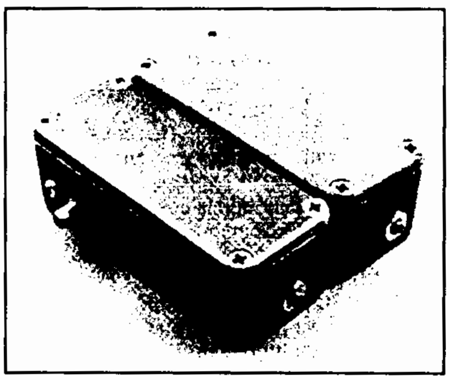


Photo D—An outside view of the DTC in two boxes.

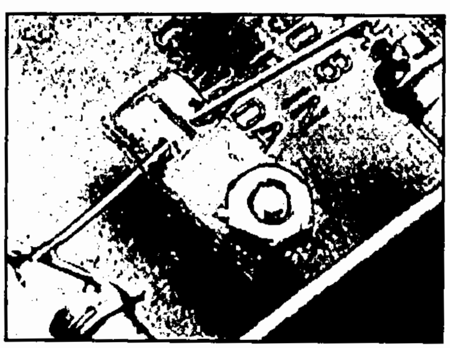


Photo E—This enlargement from Photo A shows one method to achieve tiny coupling capacitances. This method of capacitively coupling two conductors is sometimes called a "gimmick" capacitor. See text.

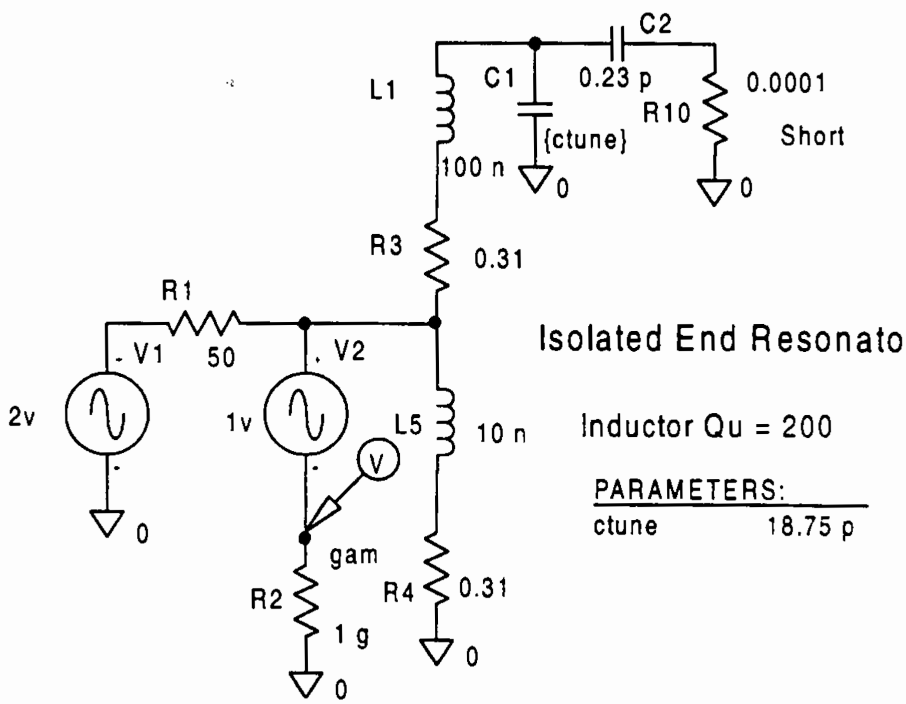


Fig 8—Calculation of reflection when looking into an end resonator without other coupled resonators.

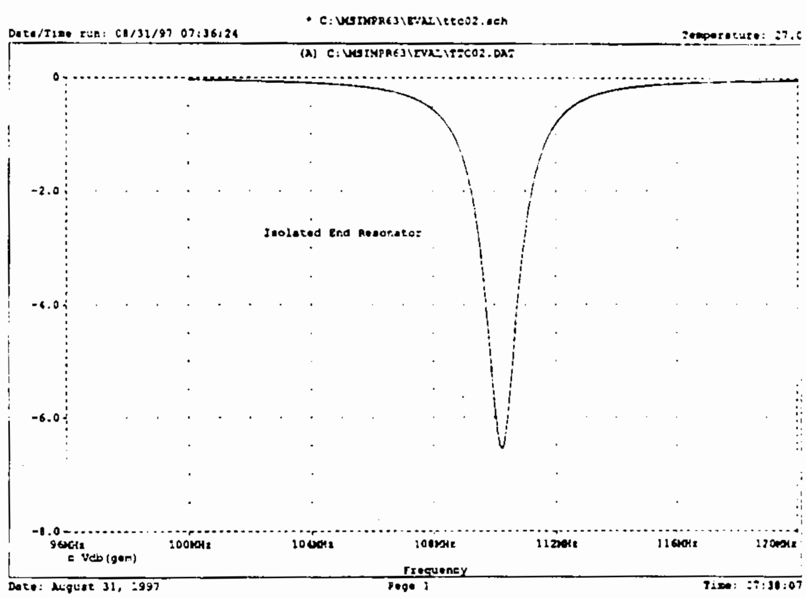


Fig 9—Reflection looking into the isolated resonator.

world, but shielding between resonators is often required. The 110 MHz example TTC is built in a single cast-aluminum box. (See Photos A and B.) Small-diameter coils were picked specifically to minimize interaction between resonators, eliminating the need for internal shields. Photos C and D show a double-tuned circuit built in two cast-aluminum boxes that have been bolted together. Large, higher Q inductors are used. This scheme can be expanded to numerous filter elements. It is often worthwhile to flip adjacent boxes so that lids alternate, side to side to accommodate cast boxes with nonparallel sides.

The small valued coupling capacitors needed in band-pass LC filters are often difficult to realize. One scheme that I have applied is sometimes called a gimmick capacitor (shown in Photo E). Two small, isolated pads are fabricated on a scrap of single sided circuit board material. (Double-sided board has excessive capacitance related to the board material.) Wires are then run from the "hot" ends of the resonators to the pads. The wires are kept a bit longer than required to reach the pads. The excess ends can then be bent close to each other, as needed, to adjust the capacitance. In the example filter, the excess wire lengths were completely trimmed away, for the stray pad-to-pad capacitance provided the needed coupling.

The measurements outlined are best done with the best instrumentation available. This may be elegant commercial gear, or simple home built tools. It is quite possible to build and adjust a TTC with a signal generator and a home-brew power meter, such as the unit recently described by K7OWJ.[7] Once the filter is finished, it can be integrated into a spectrum analyzer, an instrument that will then simplify the adjustments the next time a TTC is needed.

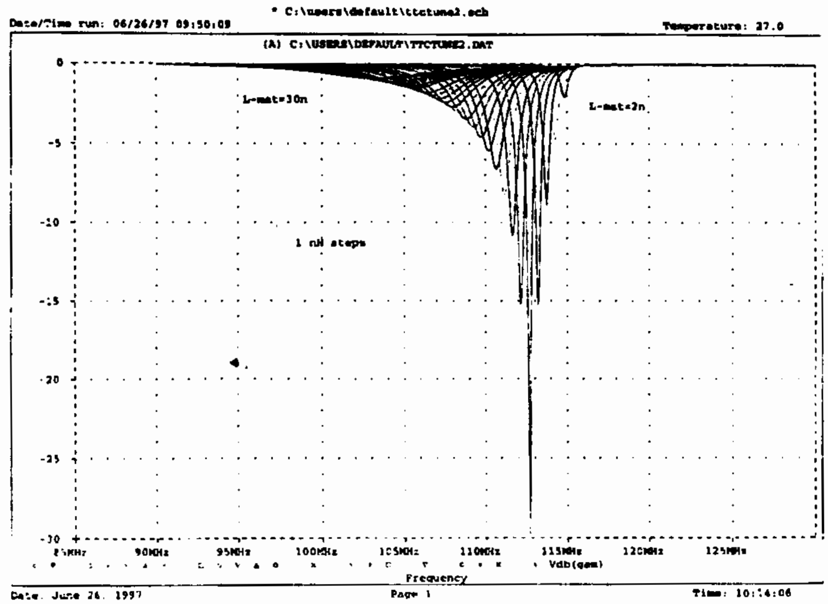


Fig 10—Reflection looking into an isolated resonator with the matching inductor changed in 1 nH steps.

References:

[1]The filters were designed with *DTTC*, a program offered with the ARRL version of "*Introduction to Radio Frequency Design*," ARRL Order No. 4920. The analysis of Fig 3 was generated with *GPLA*, also offered with the text. *ARRL Radio Designer* is also suitable for analysis of filters of this sort. *PSpice* is a program offered by MicroSim, 20 Fairbanks, Irvine, CA 92718. Also, see **www.microsim.com**. The evaluation version of *PSpice* is a very effective, yet affordable tool. ARRL publications are available from your local ARRL dealer or directly from ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the World Wide Web at **http://www.arrl.org/catalog**.

[2]Hayward, "The Double-Tuned Circuit," *QST*, Dec. 91, pp 29-34.

[3]See IRFD (See Note 1). The coils are designed with a program on the disk, *COILS.EXE*.

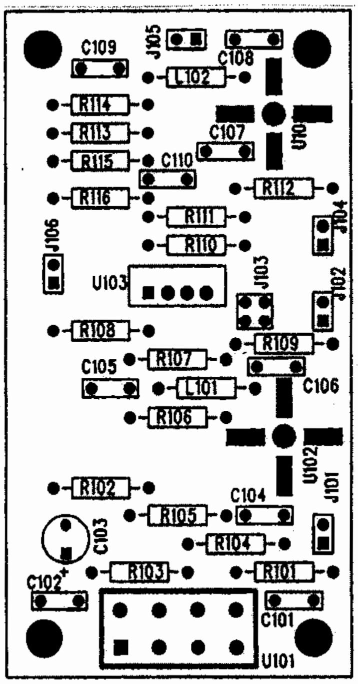
[4]The return-loss bridge is described in IRFD, in *Solid-State Design for the Radio Amateur*, (Newington: ARRL, 1977; ARRL Order No. 0402) and in several editions of the *ARRL Handbook (ARRL Order No. 1786)*, as well as numerous articles.

[5]See information on MicroSim presented in Note 1.

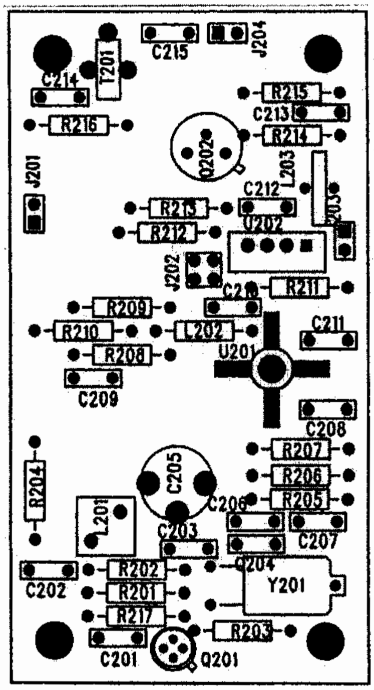
[6]Hayward, "Reflections on the Reflection Coefficient," *QEX*, Jan 1993, p 10-12.

[7]Bramwell, "The Microwatter," *QST*, June 1997, pp 33-35.

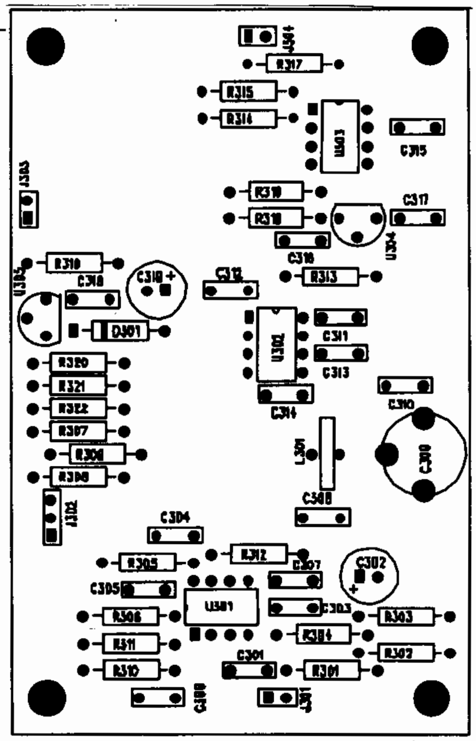
PC Board Layouts



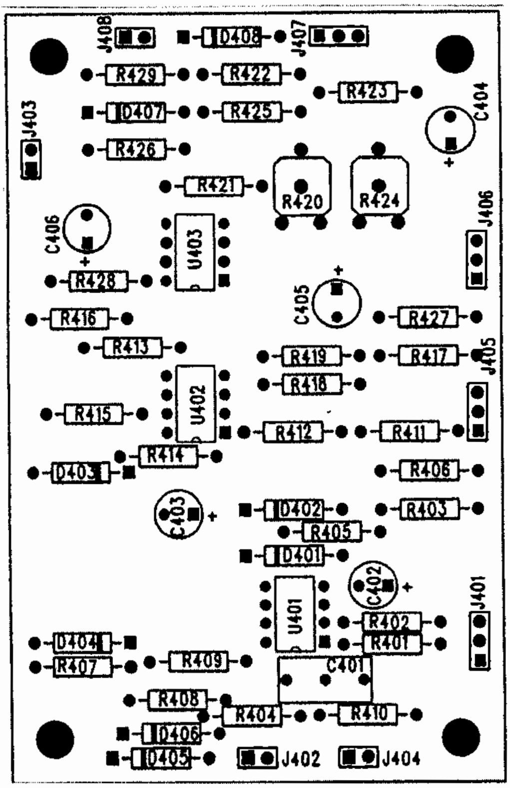
1st LO & 1st Mixer



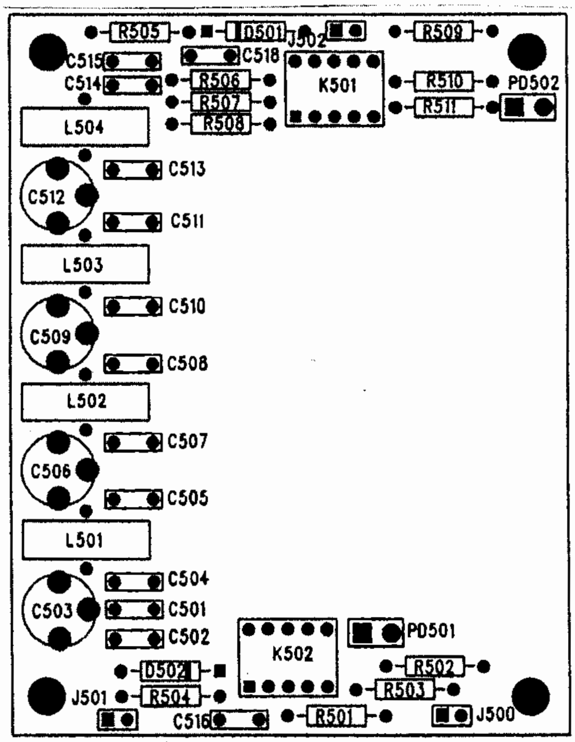
2nd LO & 2nd Mixer



IF & Log Amp



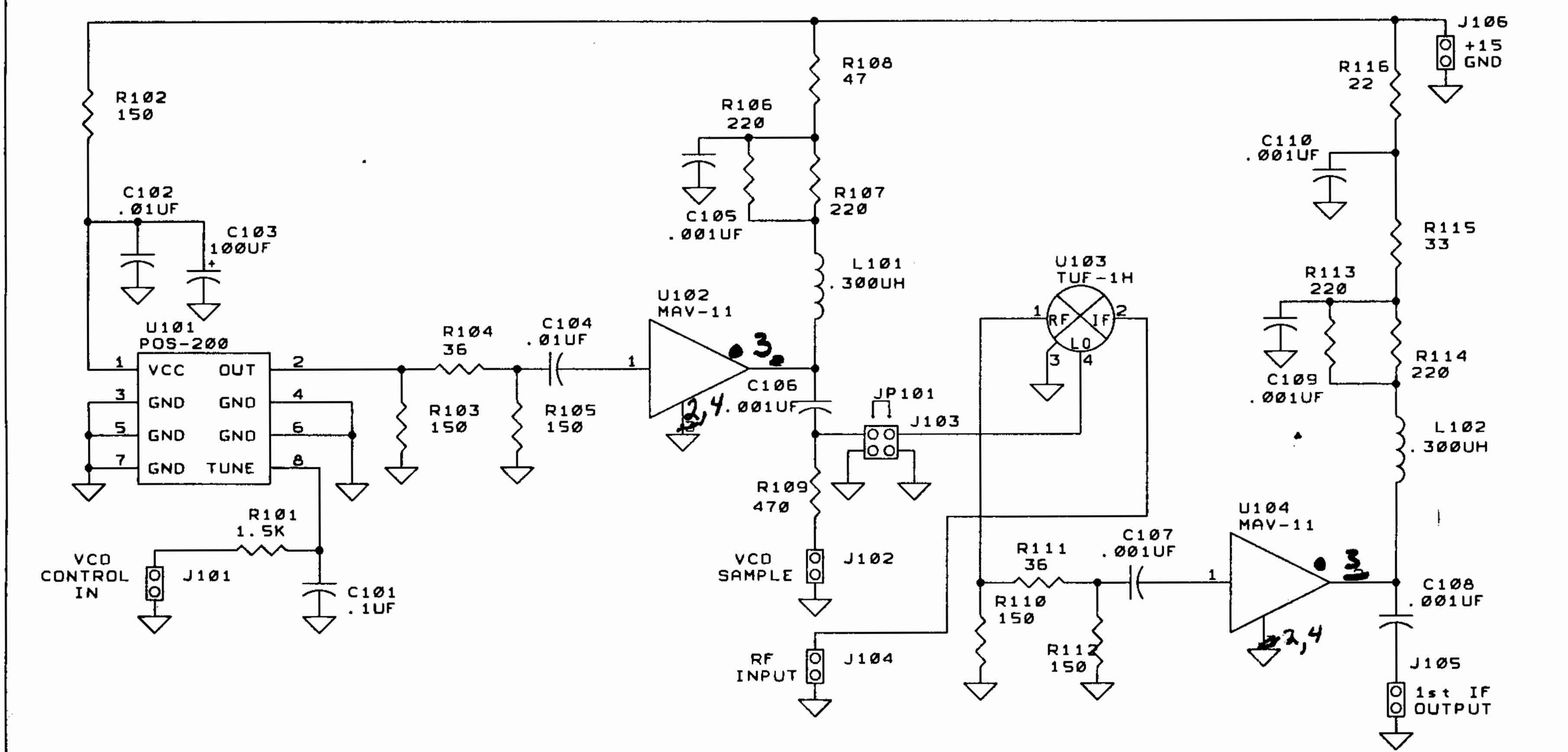
Time Base Generator



Resolution Bandwidth Filter

Parts List – 1st LO and 1st Mixer

Qty	Reference	Part Description
1	C101	.1uf, 50v
2	C102, C104	.01uf, 50v
1	C103	100uf, 25v
6	C105, C106, C107, C108, C109, C110	.001uf, 50v
2	L101, L102	.3 uh, T37-6, 10T, #30
1	R101	1.5k, 1/4w, 5%
5	R102, R103, R105, R110, R112	150, 1/4w, 5%
2	R104, R111	36, 1/4w, 5%
4	R106, R107, R113, R114	220, 1/4w, 5%
1	R108	47, 1/4w, 5%
1	R109	470, 1/4w, 5%
1	R115	33, 1/4w, 5%
1	R116	22, 1/4w, 5%
1	U101	POS-200 VCO
2	U102, U104	MAV-11
1	U103	TUF-1H



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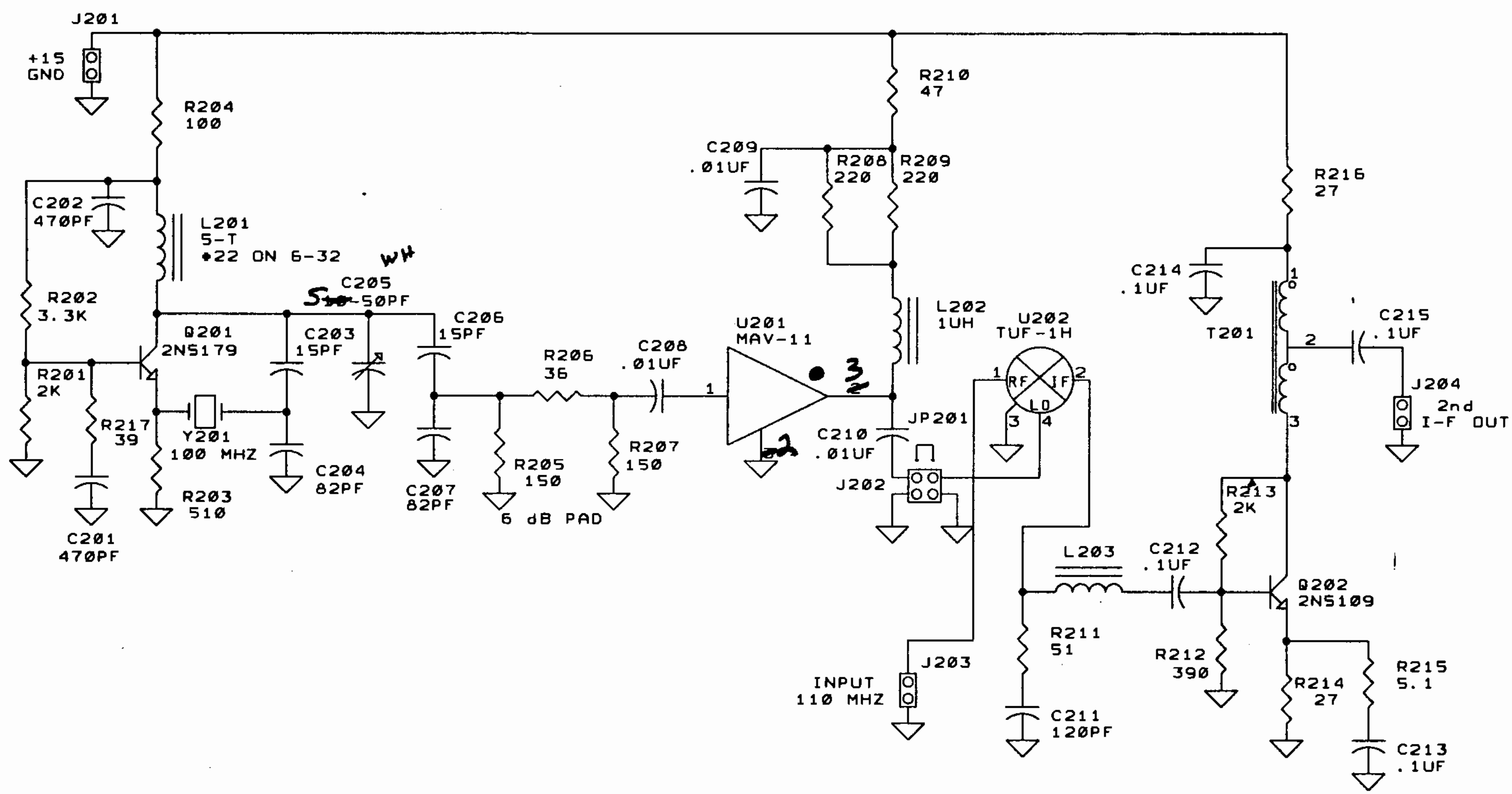
Date: March 12, 1998	Sheet 1 of 1	

Parts List - 2^{nd} LO and 2^{nd} Mixer

Qty	Reference	Part Description
2	C201, C202	470pf
2	C203, C206	15pf NP0
2	C204, C207	82pf NP0
1	C205	5 –50 pf trimmer
3	C208, C209, C210	.01uf
1	C211	120pf NP0
4	C212, C213, C214, C215	.1uf
1	L201	5T #22 on 6-32 thread
2	L202, L203	1uh, T37-6, 10T, #30
1	Q201	2N5179
1	Q202	2N5109
2	R201, R213	2k, 1/4w, 5%
1	R202	3.3k, 1/4w, 5%
1	R203	510, 1/4w, 5%
1	R204	100, 1/4w, 5%
2	R205, R207	150, 1/4w, 5%
1	R206	36, 1/4w, 5%
2	R208, R209	220, 1/4w, 5%
1	R210	47, 1/4w, 5%
1	R211	51, 1/4w, 5%
1	R212	390, 1/4w, 5%
2	R214, R216	27, 1/4w, 5%
1	R215	5.1, 1/4w, 5%
1	R217	39, 1/4w, 5%
1	T201	FT37-43, 10T Bifilar, #30
1	U201	MAV-11
1	U202	TUF-1H
1	Y201	100MHz, 5^{th} overtone



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2nd LO and 2nd MIXER		
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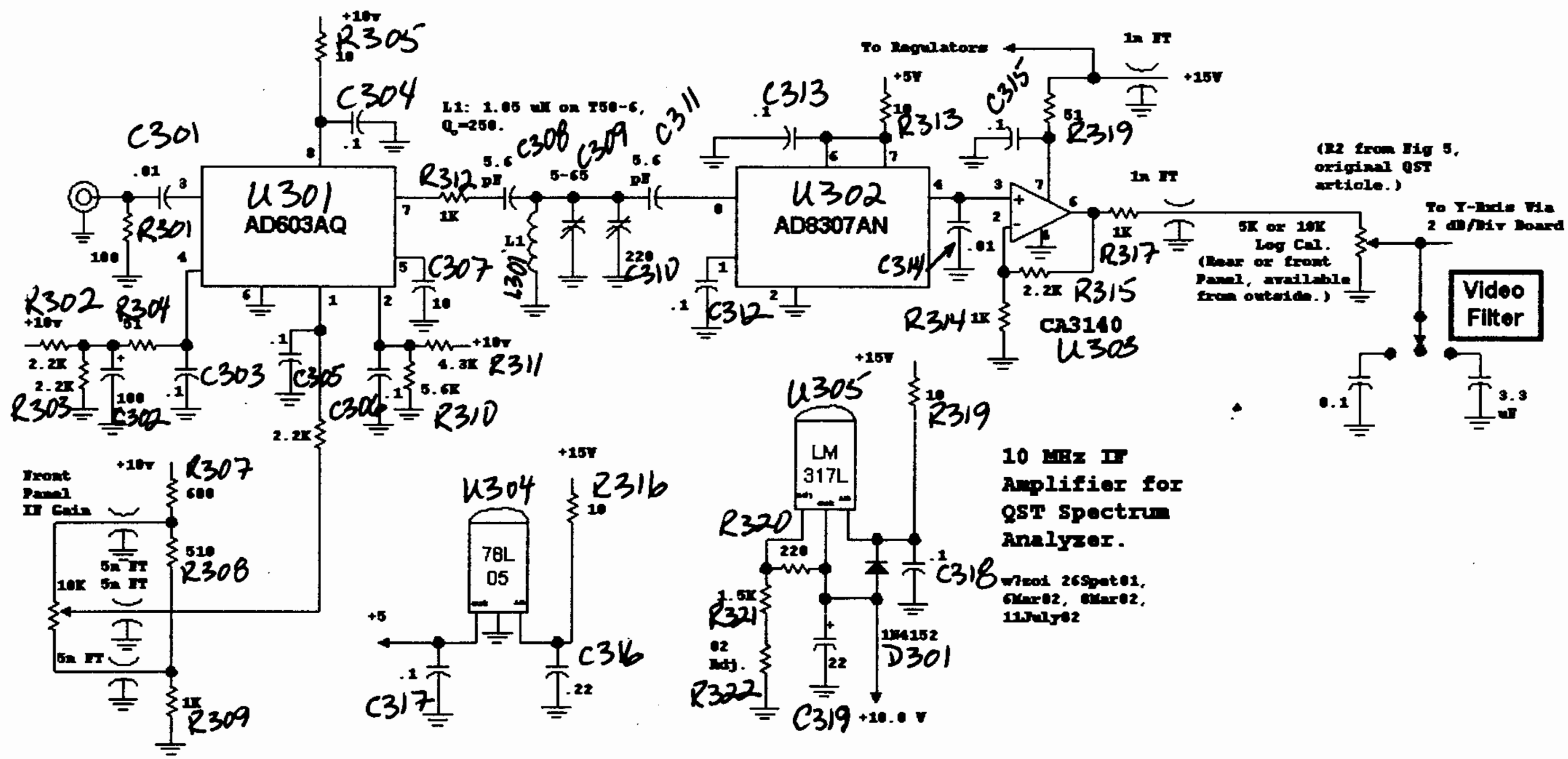
Date: June 22, 1998	Sheet 1 of 1	

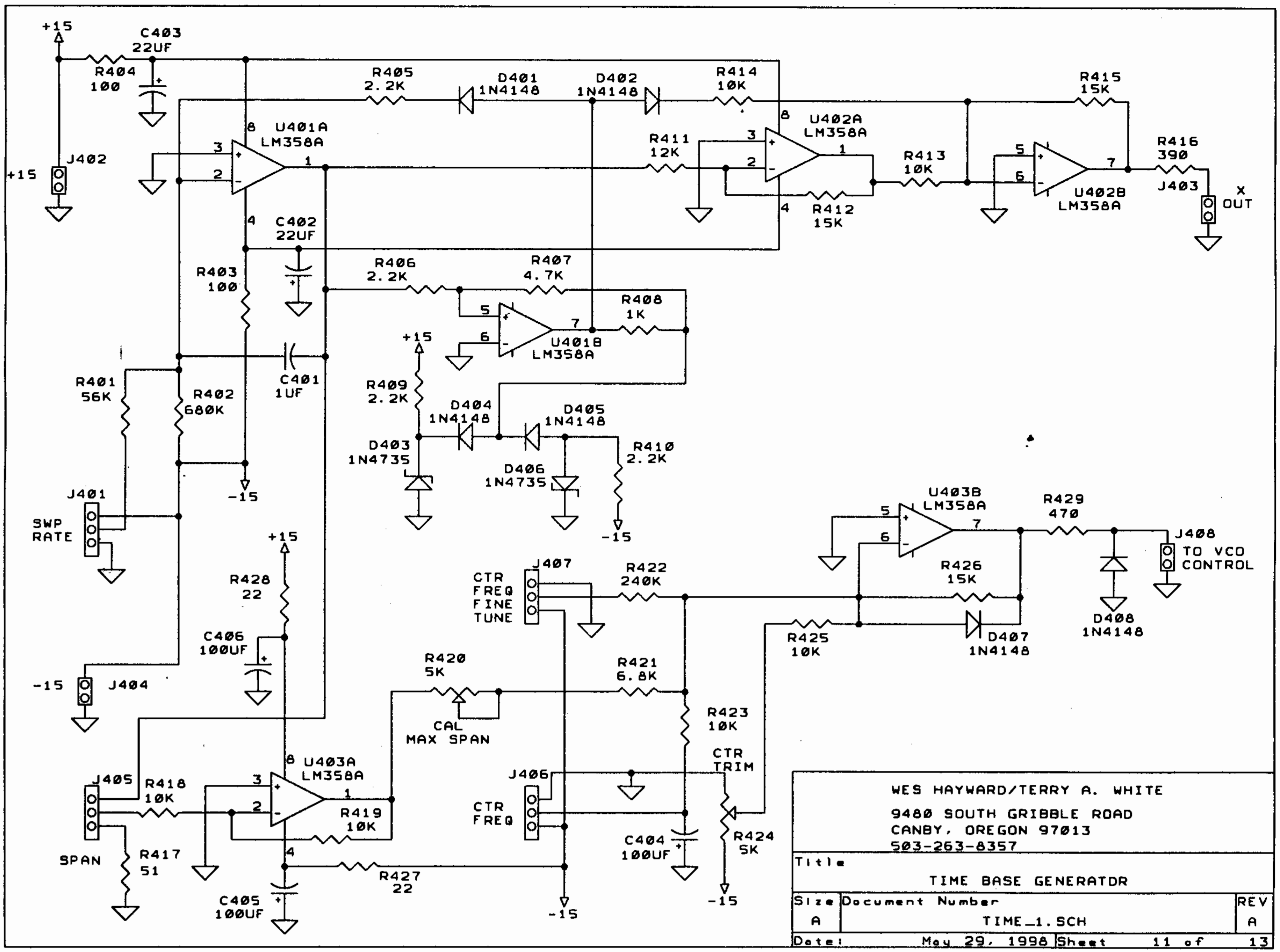
SA - New Log IF Amp		
Qty	Reference	Part Description
2	C301, C314	.01uf
1	C302	100uf
9	C303, C305, C306, C312, C313, C315, C317, C318, offboard	.1uf
1	C307	18pf
1	C309	65pf trimmer
2	C308, C311	5.6pf
1	C310	220pf
1	C316	.22uf
1	C319	22uf
1	offboard	3.3uf
1	R301	100 ohm
4	R302, R303, R306, R315	2.2k ohm
2	R304, R319	51 ohm
4	R304, R313, R316, R319	10 ohm
1	R307	680 ohm
1	R308	510 ohm
4	R309, R312, R314, R317	1k ohm
1	R310	5.6k ohm
1	R311	4.3k ohm
1	R320	220 ohm
1	R322	82 ohm
2	offboard	5k or 10k pot
1	L301	T50-6, 1.05uh 16T #22
1	U301	AD603AQ
1	U302	AD8307AN
1	U303	CA3140
1	U304	78L05
1	U305	LM317L
1	D301	1N4148 or equiv
5	FT	Feedthru cap .01 - .05uf
1	offboard	spdt switch
1		PC Board



U301
AD603AQ
U302
AD8307AN
CA3140
U303
U304
78L
05
U305
LM
317L
L1: 1.05 uH on T50-6, Q=250.
To Regulators
+15V
+5V
+10v
1n FT
(R2 from Fig 5, original QST article.)
To Y-Axis Via 2 dB/Div Board
5K or 10K Log Cal. (Rear or front Panel, available from outside.)
Video Filter
Front Panel IF Gain
1N4152
D301
+10.0 V
+5
02 Adj.
10 MHz IF Amplifier for QST Spectrum Analyzer.
w7zoi 26Spet01, 6Mar02, 8Mar02, 11July02

Parts List – Time Base Generator

Qty	Reference	Part Description
1	C401	1uf, Poly
2	C402, C403	22uf, 25v
3	C404, C405, C406	100uf, 25v
6	D401, D402, D404, D405, D407, D408	1N4148 or equiv.
2	D403, D406	1N4735, 6.2v, Zener, 1W
1	R401	56k, 1/4w, 5%
1	R402	680k, 1/4w, 5%
2	R403, R404	100, 1/4w, 5%
4	R405, R406, R409, R410	2.2k, 1/4w, 5%
1	R407	4.7k, 1/4w, 5%
1	R408	1k, 1/4w, 5%
1	R411	12k, 1/4w, 5%
3	R412, R415, R426	15k, 1/4w, 5%
6	R413, R414, R418, R419, R423, R425	10k, 1/4w, 5%
1	R416	390, 1/4w, 5%
1	R417	51, 1/4w, 5%
2	R424, R420	5k pot linear, board mounted
1	R421	6.8k, 1/4w, 5%
1	R422	240k, 1/4w, 5%
2	R428, R427	22, 1/4w, 5%
1	R429	470, 1/4w, 5%
3	U401, U402, U403	LM358A


+15
C403
22UF
R404
100
J402
R405
2.2K
D401
1N4148
D402
1N4148
R414
10K
R415
15K
U401A
LM358A
R411
12K
U402A
LM358A
R413
10K
R412
15K
U402B
LM358A
R416
390
J403
X
OUT
C402
22UF
R403
100
R406
2.2K
R407
4.7K
R408
1K
U401B
LM358A
C401
1UF
R401
56K
R402
680K
R409
2.2K
D404
1N4148
D405
1N4148
D403
1N4735
D406
1N4735
R410
2.2K
-15
J401
SWP
RATE
U403B
LM358A
R429
470
J408
TO VCO
CONTROL
R428
22
C406
100UF
CTR
FREQ
FINE
TUNE
J407
R422
240K
R426
15K
D408
1N4148
R425
10K
D407
1N4148
J404
R420
5K
CAL
MAX SPAN
R421
6.8K
R423
10K
CTR
TRIM
J405
R418
10K
U403A
LM358A
R419
10K
J406
CTR
FREQ
C404
100UF
R424
5K
SPAN
R417
51
R427
22
C405
100UF
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May 29, 1998
Sheet 11 of 13

Parts List – Resolution Bandwidth Filter

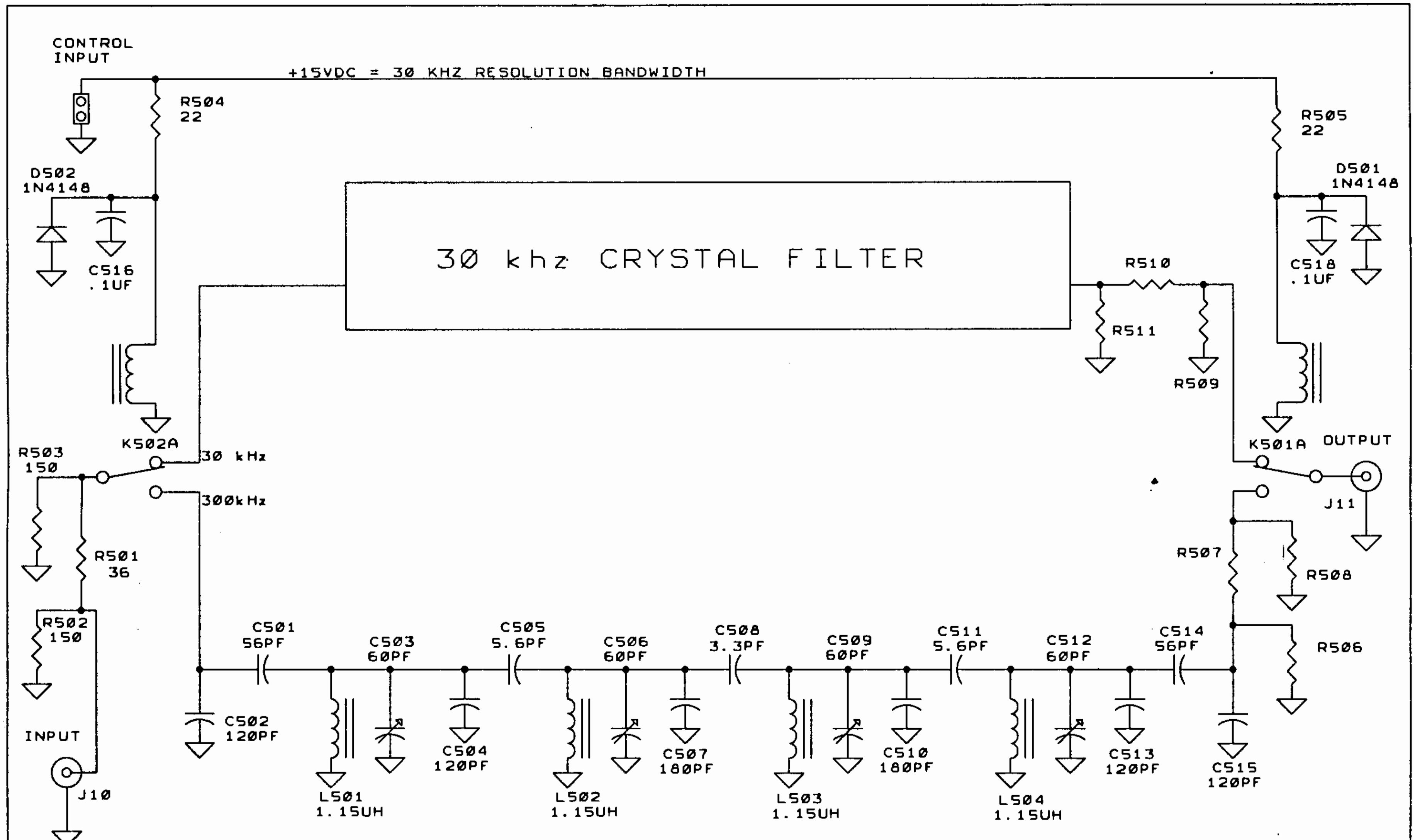
Qty	Reference	Part Description
2	C501, C514	56pf, NP0
4	C502, C504, C513, C515	120pf, NP0
4	C503, C506, C509, C512	60 pf trimmer
2	C505, C511	5.6pf, NP0
2	C510, C507	180pf, NP0
1	C508	3.3pf, NP0
2	C516, C518	.1uf
2	D502, D501	1N4148 or equiv.
2	K501, K502	Relay, Aromat TF2-12V
4	L501, L502, L503, L504	1.15uh, T50-6, 17T, #22
1	R501	36, 1/4w, 5%
2	R502, R503	150, 1/4w, 5%
2	R504, R505	22, 1/4w, 5%

Parts List – VHF Bandpass Filter

Qty	Reference	Part Description
2	L601, L605	10nH, 1" length of #18 wire
3	L602, L603, L604	100 nH, 5T #18 wound on ¼ drill bit
3	C601, C604, C607	10pf, NP0
3	C602, C605, C608	10 pf glass trimmer

Parts List – 70 MHz Low Pass Filter

Qty	Reference	Part Description
2	C701, C705	47pf, NP0
3	C702, C703, C704	100pf, NP0
4	L701, L702, L703, L704	8T, #22, ¼-20 bolt used as form



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