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OCTOBER 1976

second annual receiver issue

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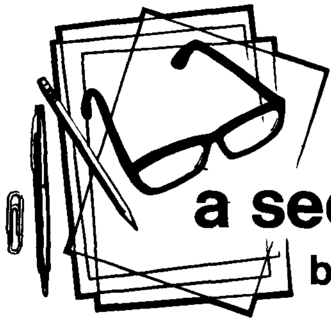

a second look

by Jim Fisk

In their quest for new sources of needed energy, researchers at the Lawrence Livermore Laboratory in California are putting together the world's largest laser in an effort to exploit the ultimate energy source — the same fusion energy as that used by the stars. The 200-foot (60m) long laser, which is scheduled to be completed by early 1977, is expected to produce 25-million watts of electricity. Initially, it will take more energy than this to make it work, so the system will operate at a net energy loss, but scientists expect to upgrade the mammoth to 100- to 300-trillion watts, and by 1980 it should make world history by generating more energy than it actually consumes. When that happens, it will signal the birth of the fusion age and the beginning of civilization's independence from scarce oil, gas, coal and atomic fuels.

Water, the fuel for fusion, has a fantastic amount of energy locked inside: the heavy hydrogen (an isotope called deuterium) in just 75 gallons of water could light New York City for nearly 10 minutes! Although scientists have known about this cheap and inexhaustible energy source for more than thirty years, harnessing it is no easy trick. Stars solve the problem by compressing the hydrogen and heating it to extreme temperatures, and then use their immense gravitational pull to confine the plasma so it can't escape. This works just fine if you're as large as a star (our sun is 864,000 miles in diameter), but here on earth the job will be done with a powerful laser which is carefully focused on a pinhead-sized pellet of heavy hydrogen fuel, instantly heating it to a tremendous temperature. Part of the pellet's outer shell is blasted outward; the inner part, however, implodes and compresses the fuel to 10,000 times its normal density. At the instant this happens the fuel fuses into lithium and emits a shower of heat and light. A thin layer of liquid lithium, which covers the spherical implosion chamber, carries away the heat which is used to drive turbines.

The huge laser, called *Shiva* after the Hindu god of destruction and reproduction, actually consists of twenty 1000-joule lasers, each about 150 feet (46m) long, arranged somewhat like a Gatling gun. Each of the individual lasers consists of a chain of seven elements: a master laser oscillator followed by six laser amplifiers which pumps out an intense pulse of light — equivalent to 10-million megawatts — lasting about a billionth of a second. Each of the 20 beams of carefully shaped and timed laser light simultaneously enter the ports of the implosion chamber and blast the fuel pellet into oblivion. The next pellet is then inserted into the center of the chamber and, once again, implodes after being blasted.

Although there are still a number of problems to be resolved, scientists are confident they will have the first successful fusion reactor in operation by early next year. It will take four or five more years to bring the system on line, but even at that, *Shiva* looks like the most promising new energy source to come down the pike in a good many years.

Jim Fisk, W1DTY
editor-in-chief

More Details? CHECK-OFF Page 134

presstop

FIRST TWO METER WORKED ALL STATES was achieved by KØMQS when he worked K6YNB/KL7 via moonbounce August 17th. The Alaska contact was particularly timely for KØMQS as he'd just nailed his 49th state — Idaho — when he contacted W7UBI on August 2. The QSO with KØMQS was, coincidentally, K6YNB/KL7's first EME contact though the DXpedition station had made some Perseids contacts.

VHF/UHF Bands Continued to offer "DX" treats with long-haul contacts almost a daily happening on two meters during early August. W2BOC has asked for log data from any stations that participated in the June 26-28 openings on two and six meters to aid in sporadic E propagation studies he's making.

SALE OF GENERAL CLASS AMATEUR LICENSES is being checked by FBI in Indianapolis and possibly other cities. An article in the *Indianapolis Star* describes a supposed ring of Amateurs who were telling prospective Amateurs that "they can bypass rigorous testing procedures by purchasing a General Class ham license for $200" from the gang.

A "Mr. Big" in the operation is described in the article as "a government employee in Pennsylvania" who has access to the FCC's computers and, for a price, is "willing to feed the data into the computers to make sure a license will be issued."

The FBI's Agent In Charge in Indianapolis confirmed that an investigation is in progress but had no further details. Contacts at the FCC in Washington were not aware of either the story or the newspaper's report of it. Apparently a number of Indianapolis area Amateurs have been visited by the FBI investigators, and one of them said that an agent told him the Indianapolis investigation is "just the tip of the iceberg!"

FCC IS COMING DOWN HARD on bootleggers operating between the 27-MHz CB band and 10 meters. FCC agents and U.S. Marshalls nailed a dozen of the outlaw operators in a July raid in northern New Jersey, and seized an estimated $10,000 in radio gear.

Amateur Input On Unlicensed and/or out-of-band operators played a big part in the investigation, and FCC/Justice Department people in other parts of the country are also using reports from Amateur radio monitors.

Most Valuable Data is that providing general operating patterns of illegal operation in a given area. Recent discussions with key FCC enforcement people cited reports of specific frequencies, operating times and general location of operators as being of prime importance, with detailed dossiers on one or two violator's activities of lesser interest. Reporters should also specify what they use for monitoring and their location — information can go to any FCC Field Office or Monitoring Station.

"WN" NOVICE PREFIXES will no longer be issued after October 1, and Novice licensees will receive the same prefixes as higher class license holders. In its announcement of the change August 19th, the Commission cited processing problems as the prime reason for the change — its computer has had some difficulty in translating a WN into a WA, WB, or (recently) WD, resulting in the issuance of some duplicate calls.

Present Novice Licensees whose licenses expire after October 1 will also be changing prefixes — the Commission plans to issue them new licenses with the appropriate new prefix — until a new license is actually received, however, a presently-licensed Novice should continue using "WN."

WD Callsigns are being issued or are on the verge of being issued in the 2nd, 4th, 6th and 8th call areas, according to the Commission's release on the Novice change, and WDs are imminent in the 5th, 9th and Ø districts.

TWO LETTER CALL REQUESTS have totaled about 400 since July 1st with 75 or so of those from two-letter-call holders wishing to change. A few requests continue to trickle in, and the fourth district has very few 1x2 calls left. Though the next group of eligible Amateurs will use up some of those remaining, the next big influx of requests is expected in January. Don't be surprised to hear N prefixes and "X" calls such as W9XA by then.

PACKRATS' SOUTH AMERICAN MOONBOUNCE DXpedition in August was a resounding success with 16 stations in 8 countries worked off the moon on 432 MHz. First contact was with K2UYH, providing him with the first E-M-E WAC, followed by W3CCX, F9FT, I5MSH, PAØSSB, LX1DB, K3PGP, VE7BBG, W1JAA, JA1VDV, K8UQA, W4ZXI, W1SL, WØYZS, KØTLH and SM5LE. In addition to E-M-E operation, over 40 OSCAR 7 Mode B contacts were also logged before the group shut down to return to the States.

AMSAT'S ASCII STA has been granted and is good through February 6. More experimental use of ASCII through the satellites is being encouraged.

TOM McMULLEN, W1SL, has joined *Ham Radio* as managing editor. Tom, a well-known VHF/UHF experimenter, has been QST's Assistant Technical Editor.

Charlie Carroll, W1GQO, until recently a member of the ARRL technical staff, has also joined *Ham Radio* as an assistant editor. Welcome aboard!

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optimum design for high-frequency communications receivers

High-performance circuits are described for communications receivers operating in the 10-kHz to 30-MHz range

The design of shortwave receivers has changed significantly in past years. Early receivers used double- or even triple frequency conversion, and selectivity was achieved at a fairly low frequency; for example, 455 kHz. **Fig. 1** is a block diagram of this type receiver. Since the i-f had to be higher or lower than the frequency bands of these receivers, the i-f was usually set below 2 MHz, which resulted in image problems. To overcome this, many tuned circuits were required in the rf stages, and oscillator tracking became quite another problem.

Present-day design avoids these expensive mechanical arrangements thanks to the availability of crystal filters in the 30- to 120-MHz range. If an intermediate frequency higher than the highest frequency of reception is used, three advantages occur:

1. To achieve constant image and i-f suppression, a simple lowpass filter with a cutoff frequency of 31 MHz can be used, guaranteeing at least 80 dB suppression. Such a filter will also substantially reduce local-oscillator power radiation.
2. With both the oscillator and intermediate frequencies higher than the highest frequency of reception, gaps in receiver coverage are eliminated (a receiver with an i-f at 5.5 MHz, for example, precludes tuning between approximately 5 and 6 MHz).
3. The ratio of the oscillator frequency range (maximum/minimum) is, by definition, less than 2:1.

Additional input selectivity can be obtained by using fairly wide cross-modulation and intermodulation filters which operate in fixed-frequency bands. Thus the receiver can use bands of constant width; that is, in 1-MHz intervals.

If we talk about a good shortwave receiver, we must define "good" in terms of receiver electrical characteristics. **Table 1** provides these characteristics for a shortwave receiver that is required for today's operation.

Many amateurs equate "good" with noise figure or signal-to-noise ratio. It has been recently agreed that a noise figure less than 10 dB is not essential for high-frequency amateur receivers. In some military and systems-oriented applications a noise figure of 6 dB or less is required for example, in systems using low-efficiency antennas such as whips. Unfortunately, sometimes some of this equipment is used with large antennas, so a large dynamic range in the receiver is needed despite the low noise figure.

What about oscillator radiation? The technique used by Southcom and being adapted by Atlas requires highly efficient shielding and filtering to avoid oscillator power feedthrough to the antenna input terminal. Commercially designed receivers for the military market must have a maximum of 15 microvolts reradiation of the oscillator signal – a requirement that practically none of the amateur receivers on today's market can fulfill.

new receiver design

Fig. 2 is a block diagram of a reciever with a first i-f of 40.525 MHz. This receiver covers 10 kHz to 30 MHz. The reason for choosing this i-f is that one of the standard i-fs in Europe is 525 kHz (455 kHz in the U.S.). By using a 40-MHz signal derived from an internal frequency standard, conversion from the first to the second i-f can be accomplished easily.

The signal at the antenna passes through a 30-MHz lowpass filter and, depending on the selection of the reception frequency, through either a high- or lowpass filter with a 2-MHz cutoff and one of eight automatically selected bandpass filters in the range of 2 to 30 MHz. The signal is then applied to a balanced power amplifier through an automatic attenuator circuit consisting of negative temperature coefficient and positive temperature coefficient resistors. The 30-MHz filter rejects image frequencies. The 2-MHz highpass filter separates the broadcast range with its often very high field strength from the high-frequency region and prevents BC signals from arriving at the mixer. An independent agc circuit (not a limiter) is used in addition to the normal agc system. This independent agc circuit

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

responds when signals equal to or above 100 mV are fed to the receiver input. This circuit acts as an automatic attenuator and protects all following stages from being overdriven without introducing any measurable distortion of its own. A push-pull power amplifier stage following the input filters uses heavy feedback to keep second- and third-order intermodulation distortion products as low as possible.

tween ±75 Hz to ±6 kHz, which permits optimum matching to the bandwidths required in the various modes of operation (slow CW, frequency-shift telegraphy or broadband telephony). Since individually switchable filters would hardly prove practical for such a large number of different bandwidths, a new double mixer with fixed-tuned filters has been designed (**fig. 3**).

The 525-kHz intermediate frequency with its side-

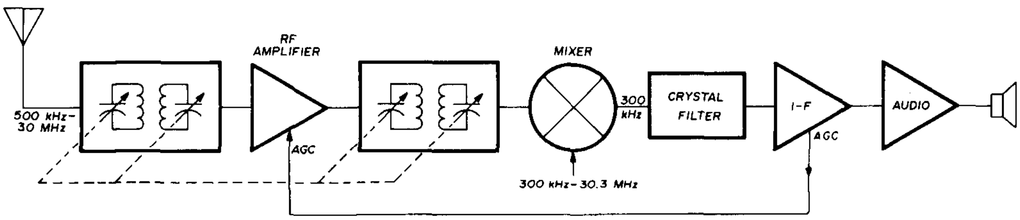


fig. 1. Block diagram of a single-conversion, low i-f, general coverage communications receiver typical of that used by many radio amateurs. The intermediate frequency is usually set below the tuning range of the receiver, as shown here, but this results in image problems on the higher frequencies.

The signal is converted to the first i-f in a high-level +17 dBm or higher double-balanced mixer and applied to a cross-modulation filter. This filter is a low-ripple, 6-pole crystal filter with a bandwidth of ±6 kHz or less, depending on its purpose. This filter suppresses adjacent signals, which might otherwise produce cross modulation in subsequent stages. A high-level double-balanced mixer is also used for converting the first i-f to the second i-f (525 kHz), where main selectivity occurs. The bandwidth of the second i-f stage is selectable in steps be-

table 1. Electrical characteristics for a modern shortwave receiver.

Frequency range	10 kHz to 30 MHz
Antenna connection	50 ohms, unbalanced
Preselection:	lowpass filter
10 kHz-2 MHz	Bandpass filters for
2 - 30 MHz	2 - 3 MHz, 3 - 5 MHz, 5 - 7 MHz
	7 - 10 MHz, 10 - 13 MHz, 13 - 17 MHz
	17 - 22 MHz, 22 - 30 MHz
Setting accuracy	Better than 50 Hz; counter display 10 Hz, 1 Hz preferred
Noise figure	about 10 dB
Modes of operation	CW, MCW, a-m, dsb, ssb, afsk
I-f bandwidth	±75 Hz to ±6 Hz in steps
Shape factor	At least 1:1.45 (6/60 dB down)
I-f and image rejection	Greater than 80 dB
Crossmodulation	Less than 2% demodulation with 100 mV emf interfering signal and less than 10% *with 5 V emf interfering signal*
Blocking through second signal	3 dB with an unmodulated unwanted signal spaced 30 kHz from, and 100 dB above, desired signal of 1μV
Intermodulation distortion	
In band	>50 dB with two 50 mV emf tones
Second order	>90 dB with two 5 mV emf tones
Third order	>80 dB with two 5 mV emf tones
S meter	Should be calibrated in μV. Input voltage range <1 μV to >100 mV.
Outputs for accessories	Audio output 0 dBm
	Agc voltage for diversity reception
	I-f outputs
	Oscillator output for counter and digital programmer

band of ±6 kHz maximum is first converted into the range 52 to 64 kHz. It is then applied to a bandpass filter with a steep-edged selectivity characteristic at 64 kHz and converted to the original i-f. If the oscillator frequency is varied in the proper direction, the shift of the reduced i-f toward the steep filter edge clips or completely suppresses one sideband. Using the same *oscillator frequency for mixing and remixing* prevents any frequency errors from occurring between input and output, which might affect receiver setting accuracy. The position of the sidebands with respect to the carrier also remains unchanged. This arrangement is followed by a similarly designed selective circuit that can limit the other sideband. Similar selective filters are used in both circuits so that the oscillator frequency is symmetrical about the 525-kHz i-f. Thus, a frequency variation by the same amount but in the reverse direction results in a symmetrical bandwidth change. The selectivity at a 500-Hz spacing from the adjusted bandwidth is at least 60 dB, and the ripple within the i-f passband is about ±1.5 dB.

If filters with the same selectivity are used, double mixing ensures an identical, symmetrical skirt selectivity of the i-f passband characteristic, constant skirt selectivity for each bandwidth, and a phase delay characteristic symmetrical with the center frequency. With a-m *reception, therefore, no distortion occurs due to delay* time. With commonly used crystal or mechanical filters this high phase and gain linearity can't be achieved under any circumstances.

This method also provides a simple means of suppressing the upper or lower sideband during ssb reception. By setting the oscillator of the first and second selective circuits to a fixed frequency, variable limiting of the remaining sideband is possible.

Both conversion oscillators are synchronized with an internal reference oscillator to increase circuit stability. This avoids an impermissible variation or shift of the center frequency, especially in narrow bandwidths.

Advertised crystal filters, which are 8- to 10-pole

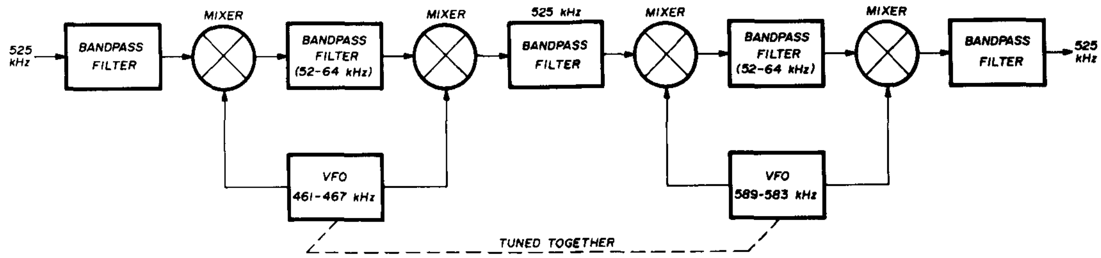


fig. 3. A system for obtaining variable i-f bandwidth through the use of a double mixing scheme. This system is used in the receiver of fig. 2 to provide selectable bandwidths from 6 kHz to 75 kHz without the expense of a large number of separate i-f bandpass filters.

designs with shape factors between 1.4:1 and 1.2:1, have two unpleasant effects:

1. The extremely steep skirt selectivity presents a problem for the agc circuit because of the high group delay and phase shift, which cannot be compensated. In almost all cases strong interfering signals at the edges of the filter response band will make the agc pump. This instability introduces distortion and overshoot.

2. Because of their high Q, and despite their fairly wide bandwidth, these filters produce appreciable ringing.

In addition, it's important to remember that, under the circumstances in which radio amateurs use their receivers, the i-f bandwidth for ssb reception should be between 1.9 and 2.4 kHz to avoid psychological fatigue caused by unpleasant hissing noises. More than ten years ago the bandwidth of the famous Collins KWM2 was restricted to 2.1 kHz for this reason.

I-f amplifier and gain control. The i-f amplifier boosts the incoming signal to an amplitude sufficient for distortion-free demodulation and ensures a constant output voltage by means of gain control. Its amplifying action is such that the inherent noise from the rf section is adequate to drive the succeeding agc amplifier to full output. If you analyze practically any amateur transceiver now on the market, you will discover that these receivers don't have enough i-f gain. The reason for this is obvious: most manufacturers want to avoid the expense of careful shielding, which is required if the gain is more than 80 dB. However, to obtain enough audio

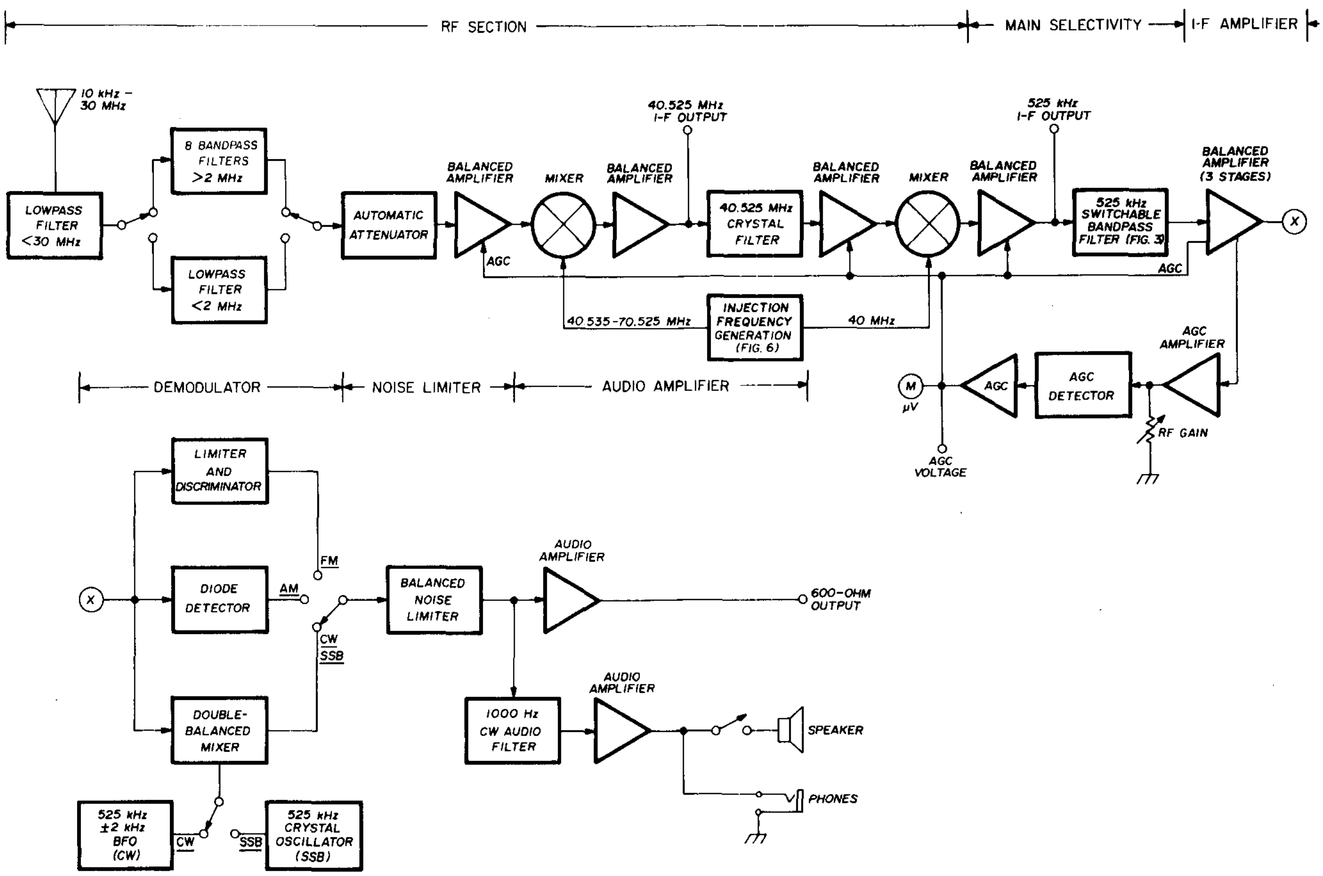


fig. 2. Block diagram of a high-performance, modern design communications receiver with an i-f at 40.525 MHz which overcomes many of the problems of previous designs. The noise figure of this receiver is about 10 dB, more than adequate for home-station use on any of the amateur high-frequency bands. Cross modulation, intermodulation distortion, and other operating characteristics of this receiver are listed in table 1.

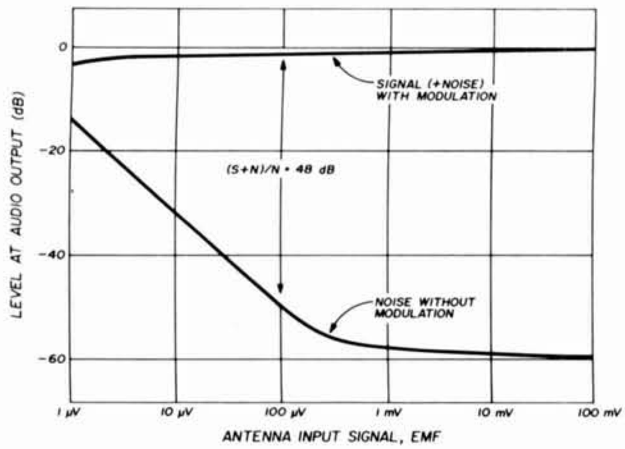
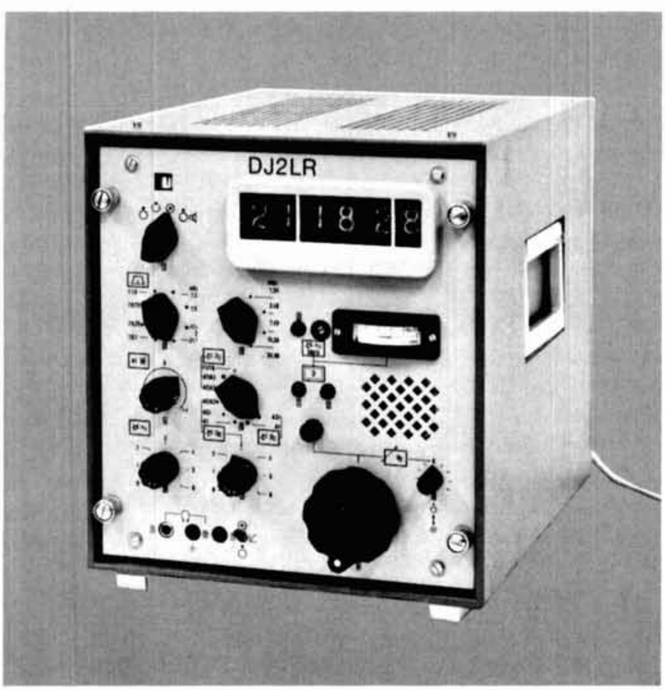


fig. 4. Optimum selection of agc action for perfect signal-to-noise performance.

output, a total gain of about 90 dB from the antenna input to the i-f output into the demodulator is required to obtain distortion-free demodulation. Unfortunately many designers use high-gain rf amplifiers and high-gain first mixers, both of which overdrive the second mixer. A sufficiently low overall noise figure can be achieved with little rf gain, as explained below.

Modern integrated circuits, such as the Plessey SL600 series or the Motorola MC1349 and MC1590, offer a good choice for a low-distortion, stable i-f amplifier with a minimum of shielding. Such a design permits sufficient agc to reduce unpleasant audio level changes. In addition, appropriate distribution of the agc voltage must be applied to the various stages to ensure a linear increase in the signal-to-noise ratio at small input voltages. **Fig. 4** shows the agc response provided by the design discussed above. The agc circuit must have high gain; through pure control action input voltages up to 100 mV emf can be reduced to a residual error of ±1 dB without additional driving (forward-acting regulation). Furthermore, the agc voltage must be applied in appropriate levels to the various stages to ensure a linear increase in the signal-to-noise ratio at small input voltages. The result is a signal-to-noise characteristic much better than previously known.

Partially home-built receiver which tunes from 1.5 to 30 MHz. The PC boards were removed from a lab model Telefunken receiver, and the circuitry was redesigned using the techniques described in this article. The third-order intercept is +30 dBm, dynamic range is 120 dB or more, and the noise figure is 8 dB. Tuning accuracy is 1 part in 10 million.



The agc response time and the type of rectification must be different for each mode of operation. Peak rectification with a short response rise time of about 20 dB per 10 mS should be used for rapid leveling of the gain to the nominal value when receiving CW or ssb signals; a too-short response time does, however, lead to blocking by interference pulses. Average-reading rectification is recommended in the a-m mode to maintain receiver dynamic characteristics.

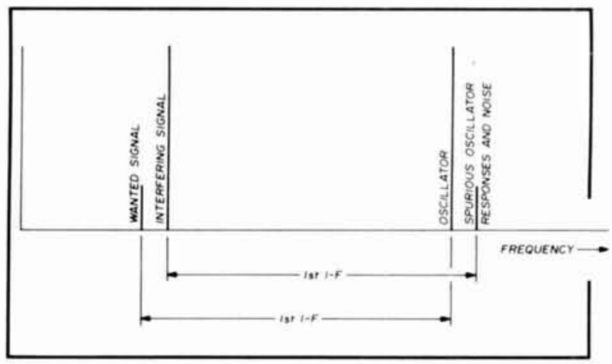


fig. 5. The noise sideband characteristics of the high-frequency injection oscillator affects receiver sensitivity because strong interfering signals at the mixer may convert the noise sidebands into the center of the i-f passband.

The agc voltage fall time should be selectable to obtain a gain variation between 5 and 50 dB per second. The slow response fall time maintains the dynamic characteristics and prevents any amplification of unwanted noise during CW pulse intervals. It also takes effect with a-m signals suffering selective carrier fading; it does not eliminate the distortion but prevents audio volume from increasing.

The modulation frequency of the incoming signal, which is present in the control lines during agc operation, must be filtered – without affecting the dynamic control characteristics – to such an extent that no additional distortion arises in a-m reception due to inverse modulation. In ssb reception, with intermodulation and little difference between the sideband frequencies ($f = 100\ Hz$), intermodulation products ($2f_1 - f_2$) because of inverse modulation must be suppressed at least 50 dB.

Demodulator and af amplifier. The demodulators for the various operation modes are connected to the output of the i-f amplifier. A-m signals are rectified in a diode fig. 6. The frequency synthesizer with linear master oscillator which is used in the high-performance communications receiver. A complete description of this circuit, which uses four separate oscillators, is contained in the text.

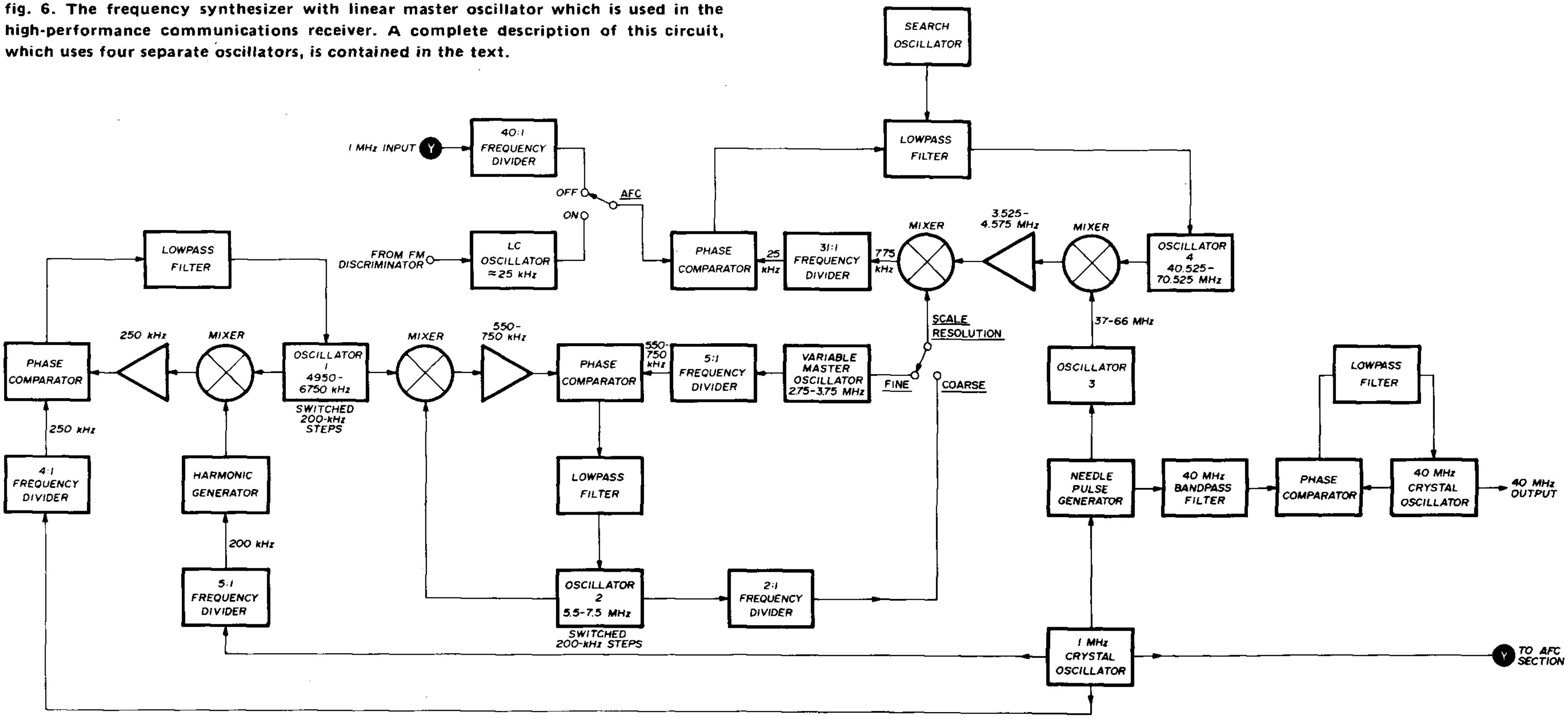
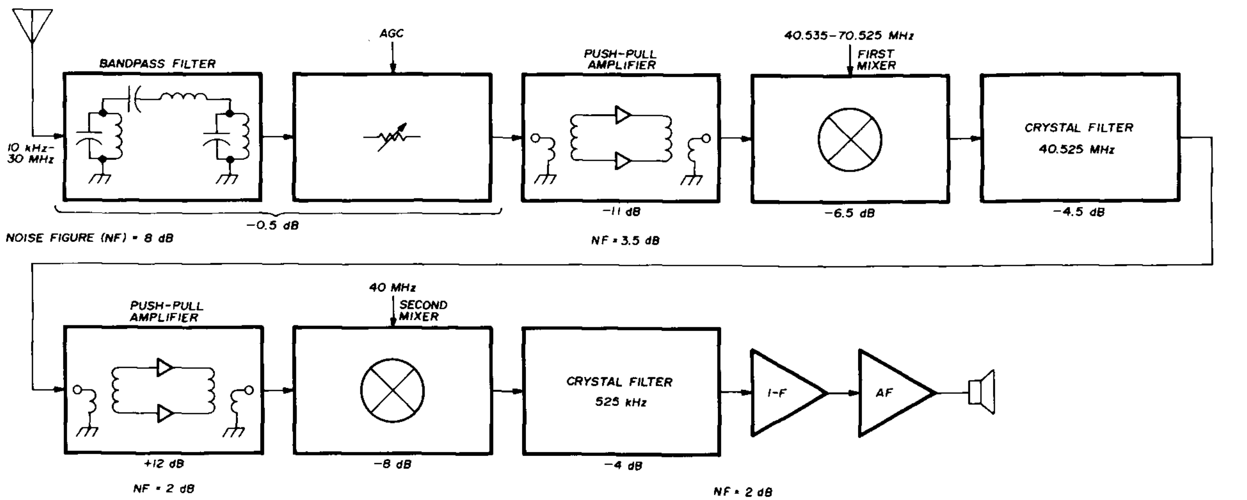


fig. 7. Gain distribution of a high-performance, high-frequency receiver. High dynamic range, as shown in fig. 8, is obtained by using as little gain as possible while keeping the overall noise figure below 10 dB.



circuit. The i-f, CW, and ssb signals are converted into audio signals in a double-balanced mixer which is driven by a vfo for CW signals, and by a high-precision crystal oscillator for ssb signals. Frequency-modulated waves (maximum bandwidth 6 kHz) are demodulated in a discriminator preceded by a limiter.

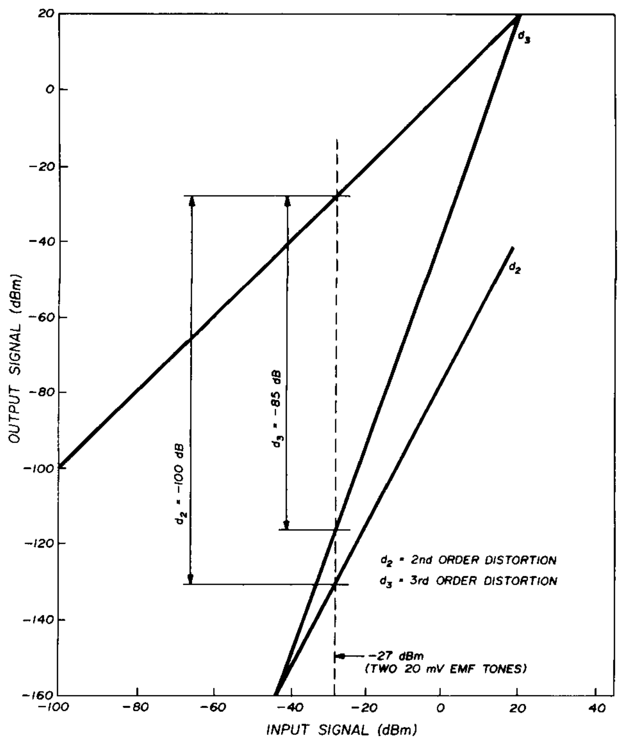


fig. 8. Dynamic performance of the receiver shown in fig. 7. The third-order intercept point is at +20 dBm. At −27 dBm input (two 20 mV emf signals), third-order products (d_3) are down −85 dB and second-order distortion (d_2) is down −100 dB.

The audio-frequency signal is applied through the function selector from the demodulators to an adjustable, balanced noise limiter, then to audio amplifiers. The amplifier with an output impedance of 600 ohms is associated with a balanced-line output. Starting from 0 dBm (1 mW into 600 ohms with 100% modulation), its output level is adjustable by ±10 dB. The second amplifier feeds a built-in speaker. It also contains a switch-selected 1000-Hz filter (bandwidth 200 Hz), which improves CW audio quality.

oscillator

The receiver frequency stability and setting accuracy depends on oscillator accuracy. The oscillator characteristics also affect receiver sensitivity, since strong interfering signals arriving at the mixer as a result of the receiver input broadband characteristic can convert noise sidebands and spurious oscillator responses into intermediate frequencies (**fig. 5**).

The oscillator frequency of the synthesizer (**fig. 6**), whose output is 40.525 to 70.525 MHz (**oscillator 4**), is

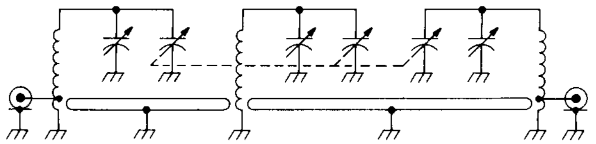


fig. 9. Tunable preselector which may be used to suppress local, high-level signals. Bandpass curve for this preselector is plotted in fig. 10.

derived in several phase-controlled circuits from selected harmonics of a 1-MHz crystal oscillator and the frequency of the variable-frequency master oscillator. The crystal in the 1-MHz oscillator is housed in a proportionally-controlled oven for stability. The frequencies derived from this 1-MHz oscillator are more accurate than those of the variable-frequency master oscillator, which determines receiver setting accuracy and stability.

While some receiver designs use frequency synthesizers which are switchable in steps down to 1 Hz, this much resolution is somewhat prohibitive for a search receiver. The lock time of such a frequency synthesizer would not permit easy tuning and, therefore, this type of receiver should only be used for channelized point-to-point communications. The method used here, derived from the Rohde & Schwarz EK07 shortwave receiver, was designed in 1957. A similar method has been used in receivers such as the National HRO500 and HRO600.

Basically, the 40.525 to 70.525 MHz frequency range is converted to an auxiliary i-f of 2.75 - 3.75 MHz. A frequency- and phase-sensitive double-balanced modulator is used as a comparator together with the master oscillator to determine oscillator 4 output frequency. The master oscillator, which can be tuned over a 1-MHz range (2.75 - 3.75 MHz) capacitor, exhibits a linear frequency characteristic. This linearity permits the use

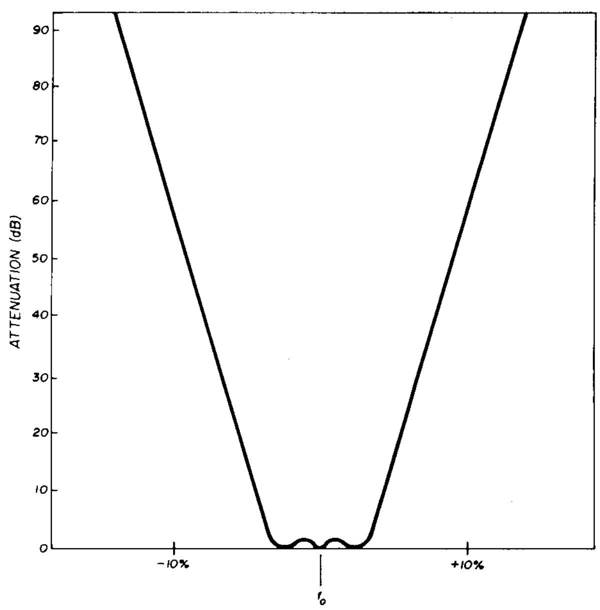


fig. 10. Selectivity curve for the preselector circuit of fig. 9.

of a directly calibrated scale with an ultimate resolution of 50 Hz.

To increase the variable master-oscillator frequency stability, all frequency-determining parts are housed in an oven and maintained at 60°C (140°F). Short needle pulses are derived from the 1-MHz oscillator output and are used to lock in oscillator 3 in 1-MHz steps. This frequency is required in a down-converter system to to 775 kHz. This frequency is then applied to a 31:1 frequency divider. A phase comparator operating on the scanning principle compares the 25-kHz frequency difference with the output from the 1-MHz crystal oscillator. This output voltage adjusts oscillator 4 (residual superimposed ac voltages have been suppressed by the lowpass filter). If this circuit is not controlled, as for instance when the MHz range is changed, a search oscilla-

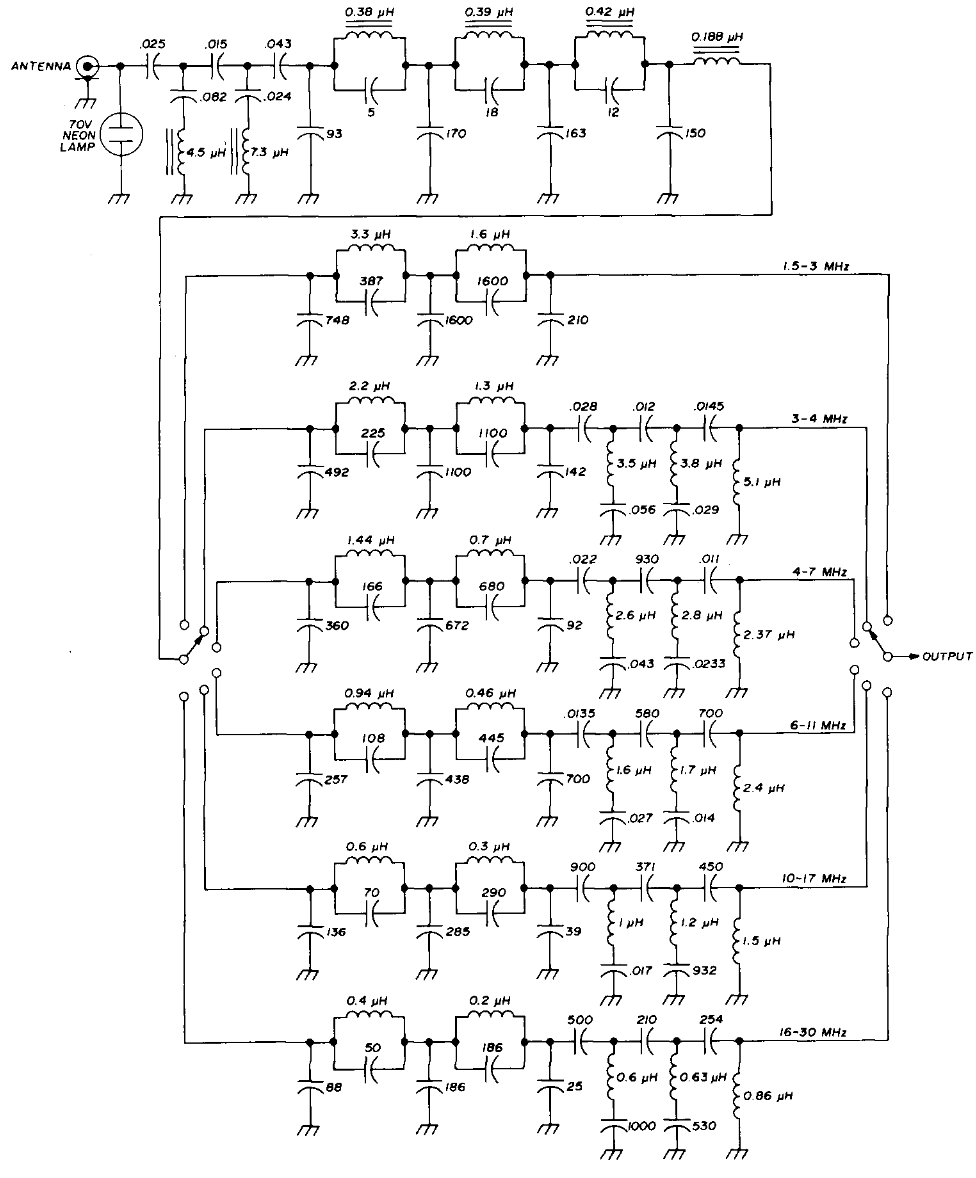


fig. 11. Highpass-lowpass elliptic filter (1.45 MHz to 32 MHz) for use at the front end of a high-frequency communications receiver.

obtain 2.75 to 3.75 MHz for the master oscillator. The same needle pulses are used in a phase-locked loop system to synchronize the frequency of the 40-MHz crystal oscillator, which converts the first i-f (40.525 MHz) to the second i-f (525 kHz) with the 40th harmonic of the 1-MHz crystal oscillator. Oscillator 4 consists of six single oscillators, each of which is tuned over a range of about 5 MHz by tuning diodes.

The frequencies of oscillators 3 and 4 are converted tor generates a sawtooth voltage that sweeps the oscillator range until its frequency is in the lock-in range of the control circuit. The frequency divider reduces the control circuit gain to the 31st part, thus increasing stability.

When the afc circuit is switched on, the 25-kHz reference frequency is not derived from the 1-MHz crystal oscillator but from an LC oscillator, whose frequency can be varied by a varactor. If the receiver is not exactly tuned to the carrier, the output voltage of a narrowband fm discriminator at the output of the i-f section changes the LC oscillator frequency and hence that of oscillator 4, so that the carrier lies at midband except for a residual control error of about 5% of the original deviation. The afc circuit permits adjustment to within ±1.5 kHz.

Free-running oscillator 4 has a high signal-to-noise ratio; for example, more than 140 dB per hertz of bandwidth at a 30-kHz spacing from the carrier. The control circuit has a narrow bandwidth to minimize its effect on the signal-to-noise ratio (circuit gain of 1 at about 5 kHz). Low-frequency disturbances, such as shock or noise caused by operating the switches or by the built-in monitoring loudspeaker, are avoided by mounting parts of the oscillator on shockmounts.

The mode of operation described is ideal for rapidly scanning wide frequency ranges, since a range of 1 MHz is covered in ten rotations of the tuning knob. The mode with high-scale resolution provides the full setting accuracy of the receiver. For this purpose, the tuning knob for the 100-kHz ranges is turned from its end position to the required 100-kHz step. The signal fed to the control circuit of oscillator 4 is a mixture of the tenth part of the master-oscillator frequency and a frequency derived from the 1-MHz oscillator. This signal is switch-selected in 100-kHz steps.

A frequency divider and a harmonic generator produce a 200-kHz spectrum from the 1-MHz frequency of the oscillator. Oscillator 1, like oscillator 2, is switched in 200-kHz steps by the 100-kHz tuning knob and converts one of the spectral lines in the mixer to 250 kHz. The subsequent phase bridge compares this frequency with a 250-kHz frequency derived from the 1-MHz oscillator by a 4:1 frequency divider, thereby producing the control voltage for oscillator 1. Oscillator

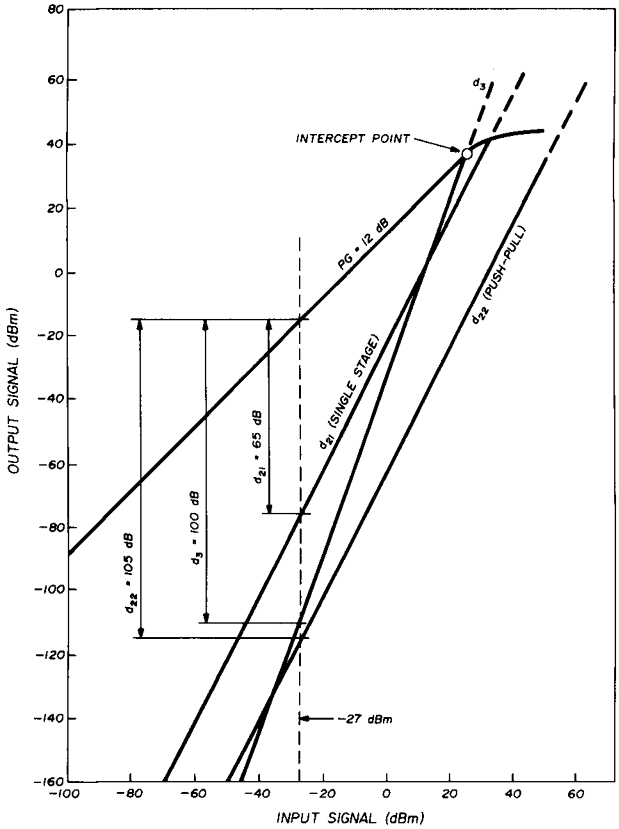


fig. 13. Dynamic performance of a new type of push-pull rf amplifier stage (figs. 14A, 14B, and 14C). Power gain is about 12 dB. With an input of −27 dBm (two signals), third-order distortion products are down 100 dB and second-order distortion is down 105 dB. Third-order intercept point occurs at an input of about +32 dBm

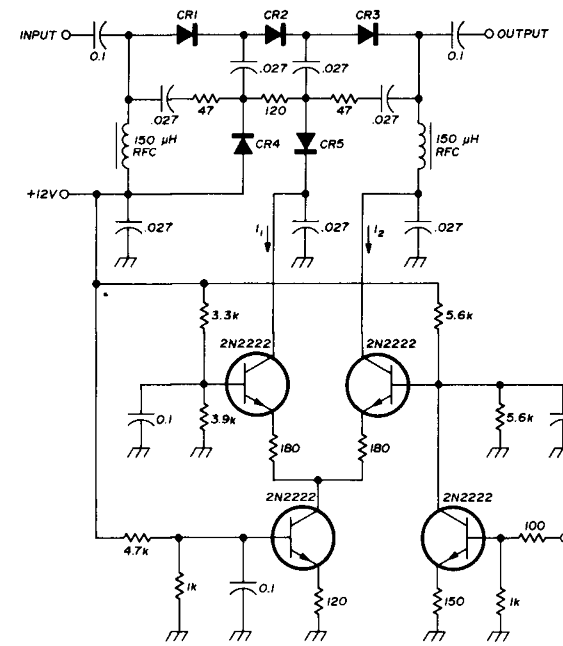
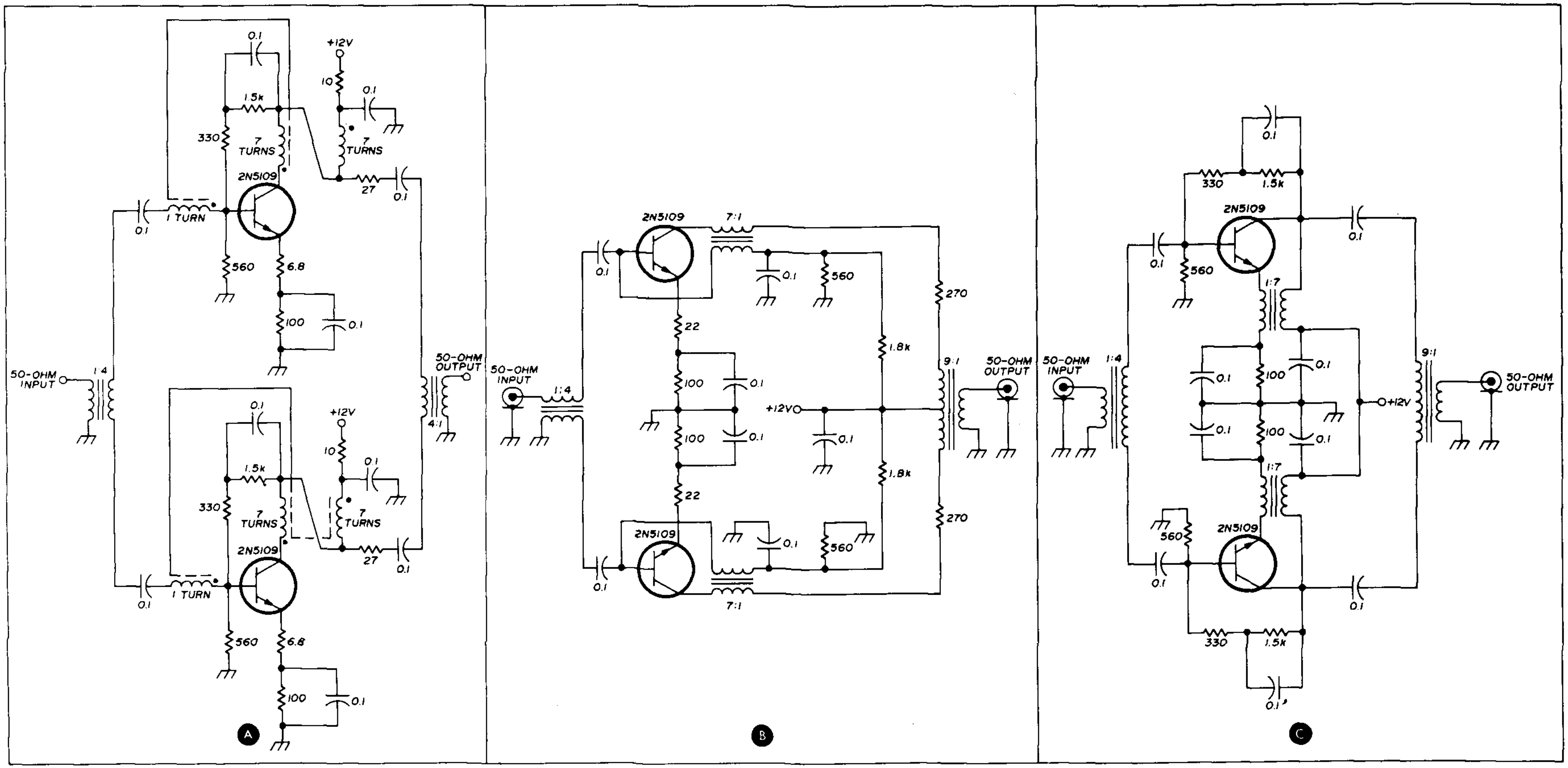


fig. 12. PIN diode attenuator with dc amplifier for use in a high-performance communications receiver. All diodes are Hewlett-Packard PIN diodes, type HP5082-3081.

1 thus operates with the high accuracy of the 1-MHz crystal oscillator on one of the frequencies divided into 200-kHz steps between 4950 and 6750 kHz.

In a further phase-control circuit, the frequencies of oscillators 1 and 2 are converted to the 500 - 700 kHz range and compared in the phase bridge with the master oscillator frequency, which is divided in a ratio of 5:1. After separating the ac voltage components in the low-pass filter, the output voltage of the phase bridge controls the frequency of oscillator 2. By dividing the oscillator frequency in half, the frequency of the master oscillator, 2750 - 3750 kHz, is again obtained, but this time in a switch-selected, 100-kHz range. Because of the frequency divisions (5:1 and 2:1), the master oscillator covers only a tenth of the range so that receiver scale resolution and accuracy are increased by ten times. With a counter, the scale resolution may be 10 Hz or better and the BCD outputs may be used to program a direction finder through a remote-control system.

This type of synthesizer permits continuous tuning for many purposes. However, quasi-continuous tuning in steps of 10 Hz is admissible, which simplifies the whole synthesizer dramatically. This synthesizer has been described in detail to show what must be considered when building a practically spurious-free, low-sideband-noise local oscillator.



A. Circuit using current and voltage feedback for improved linearity and transformers for stabilizing input and output impedance from 100 kHz to 200 MHz. Amplifier has low output impedance; the two 27-ohm resistors are used to increase this value.

B. An improved version of the circuit in A. Output impedance is very high. Agc can be applied by replacing the two 270-ohm resistors with a single pin diode shunt regulator. This circuit is less expensive and simpler than a constant-impedance, T-attenuator.

C. Low-noise version with emitter feedback for extremely high input and output impedance. Amplification is 7 (1:7 turns ratio) and input impedance is 300 ohms divided by 7, or 47 ohms. Output impedance is several hundred ohms. Noise figure of less than 2 dB can be obtained.

fig. 14. Three push-pull arrangements for high linearity input rf stages.

The receiver described above is the result of combining various optimized stages and careful planning. To predetermine the overall technical performance, it's required to start with a block diagram containing vital information, such as gain, dynamic range expressed in second, third, and higher-order intermodulation distortion, and noise figures.

Fig. 7 is a block diagram of a receiver with a first high i-f and shows the amplification or losses of each stage. As shown, the input bandpass filter and the automatic attenuator have a 0.5-dB loss. The push-pull rf amplifier provides approximately 11 dB gain, which will compensate for the losses of the passive, high-level, double-balanced mixer and the losses of the 40.525-kHz crystal filter. The overall gain from the rf input to the 40.525-MHz push-pull amplifier is 0 dB. The stages from 40.525 to 525 kHz are almost identical; however, the second-order intermodulation of the second rf amplifier does not exceed the high values of the first. This is not necessary because of the cross-modulation crystal filter. The second high-level double-balanced mixer has somewhat higher losses but provides the same or even better intermodulation distortion suppression. The first 525-kHz filter is about ±6 kHz wide, and the overall gain up to this point is 0 dB.

The overall noise figure, which can be calculated from the block diagram, is 8 dB. The second and third intermodulation distortion performance of the rf input is shown in **fig. 8.** The main idea of obtaining a high dynamic range lies in the concept of having as little gain as possible while keeping the overall noise figure below 10 dB. This is achieved by carefully selecting the characteristics of each of the individual stages, as will be described later.

For a few rare applications it may be necessary to suppress frequencies in the immediate vicinity of a desired signal, for example, where a transmitter and a receiver are used simultaneously. Under these rare circumstances an rf preselector, as shown in **fig. 9,** is an absolute necessity. Because of the preselector's high selectivity, a 10-volt emf signal, ±10% away from the frequency of reception should not create intermodulation distortion more than -9 dB down. However, for these purposes special transceivers with duplex capabilities are used. These preselective filters have about -4 to -5 dB insertion loss. A typical response is shown in **fig. 10.**

circuit analysis

Because of the new approach in the design of this receiver, the circuits of a number of stages will differ substantially from those commonly used. To give a better understanding of the overall system, the most important stages are shown and explained here.

Input filter. In most cases the highly selective preselector of **fig. 9** is not required, so a combination of elliptical filters selected for the frequency ranges is adequate. Third- and higher-order, odd-number intermodulation distortion products cannot be overcome by selectivity unless filters are 10 kHz wide or less. Second-order

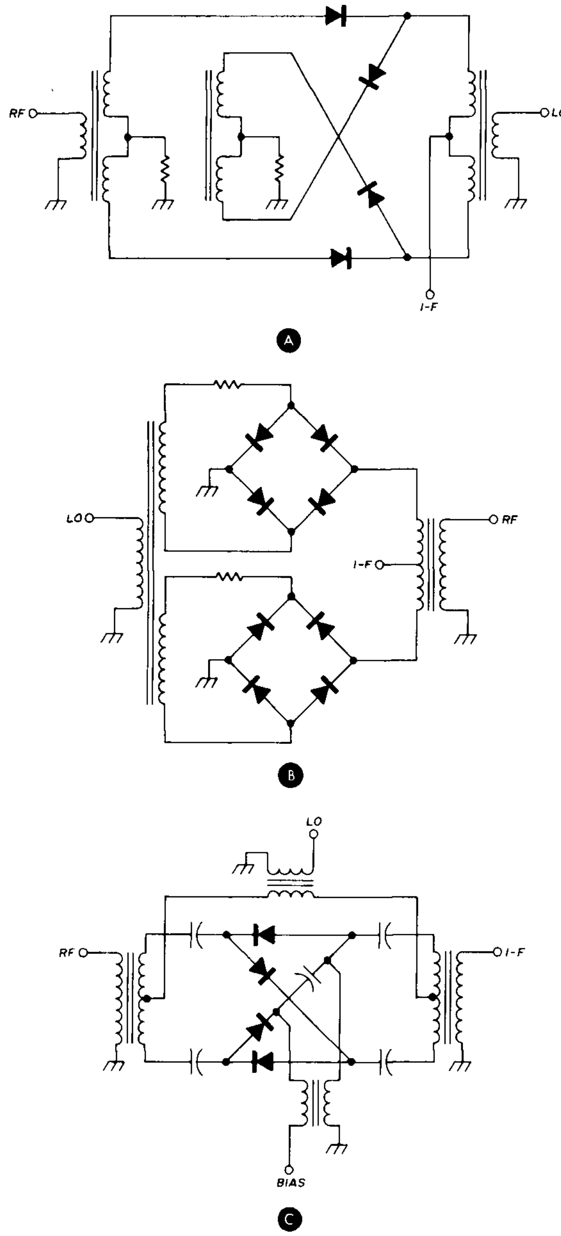


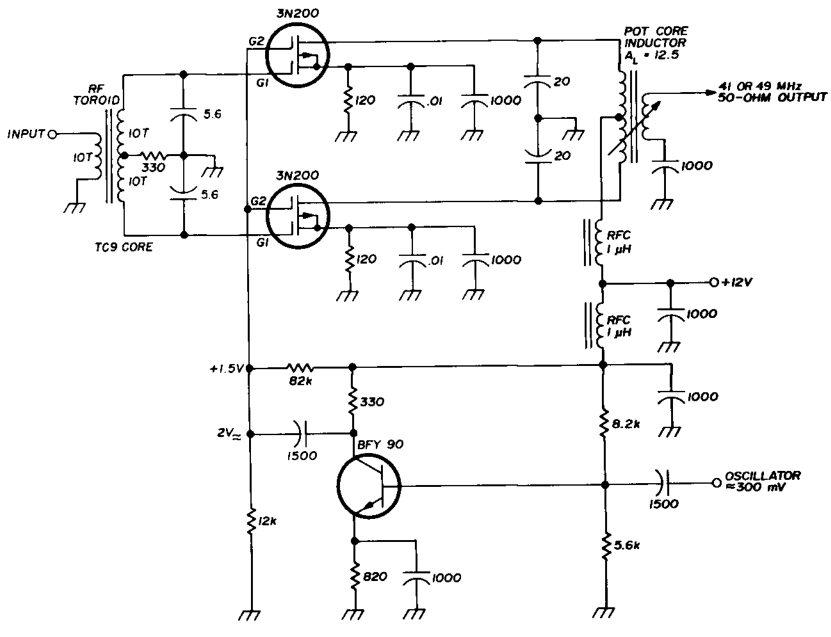
fig. 15. Double-balanced mixers. A and B are medium- and high-power configurations; C is a double-balanced mixer for uhf/shf with adjustable bias.

intermodulation distortion products can be improved by about 40 dB with these switchable filters.

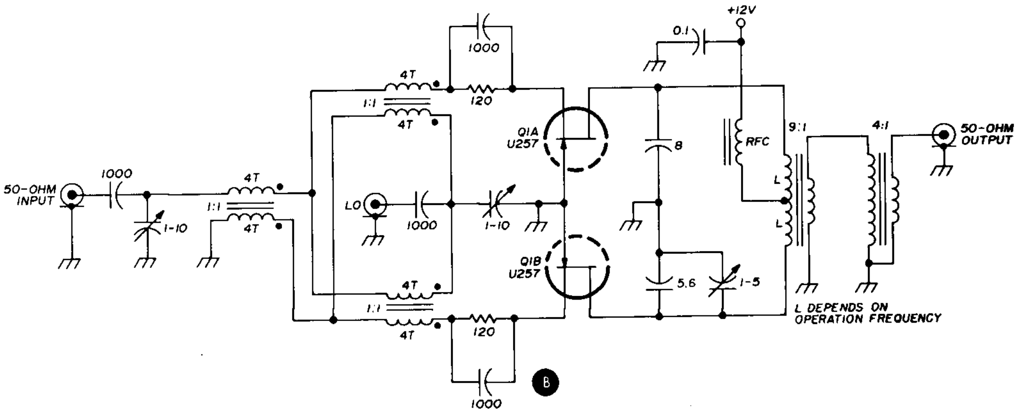
Special Bessel-Cauer elliptical filters are used with a given Chebishev response in the passband. This is absolutely necessary, because 50-ohm impedance matching is required in the stopband and passband so that filters can be cascaded. **Fig. 11** shows a modern input filter covering the range 1.5 to 30 MHz in six steps.

A highpass-lowpass input bandpass filter suppresses unwanted broadcast signals, and the lowpass filter section, together with the following lowpass section of the individual bandpass filter, guarantees more than 90 dB image suppression. The advantage of this technique is that these filters can be aligned very easily with the help

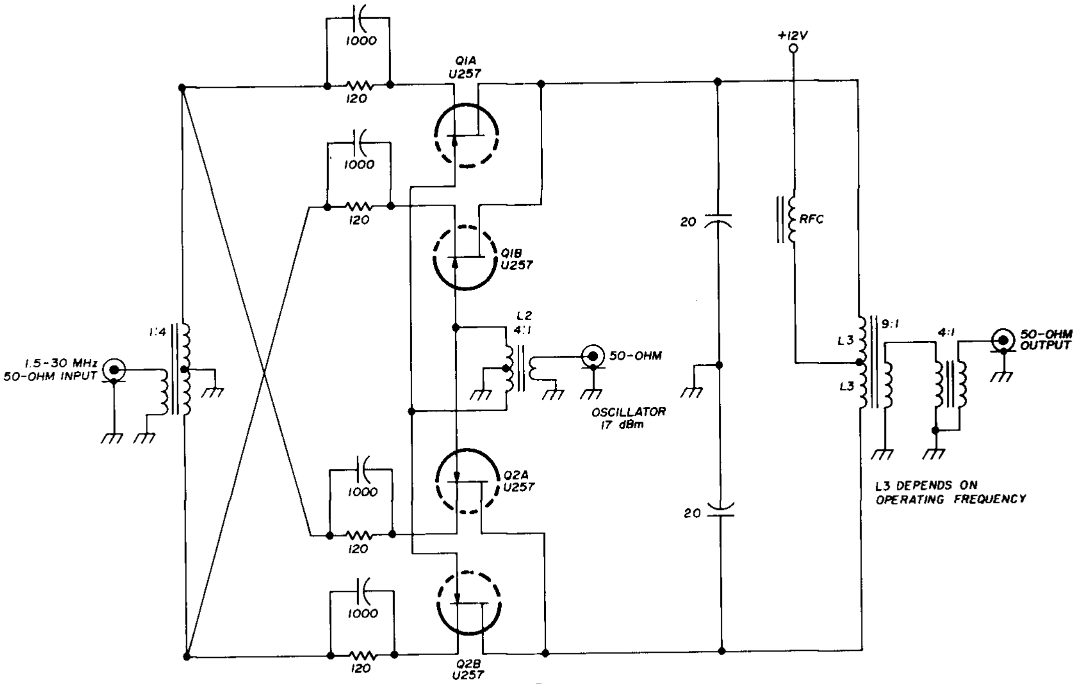
fig. 16. Examples of fet mixers. A balanced mixer with high input impedance (1 kilohm) is shown in A. A two-tone, 176 mV emf signal produces third-order IMD 68 dB down. A balanced fet mixer with 50-ohm input is shown in B (same performance as the circuit in A). C is a double-balanced mixer suggested by Ed Oxner of Siliconix. Its performance is 3 dB better than that of the circuit in B. An improved version of this mixer is shown in D (on facing page), also by Mr. Oxner.



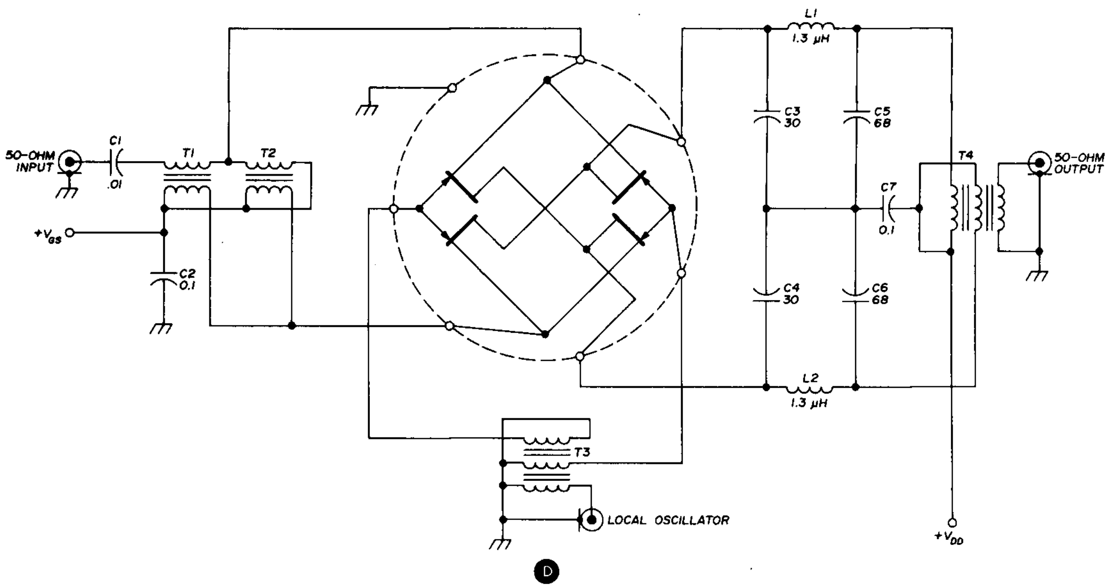
A



B



C



of a sweeper generator and an oscilloscope, and are much less complex than a mechanical tracking arrangement.

Input attenuator. Modern communications receivers should have low-distortion, automatic input attenuators that are activated at input signal levels above 100 μV emf. For the frequency range from 1 MHz to 30 MHz, pin diodes such as the HP5082-3081 are highly recommended in a double-T configuration (**fig. 12**), where the line impedance must remain fairly constant because of filter matching. The intermodulation distortion products of an attenuator such as this are about 85 dB down for two 1-volt emf signals. This means that the attenuator does not add any serious distortion to the system.

Rf input stage. In the past only certain vacuum tubes were considered to be sufficiently linear for rf front ends. However, recently it has been found out that if both voltage and current feedback are used in a transistorized rf stage, much better linearity can be obtained than with any receiving-type tube. The second-order intermodulation distortion products can be suppressed by almost another 40 dB over a single stage when using a push-pull arrangement. **Fig. 13** shows the intermodulation distortion performance of a single stage and a push-pull stage, using the highly linear 2N5109 uhf power transistors made by RCA. The input and output impedances are stabilized with a third feedback network using a wideband transformer.

Fig. 14 shows three push-pull arrangements suitable for this purpose. **Fig. 14A**, which was published earlier,[2,3] uses voltage and current feedback to minimize intermodulation distortion, while the transformers act as impedance stabilizing devices. The cores recommended are F625-0-TC9 for 1.5-30 MHz or F625-9-Q1 for the higher frequencies (made by Indiana General).

Fig. 14B shows an improved version of the circuit in **fig. 14A.** While the latter produces a 50-ohm output impedance, the circuits of **figs. 14B** and **14C** are basically constant-current devices. Because of this, a push-pull pin diode attenuator using only two diodes can be used (5082-3081 made by Hewlett-Packard). With the constant-impedance attenuator, **fig. 12**, second-order intermodulation distortion products are also further suppressed. However, because of the unbypassed emitter resistors, the additional noise contribution will not permit noise figures below 3 to 4 dB.

Fig. 14C show a low-noise arrangement in which

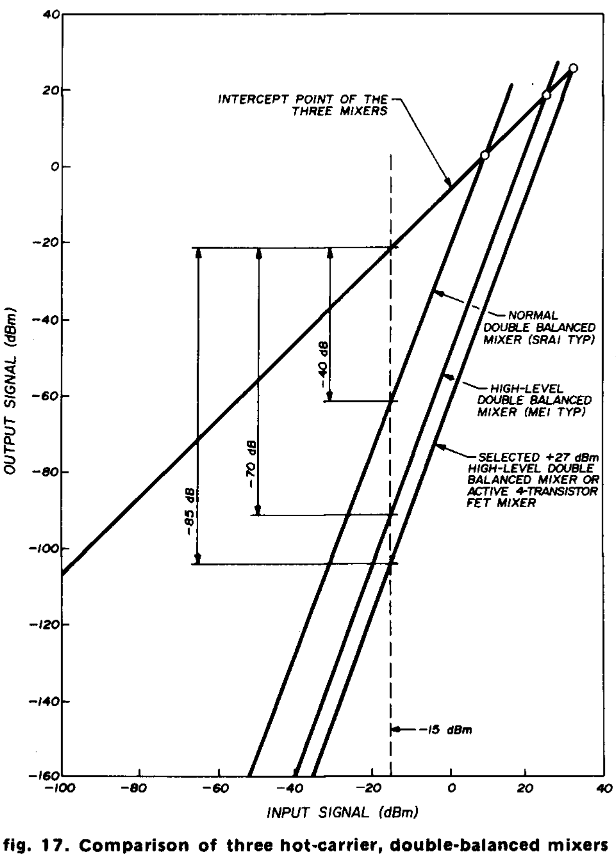


fig. 17. Comparison of three hot-carrier, double-balanced mixers showing third-order IMD performance.

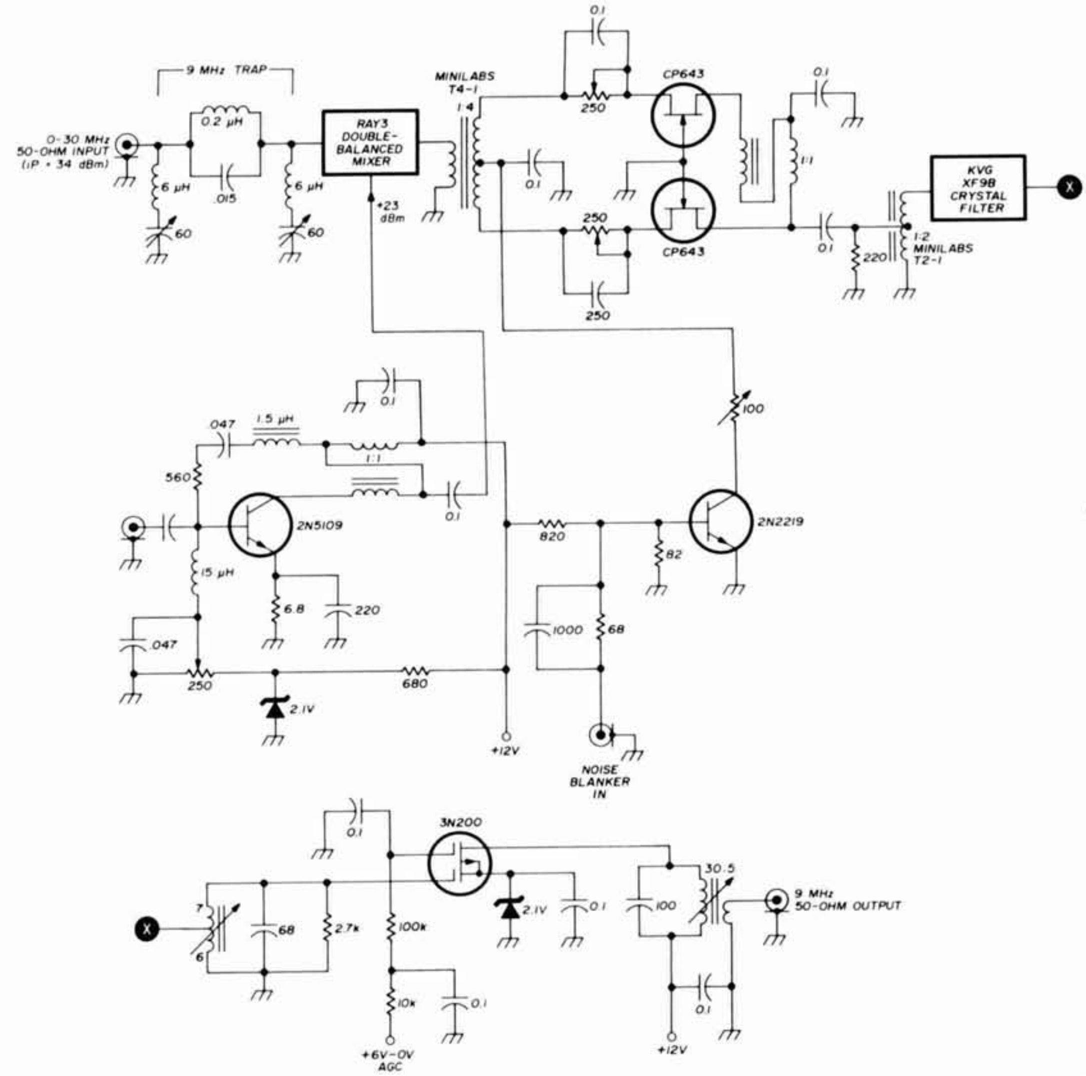
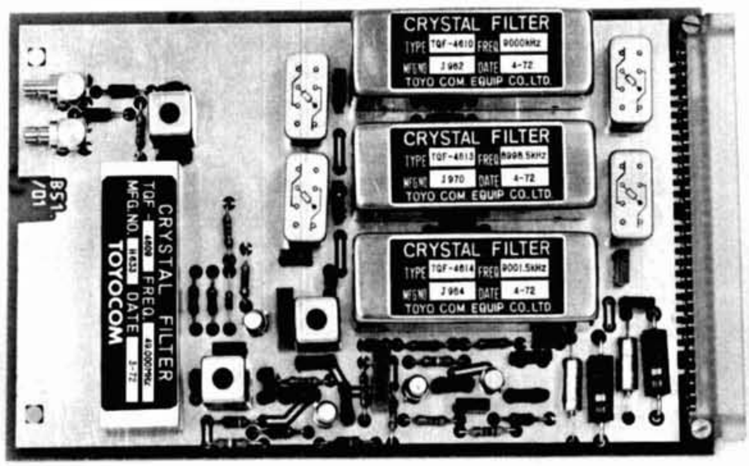


fig. 18. Improved version of Martin's rf input stage. The 9-MHz trap suppresses i-f feedthrough, and the push-pull fets increase dynamic range by 4 dB relative to the original design. In addition, a noise blanker can be used with 1-μs switching time because of the push-pull arrangement.



I-f section of the high frequency receiver built by DJ2LR. Crystal filters are switched into the circuit with small relays. The i-f amplifiers are located along the bottom portion of the board.

emitter feedback through a transformer is accomplished. As a result, the input and output impedances are extremely high, and the voltage feedback permits choosing suitable values. However, this push-pull stage will provide noise figures of 2 dB or less, while still having the same basic dynamic range.

In some cases it's possible to build a receiver without an rf input stage. Especially for fixed operation, the antenna system will have a noise temperature much higher than 300 KT_0 , which means that the various noise sources being picked up by the antenna system will provide a noise floor significantly higher than the receiver-input noise. Under these circumstances, receiver sensitivities up to 20 dB do not degrade overall performance. For applications like mobile radios, however, such sensitivity is not good enough. Since mobile antenna efficiency is between 10 and 30%, the noise pickup is reduced by the same amount and, therefore, noise figures less than 10 dB are vital. In cases of fixed stations new circuitry can be used, which is discussed in the next section.

Double-balanced wideband mixers. It has been found that push-pull arrangements have advantages over single-stage mixers; however, because of the high input impedance of tubes, few attempts have been made to use push-pull or double-balanced mixers with vacuum tubes in high-frequency communications receivers. Because of some other disadvantages the use of beam-deflection tubes was not a breakthrough. Since the introduction of low noise hot-carrier diodes, double-balanced mixers for high-level operation have been constructed and various configurations have been used, as shown in **fig. 15.**

Only recently double-balanced mixers with field-effect transistors have been used which exhibit excellent third-order intermodulation distortion suppression. **Fig. 16** shows four tested configurations with fets. **Figs. 16B** and **16C** use matched pairs.

Fig. 17 shows the third-order intermodulation distortion suppression of low-level double-balanced diode mixers, medium-level double-balanced mixers, and high-level double-balanced mixers, including double-balanced, field-effect transistor mixers. As can be seen, the fet mixer shows significantly higher signal-handling capability. Most circuits using field-effect transistors in double-balanced mixers were originated by Ed Oxner (ex W9PRZ) of Siliconix Incorporated.

Because of the inherent gain capability, fet double-balanced mixers offer advantages in test instruments but their application in shortwave receivers remains debatable. These circuits are very sensitive to load termination, and the input impedance of most crystal filters found on the market changes significantly over frequencies outside the passband. To reduce this effect, a resistive termination is required at the input (valid for all mixers, independent of configuration). However, the remaining impedance jump of 1 to 2 for out-of-passband operation reduces performance greatly.

Field-effect transistors are high-input-impedance devices. Fet mixers require about 2 volts rms across 50 ohms, which is equivalent to the injection required by a high-level double-balanced mixer. Because of the difficulty in providing wideband impedance matching in fets at high impedances, in my opinion medium- to high-level double-balanced mixers offer an advantage over fet mixers.

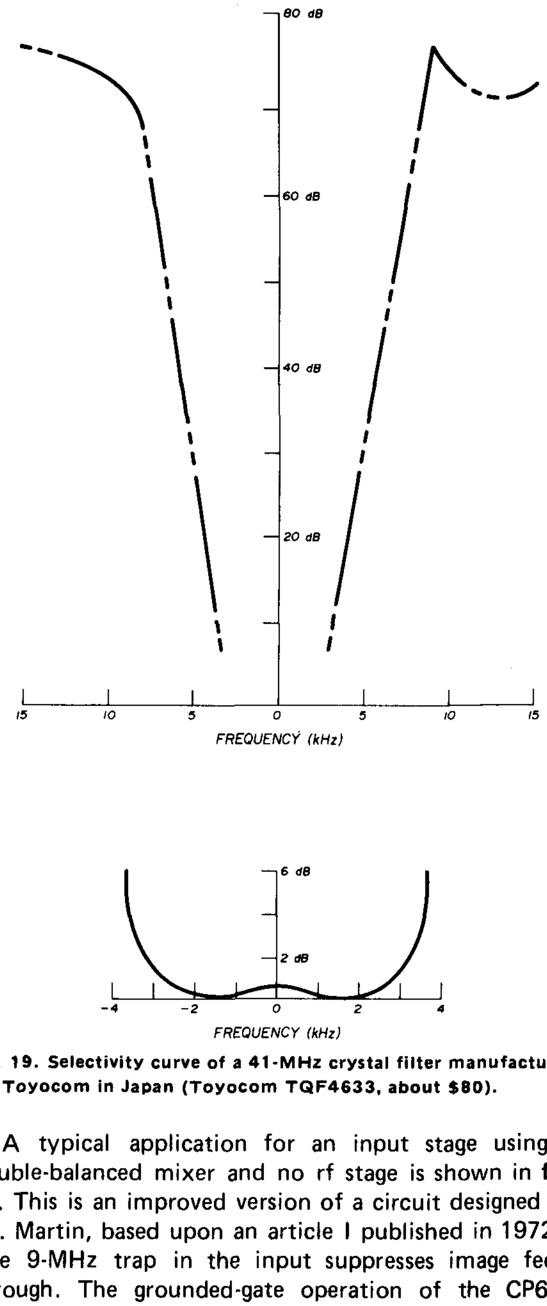


fig. 19. Selectivity curve of a 41-MHz crystal filter manufactured by Toyocom in Japan (Toyocom TQF4633, about $80).

A typical application for an input stage using a double-balanced mixer and no rf stage is shown in **fig. 18.** This is an improved version of a circuit designed by Mr. Martin, based upon an article I published in 1972.[4] The 9-MHz trap in the input suppresses image feed-through. The grounded-gate operation of the CP643 transistors (Crystalonics) provides a wideband resistive input termination for the mixer. The magnitude of the input impedance to the CP643s can be set by adjusting the 250-ohm potentiometers.

The 2N5109 transistor amplifier provides about 20 dBm injection for the RAY3 high-level double-balanced mixer (Mini-Circuits Laboratory). This circuit, which is similar to that found in the Atlas transceiver, has a distinct advantage: because of proper termination of the double-balanced mixer, unwanted intermodulation distortion products are at least 15 dB below those found in the Atlas circuit. The reason for this is that the Atlas circuit uses a tuned circuit between the crystal filter and the first i-f transistor (2N3866), and this tuned circuit does not provide proper matching to attenuate spurious products. All manufacturers of double-balanced mixers specifically require proper resistive termination for optimum performance.

The KVG XF9B crystal filter is a popular ssb filter familiar to most amateurs. The first i-f stage after the crystal filter shown here provides enough gain and agc action for most applications.

The RAY3 double-balanced mixer requires a 50-ohm termination, which is obtained by an rf transformer, model T4-1.* Using a push-pull arrangement not only increases total dynamic performance by 3 dB but also provides the possibility of using the transistors as a switch for noise blanking. For high-efficiency noise blanking a switch time of 10 microseconds is essential. A similar circuit, which is extremely efficient, is found in the Swan 200 series of amateur equipment.

To reduce costs, or if the CP643 is difficult to obtain, two BF246C transistors, made by Texas Instruments, may be substituted. These are high-current vhf fets with an I_{DSS} of about 100 mA. Since the optimum load for this push-pull amplifier is 220 ohms and a crystal filter requires 500 ohms, a 2:1 transformer, model T2-1 (made by Mini-Circuits Lab), must be used.

The 220-ohm resistor reduces the high-impedance characteristic effect of the filter as described in reference 3. The output termination of the crystal filter is provided by the tuned circuit, which is heavily damped by the 2.7 kilohm resistor. This is necessary to provide stable operation.

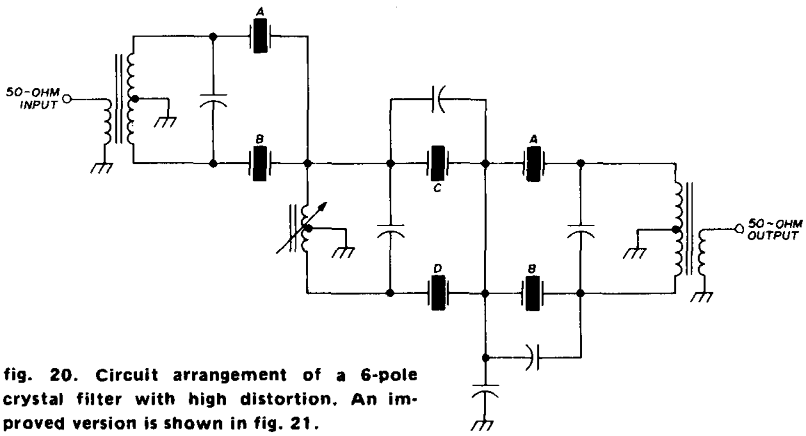


fig. 20. Circuit arrangement of a 6-pole crystal filter with high distortion. An improved version is shown in fig. 21.

The field-effect transistor provides 60 dB gain variation, and the overall gain from input to output at 50 ohms is 28 dB. In cases where a tuned circuit is used as a first i-f amplifier between the double-balanced mixer and an input stage, the dynamic range of the mixer will be heavily degraded. This is one of the reasons why the Atlas circuit does not provide the dynamic range that would be expected from the mixer alone. I believe this circuit is a good suggestion for experimenters who are still willing to build their own communications receivers.

Vhf crystal filters. Until recently low-loss 6-pole crystal filters were not available. The most significant feature other than the shape factor of these vhf filters is the internal loss, which must be held below 5 dB (typical 4.5 dB) for narrow-band applications, i.e., ±3.5 kHz. **Fig. 19** shows the typical response of a 41- or 49-MHz crystal filter.

When it was discovered that mechanical filters with magnetostrictive transducers create heavy intermodulation distortion at high input voltages, these filters were updated and now use piezo-electric transducers, which avoid this distortion. A similar effect can be observed at

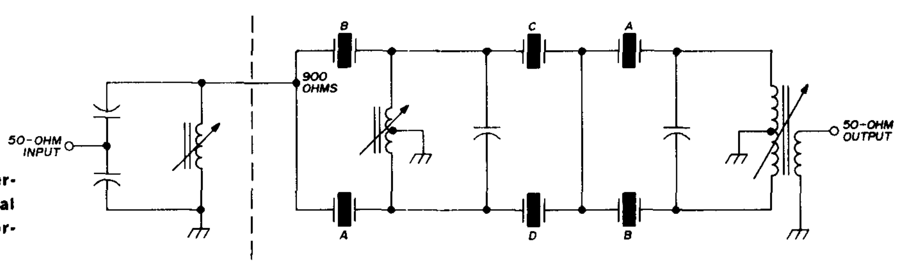


fig. 21. Improved version of a 6-pole crystal filter with low distortion.

*The RAY3 mixer and T4-1 transformer are made by Mini-Circuits Laboratory, 837-843 Utica Avenue, Brooklyn, New York 11203.

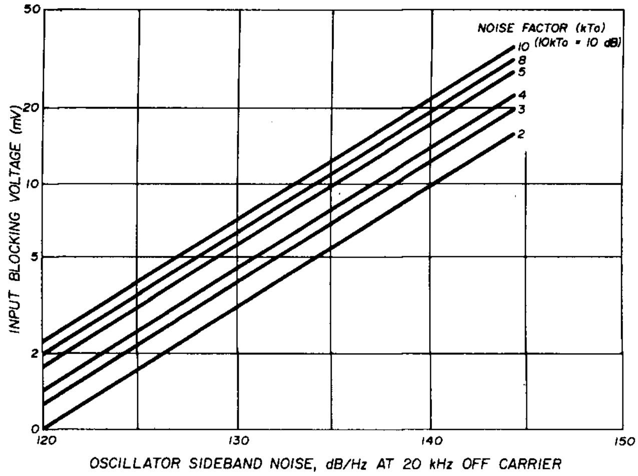


fig. 22. Receiver blocking effect caused by oscillator sideband noise. Blocking can be improved dramatically by ensuring that oscillator sideband noise is suppressed by careful circuit design.

the input transformer of crystal filters using toroids. Two 1-volt input emf signals at 50 ohms must not produce third- and second-order intermodulation distortion products of 85 dB or less. **Fig. 20** shows a typical configuration of a 6-pole vhf crystal filter with high distortion because of the input toroidal transformer, while **fig. 21** shows an improved version using a tuned input transformer with almost no measurable distortion.

Oscillator sideband noise. As explained earlier, oscillator sideband noise contributes significantly to the large-signal-handling capability of a receiver. Blocking and 3-dB compression are often confused in the literature. Compression is the effect caused by sideband noise, i.e., 20 kHz off the vfo carrier frequency, which causes blocking. This blocking effect can be expressed in terms of the sideband noise as it affects receiver sensitivity:

> Assume sideband noise of 145 dB per hertz located 20 kHz from the vfo carrier frequency and a receiver noise figure of 10 dB. An input signal of 50 mV will cause 3-dB blocking or desensitization of the receiver, while 3-dB compression may occur at a point as high as 1 volt on the input signal. This relationship is shown in **fig. 22.**

When designing oscillators with very low sideband noise, either selected fets operating in a saturated mode or medium-power 2N3866 transistors should be used.

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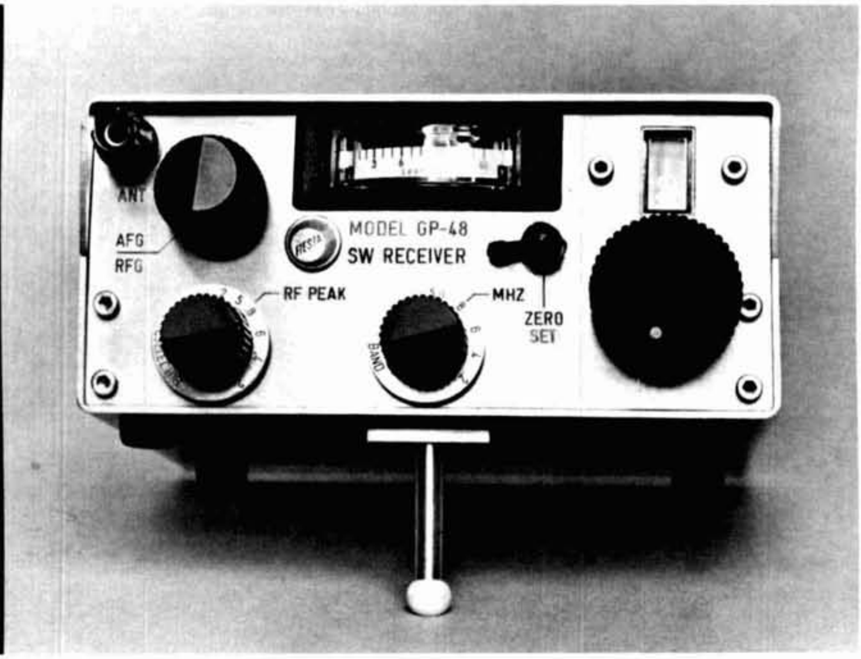

double-conversion hf receiver

with mechanical frequency readout

Although designed for shortwave broadcast, this receiver is easily modified to provide coverage between 3.2 and 30 MHz

Previous articles have described my ideas on monoband versus multiband receivers.[1,2] The concepts developed in these articles are still true, but response to another article[3] indicates many experimenters aren't interested in building a one-band receiver and are willing to accept spurious responses in a multiband unit. This is particularly true for amateur-band operation where it's a simple matter to change frequency if a contact is made where a spurious response exists in the receiver. My operation is largely confined to listening to shortwave broadcast. Tuning a signal on these bands is a different situation. Shortwave broadcast stations emit signals on one frequency. If one of these signals happens to fall on a spurious response in the receiver, reception is difficult and a rare DX station may be lost.

This article presents the results of my experiments to produce a miniaturized, solid-state receiver with a mechanical counter digital frequency readout. The receiver covers 4.2 - 5.2 MHz in five bands, but with modification to front-end inductances and suitable crystal selection, it can provide general coverage between 3.2 and 30 MHz.

conversion method

An extensive search was made of commercial and military communications gear. Some mixing combinations, excellent as they are, were discarded because of their complexity. This was particularily true of the Collins 51J4, whose mechanical complexity is high and for the Racal 217 whose electrical complexity is high. Other combinations were also discarded. The Collins 75A4, for instance, would be difficult to duplicate because of the complexity required to build a linear vfo over 1 MHz in the 2-MHz region.

The Collins 75S3 series was screened in depth. I operated a 75S3 inside and outside the amateur bands for several years and found only one spurious response

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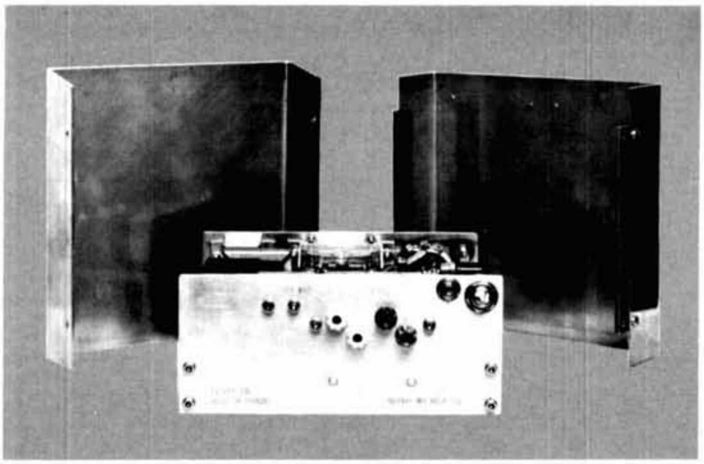
at 5000 MHz caused by the second harmonic of the vfo when tuned to 2500 MHz. This occurred when 5 MHz was tuned in the 4.8-5.0 MHz band; by switching to the 5.0-5.2 MHz band, 5000 MHz was received with no problem.

Among the advantages a duplication of the Collins 75S3 would offer is its straightforward construction. And, even more important by amateur standards, the vfo tunes only 200 kHz making it easy to linearize. The price you have to pay for these advantages, however, is that if the vfo frequency excursion is only 200 kHz, each high-frequency crystal provides coverage for a band 200-kHz wide. In other words, if you wish to use the receiver over a wide frequency range, the quantity of crystals required is somewhat high. The model GP48



Rear view showning stainless-steel cabinet construction details together with the back-panel 110-Vac input (white jacks) and 6-Vdc input (dark jacks). Two phone jacks in parallel allow headphone operation with standard and miniature plugs. This receiver required about $150 worth of parts and 150-200 hours construction time.

receiver described here has only five bands and therefore five crystals; earlier models had wider coverage: the GP45 had ten bands and the GP46 36 bands. At about $4 per crystal, the GP46 had about $150 worth of crystals.

I built some solid-state receivers around the Collins 75S3 mixer combination. I was quite pleased with these experimental efforts; then I made another unit (dubbed GP42). The GP48 is the sixth of a series. Its schematic appears in **fig. 2**. This receiver covers 4.2-5.2 MHz or, generally speaking, the 60-meter shortwave broadcast band. With suitable crystal selection and front-end coil inductances, any frequency between 3.2 and 30 MHz can be covered with the exception of 5.2-6.5 MHz.

circuit considerations

The basic circuit is closely related to the receivers in references 1 and 3, whereas the basic mixing scheme is derived from the Collins 75S3 line. There are, however, some differences worthy of mention. The Collins front end is slug tuned, with parallel capacitors switched in to

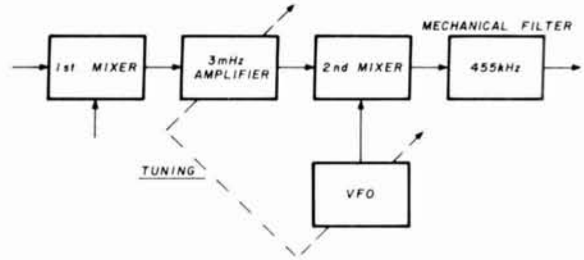


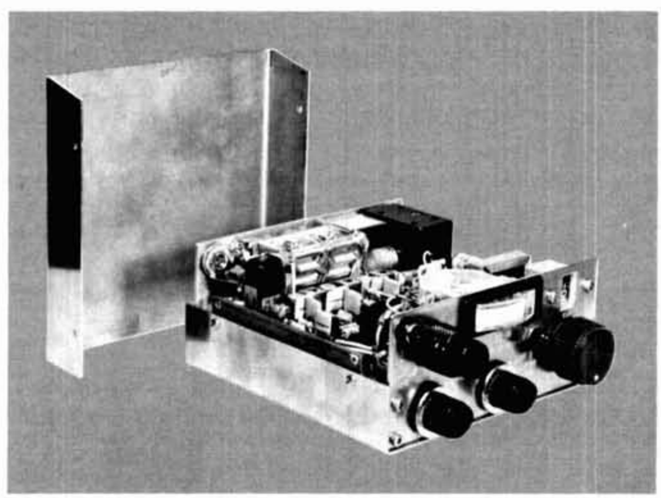
fig. 1. Block diagram or receiver 3.MHz i-f strip showing tuning method.

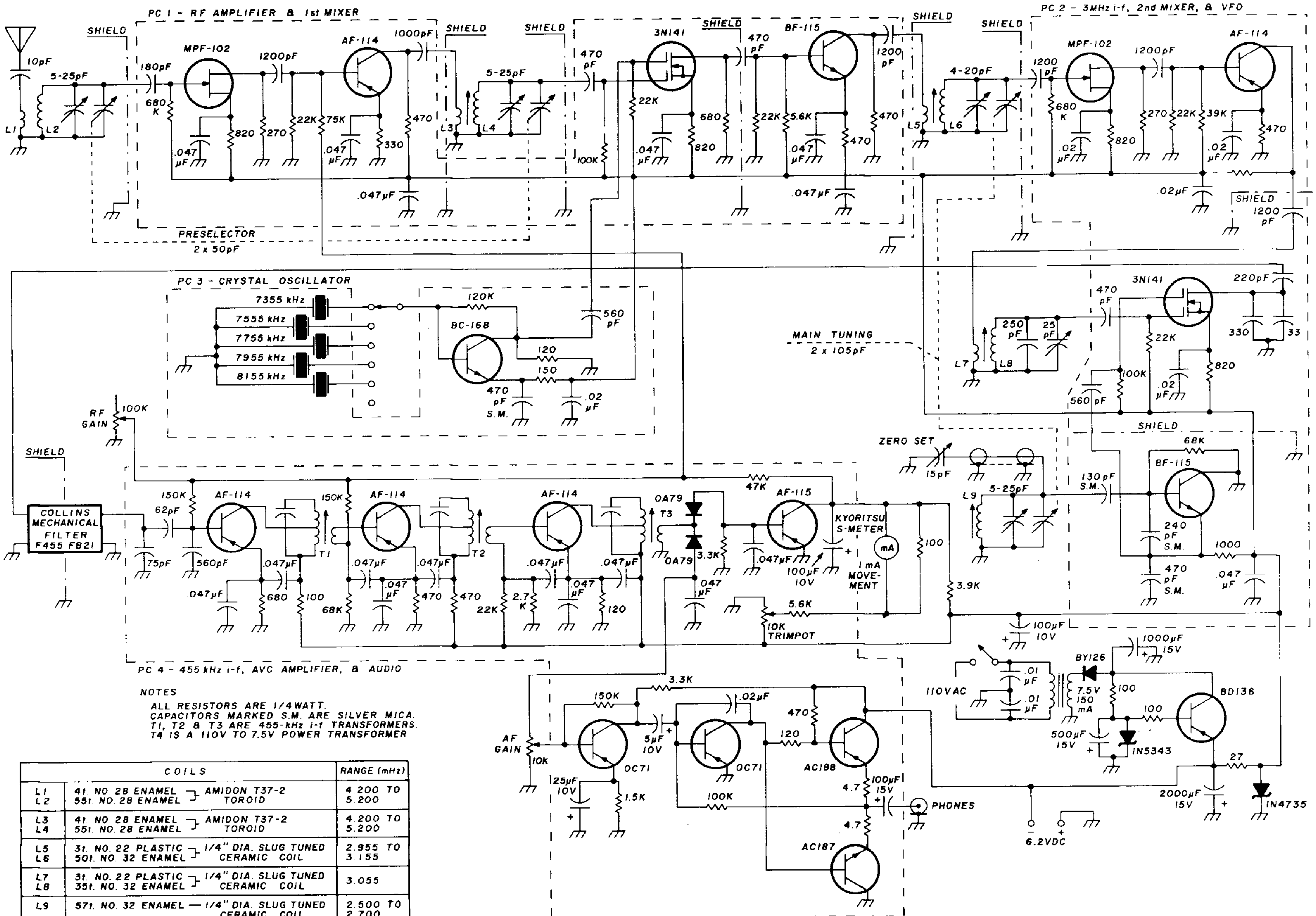
cover the low-frequency bands. While I've built similar units, I preferred the usual variable capacitor combination for this project, as most builders would probably not be able to homebrew a backlash-free multicore mechanism. For those who prefer the permeability tuned version, some surplus outlets offer powdered-iron cores with high length-to-diameter ratios and with a spring (rather than the usual threaded pin) termination. The permeability, μ, of the cores is unknown, but it could be found with a grid dip meter or by breadboarding the front end before final assembly.

The Collins 3-MHz i-f is fixed tuned with a remarkably flat response over the 2955-3155 kHz range. I didn't wish to compromise on sensitivity and, after some experimenting, I decided to make the 3-MHz i-f partially tunable (**fig. 1**). The vfo capacitor is a two-gang affair that tunes the vfo and the input to the 3-MHz i-f amplifier, whereas the second mixer is fixed tuned. This improves the bandpass flatness and, by having a fixed-tuned mixer, buffer stages aren't needed between the vfo and the mixer. In fact, since the mixer is fixed tuned and operates over a narrow frequency range (200 kHz), it offers a reasonably constant load to the vfo.

This brings up another difference between my design and that of the Collins 75S3 line; the vfo is capacitance-tuned in my case. I have already published an article on

View with upper half of cabinet removed. Front panel controls are (from left) antenna input jack, af/rf gain control, preselector tuning control, band switch, zero-frequency set control, and main-tuning control.



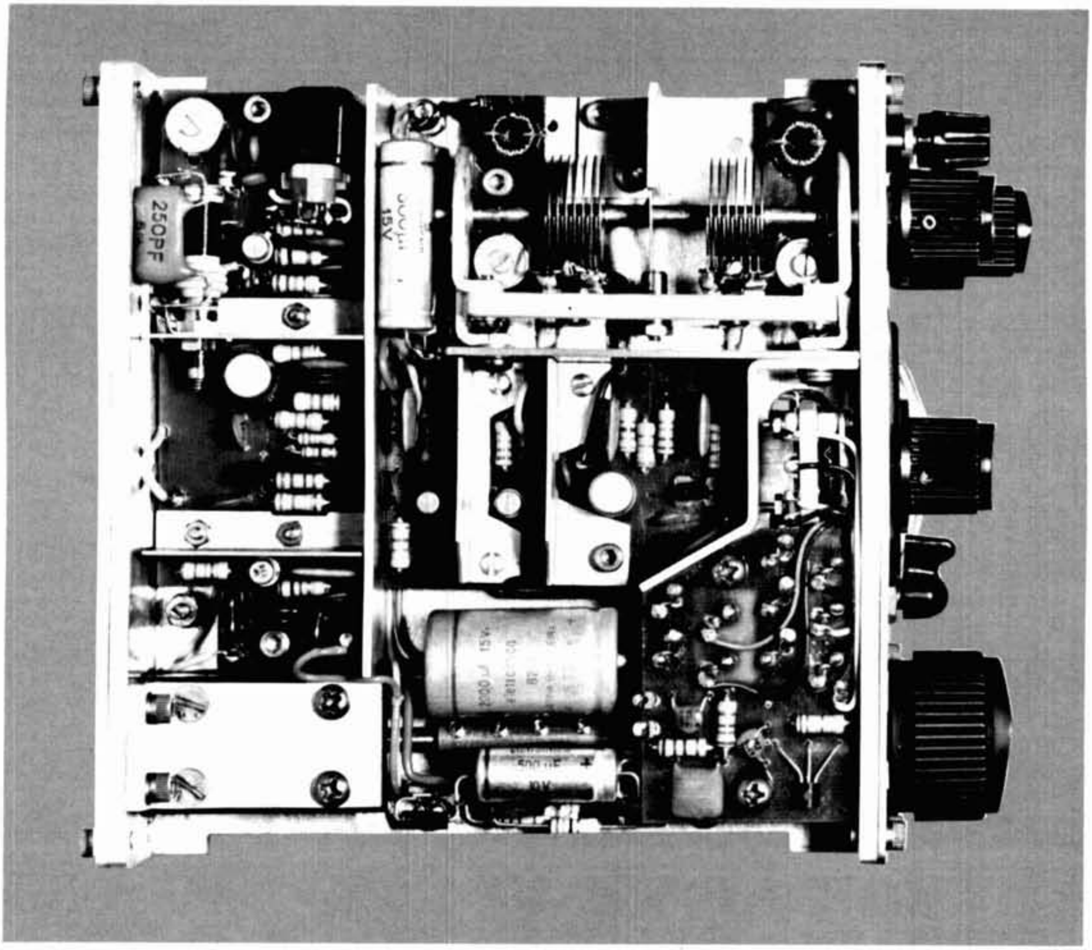


NOTES

ALL RESISTORS ARE 1/4 WATT.
CAPACITORS MARKED S.M. ARE SILVER MICA.
T1, T2 & T3 ARE 455-kHz i-f TRANSFORMERS.
T4 IS A 110V TO 7.5V POWER TRANSFORMER

COILS		RANGE (mHz)
L1 L2	4t. NO. 28 ENAMEL 55t. NO. 28 ENAMEL — AMIDON T37-2 TOROID	4.200 TO 5.200
L3 L4	4t. NO. 28 ENAMEL 55t. NO. 28 ENAMEL — AMIDON T37-2 TOROID	4.200 TO 5.200
L5 L6	3t. NO. 22 PLASTIC 50t. NO. 32 ENAMEL — 1/4" DIA. SLUG TUNED CERAMIC COIL	2.955 TO 3.155
L7 L8	3t. NO. 22 PLASTIC 35t. NO. 32 ENAMEL — 1/4" DIA. SLUG TUNED CERAMIC COIL	3.055
L9	57t. NO. 32 ENAMEL — 1/4" DIA. SLUG TUNED CERAMIC COIL	2.500 TO 2.700

fig. 2. Receiver schematic. Crystal frequencies shown are for receiver coverage between 4.2-5.2 MHz. Receiver range may be changed to include 3.2-30 MHz by substitution of suitable crystals and front-end inductances.



Bottom view. Mechanical filter is at top left next to the mixer coil and its circuit. The 3-MHz two-stage i-f amplifier is at left center; the vfo circuit is at the bottom next to the gear reducer. At top right is the two-gang front-end variable capacitor with its toroids and trimmers. The rf and mixer circuits are at center; the band switch and crystal oscillator are at right (crystals are under the PC board). At bottom center is the regulated power supply.

a permeability tuned vfo[4] as well as some generalized considerations on linear vfos.[5] Again, the average amateur will find if difficult to come up with a backlash-free permeability-tuned vfo and even more difficult, as in this case, to gang two slugs together.

construction details

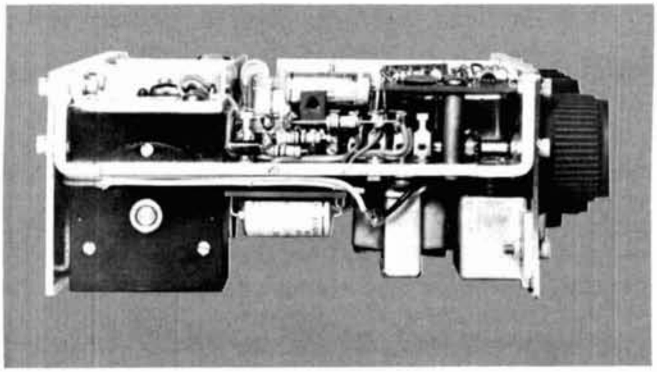
The photos show different views of the GP48; most of the construction details can be seen in these pictures. The receiver dimensions are 6 inches wide by 2.5 inches high by 6 inches deep (15 by 6 by 15 cm). All front and back panel lettering was made with a pantograph; knob skirts were made from 16-gauge stainless steel and were similarly engraved. Aluminum shields are 1/32 inch (0.8mm) for PC-board mounting and 1/16 inch (1.6mm) for chassis mounting. Extensive shielding and capacitive decoupling assure stable operation at high gain. High-quality components are used throughout.

The gear-reduction unit, which has a 100:1 ratio, is permanently lubricated in a sealed container. It's a British import made by Muffett, Ltd. The counter, made by Veeder Root, has three vertical digits. A vertical counter eliminates parallax problems when the receiver is tilted upward. The receiver tuning rate is 10 kHz per knob revolution, which makes for exceptionally smooth action. The brass bevel gears for the counter, made by Boston Gears, have a 1:1 ratio.

The vfo double-gang variable capacitor is also a British import, available in the U.S. from the J.W. Miller Company. The capacitor has very low torque, with ball bearings at both ends. It is coupled by a flexible (bellows) joint to the worm gear reducer. The S-meter is a Japanese import with a 1-mA movement. The meter face was replaced with one reading in S-units.

The receiver cabinet, which is in two halves, is made of 18-gauge stainless steel, secured by 1/8 by 3/4 by 4½ inch (3 by 19 by 114 mm) spacer bars and four 4-40 (M3) binder-head screws. Four glass-epoxy PC boards are used. The i-f/af board is mounted on top of the chassis, while the vfo/3-MHz i-f/second-mixer, together with the front-end rf amplifier/first mixer and crystal oscillator/crystal holder, are mounted below chassis.

The vfo circuit was derived from previous receivers and its circuit adapted to cover the 2500-2700 kHz



Right side. Gear reducer is at left; regulated power supply at center top. At top right is the crystal oscillator PC board with the crystals barely visible below. The power transformer is at the bottom right next to the mechanical counter. Both front and back panels are fastened with 6-32 (M 3/5) Allen-head stainless steel screws.

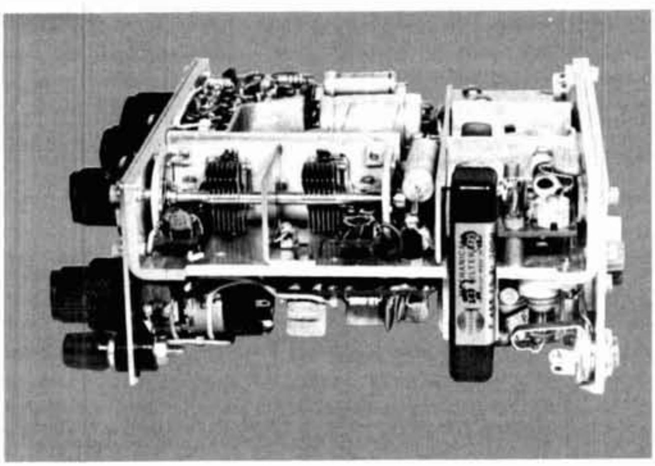
range; the variable capacitor (105 pF) is correct for the frequency excursion in question. The audio strip ends with a complementary pair wired to avoid transformers. Output power exceeds 0.5 W – more than enough for headphones. The power supply is electronically regulated providing, through back panel jacks, the dc power to run other gear, or conversely, to be used to feed the receiver from a dc source.

circuit details

Because of the extremely high gain of this set, proper shielding and bypassing must be used to achieve stability. All tuned circuits are shielded from the circuits of the corresponding active devices. Heavy filtering is used on all power lines. The vfo coil is shielded to minimize proximity effect. The vfo capacitor is not shielded because it must be accessible to allow for easy filing of its plates during the linearization procedure (described below). It must be remembered, however, that during the filing process, a small temporary shield must be fastened to the top of the vfo variable capacitor to simulate the effect of the metal receiver cabinet. Failure to do so will result in degrading vfo linearity because of the proximity of the cabinet when set in place.

It is important to understand that when a vfo is coupled to a counter, a simple change such as moving a wiring harness can degrade the calibration by more than one kHz at the band edge. Therefore, before starting to linearize the vfo capacitor all screws, nuts and components in general must be securely tightened in place and no circuit changes should be made during or after linearization. The easiest way to achieve perfect linearity is to file down the capacitor plates to within 1 kHz or so of the nominal frequency desired; from then on, the fine part of the linearization procedure is to slightly bend and adjust the side plates of the capacitor rotor. This procedure has the advantage of being reversible whereas filing is not.

Left side. Front-end double-gang variable capacitor, toroids, and trimmers are at upper left, installed on small printed circuit boards.



selectivity improvements

A major difference between amateur-band operation and shortwave broadcast listening is in selectivity. In the first case interference may be generally avoided by changing frequency; obviously this isn't possible when receiving shortwave broadcast stations. The minimum bandwidth for a-m reception is stated in the literature as 3 kHz or thereabouts. Any serious listener knows this number is way too high. I've used a 2.1-kHz bandwidth (through a mechanical filter) for many years but have changed to narrower bandwidths as conditions demand. For serious shortwave broadcast listening, I'd say the ideal receiver bandwidth is 1.8 kHz.

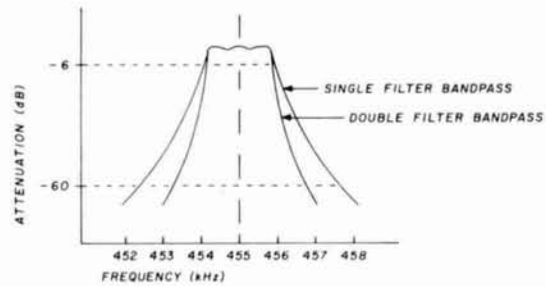
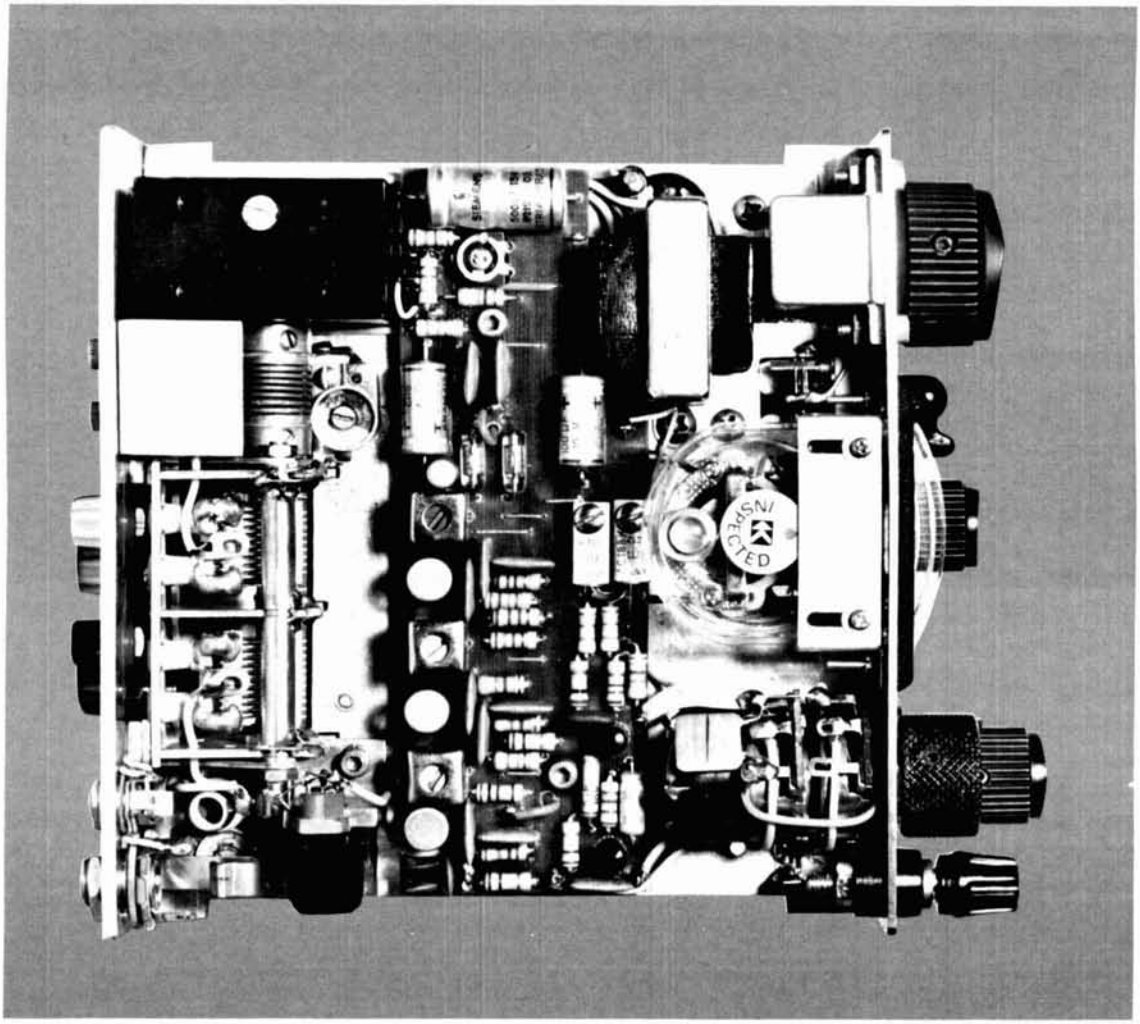


fig. 3. Skirt response of single and double filter arrangements. When cascaded, each filter must have an identical midband frequency for best results.

While mechanical filters are available with a skirt ratio of 2:1 (-60 to -6 dB), a nearby station can still cause interference because of the bell-shaped response of these filters. A remarkable improvement can be achieved by cascading two identical filters, but some words of caution are in order. Using this method, the skirt response at -60 dB can be improved by a factor of $\sqrt{2}$, or about 1.4 times, while maintaining the response constant at -6 dB (see **fig. 3**). A definite improvement in obtained; so much in fact that comparing two receivers tuned to the same station, one with a 1.4-kHz mechanical filter and the other with two cascaded 2.1-kHz mechanical filters (both bandpass figures at -6 dB), one's reaction is that the second receiver is more selective than the first.

To realize the full potential of this combination it's



Top view. Gear reducer is at top left. One of the 3-MHz i-f coils is visible below the variable capacitor. PC board in center includes the three i-f amplifiers, avc amplifier, and audio strip. At top right are the power transformer, mechanical counter, S-meter, rf/af control, and antenna input jack. Aluminum bracket, top left, shields the vfo coil to minimize proximity effect when the steel cabinet is installed.

imperative to use two filters of identical midband frequency to avoid stagger tuning with consequent skirt degradation. Each filter must be individually shielded to avoid ground loops and to ensure the signal travels through the filters and not around them. You can't take excessive precautions in this respect, because such arrangements are critical beyond imagination. The GP46 receiver (mentioned earlier) was based on this concept with entirely satisfactory results. Another i-f stage should be added to compensate for the insertion loss of the second filter.

frequency readout

While an electronic frequency readout could be used, a mechanical counter is less expensive and physically smaller. For portable work, the mechanical counter is also better since it requires no power. I believe that the vfo linearization work, however tedious and delicate, is still more advantageous than building an electronic counter. Such a counter, however, could always be added to the receiver as an external unit. The advantage in this case is that the counter can be used for other projects. I've found that an external counter offers less interference than built-in units, as their 100-kHz clocks tend to show some leakage into the receiver circuits.

acknowledgement

I'd like to thank PY2GP for his continued support in the realization of this and other projects.

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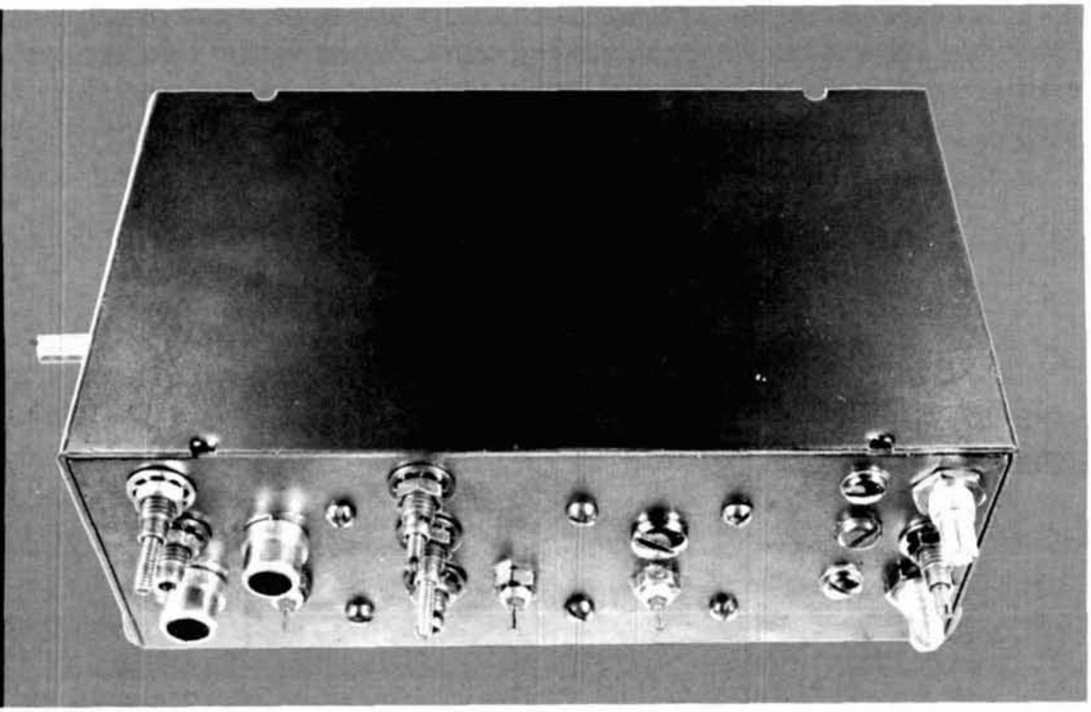

multiband high-frequency converter

A VVC-tuned converter that extends your 80-meter receiver tuning range to include 40-10 meters plus WWV

Many receiver designs have been published for the 80-meter band. This article provides complete design and construction details for a multiple-band converter to extend the range of this type of receiver. This design provides conversion of the hf bands and 10-MHz WWV signals to 3.4-4.0 MHz for i-f amplification and detection.

Fig. 1 illustrates the converter design in which mos field-effect transistors are used in the rf amplifier and mixer stages because of their superior spurious-response rejection and signal-handling capability. The dual-gate protected mosfet features rf gain or agc control in the rf-amplifier (Q1) stage and outstanding mixing characteristics for conversion to the i-f output (Q2).

Overall gain of the converter is shown in **table 1.** Coil Q for each stage is fairly high; as a result bandwidths are narrow, so some method of tuning is necessary to cover each band. Tuning is by variable capacitance (VVC) diodes in each of the rf and mixer tuned circuits.

circuit description

Fig. 2 is a schematic of the rf and mixer stages. Selection of each band is by band switching the appro-

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priate tuned circuit in the rf and mixer input stages. The mixer output uses a VVC that tunes a full MHz. The individual tuned circuit design centers are for the middle of each amateur band. The VVC will allow the coils to cover each band with a margin of several hundred kHz at

end. Both the rf and mixer stages depend on source resistance for gate bias. The source bias in Q2 is extremely important since it establishes the transfer curve linearity that provides an optimum combination of mixing and spurious response rejection (more informa-

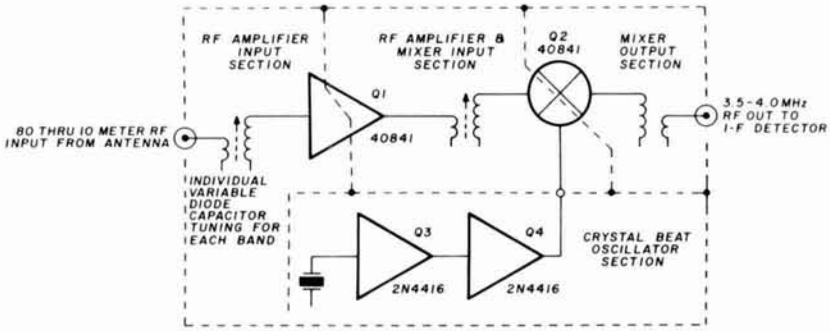
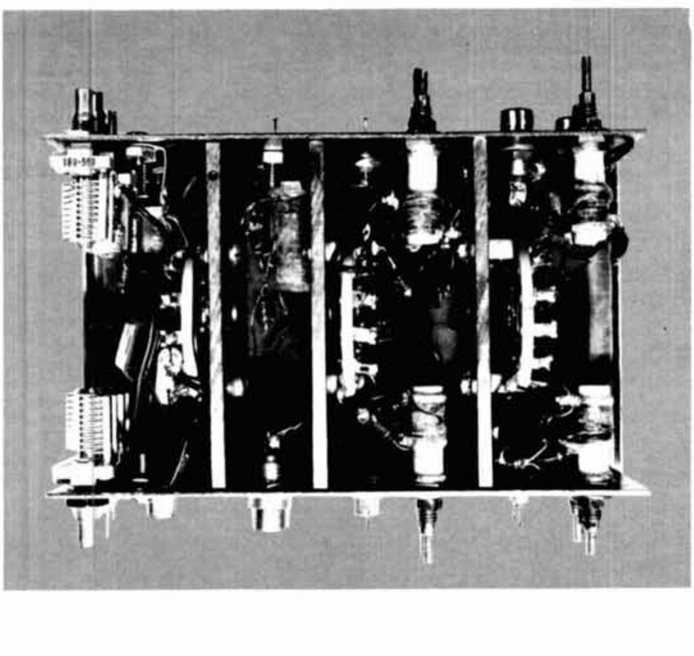


fig. 1. Converter block diagram. Design covers 80-10 meters and 10 MHz for WWV.

the band edges. The unloaded Q of the mixer output is approximately 150; the bandwidth is quite sharp and will require simultaneous tuning with the rf stages.

The total converter bandwidth is difficult to specify because of the variety of interpretations. The half-power bandwidth – that is, the width for 50% decrease in converted signal amplitude – was measured at less than 200 kHz on all bands. This sharp selectivity also indicates that for a typical 500-kHz band-segment width, several peaking adjustments are required from end to

Converter interior. Shielded partitions contain (l to r) oscillator, mixer output, rf amplifier output, and rf amplifier input sections.



tion about this source bias and mixer linearity is given in the mixer discussion that follows).

Fig. 3 is the high frequency beat oscillator schematic. This circuit is an adaptation of those in references 1 and 2. A conventional LC network is illustrated for 10-meter

table 1. Performance summary

band	tuning range (MHz)	i-f output (MHz)	converter gain (dB)
80M	3.5-4.0	3.5-4.0	0
40M	7-7.3	3.7-4.0	37
20M	14-14.25	3.625-3.875	41.7
15M	21-21.45	3.5-3.95	36.5
10M	28.5-30 (any 500 kHz segment)	3.5-4.0	34.5
WWV	9.9-10.1	3.65-3.85	28.9

Sensitivity: <.15V rms using the i-f system described in reference 3 on all bands.
Bandwidth: ±100 kHz for 50% decrease in signal amplitude without peaking adjustment.
Spurious signal rejection: ≅50 dB attenuation at ±1 MHz.

beat-oscillator conversion. This signal could have been generated by a crystal as for the other bands; however, fundamental crystals above 20 MHz are expensive and restrict the circuit to some fixed segment of the 10-meter band. Down conversion is accomplished for all bands using a beat oscillator signal approximately 3.5 MHz less than the lower edge of the receiving frequency. For 7.0-7.3 MHz conversion we use the top end of the 80-meter band, 3.7-4.0 MHz, as the i-f output to minimize beat-oscillator feedthrough in the mixer output stage. The 20-meter band (14.0-14.25 MHz) is down converted to the middle of the 80-meter i-f range to maximize circuit selectivity in subsequent stages.

Regular 32-40 pF, parallel-tuned, fundamental-

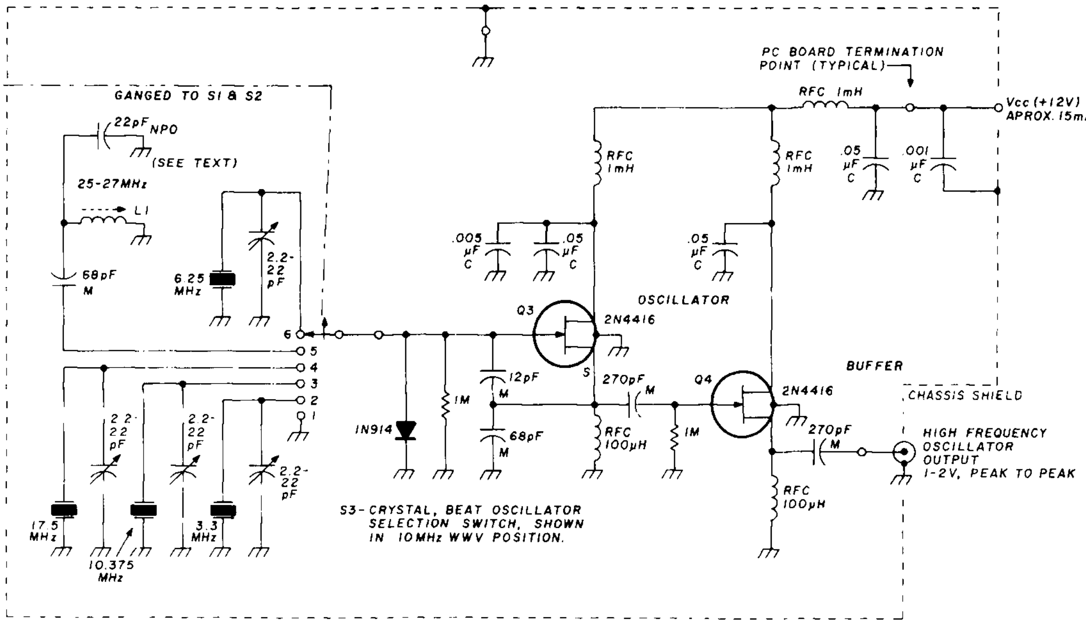


fig. 3. Oscillator schematic. L1 consists of 11 turns of AWG 26 (0.3mm) enameled wire on J.W. Miller 4500-2 form.

frequency crystals should be used. The 2.2-22 pF parallel trimmer capacitors across the crystals may be deleted if you're willing to accept some error in the converted signal output. This error in most cases will be less than ±5 kHz if the capacitors are not used. The total capacitance across the crystal is the sum of the parallel component capacitance and the gate source capacitance. Since all of these components contain variations in capacitance tolerance, some compensation or trimming is required for exact beat-oscillator frequency generation.

Where digital readout schemes are used for frequency display and consideration has been given for band-edge

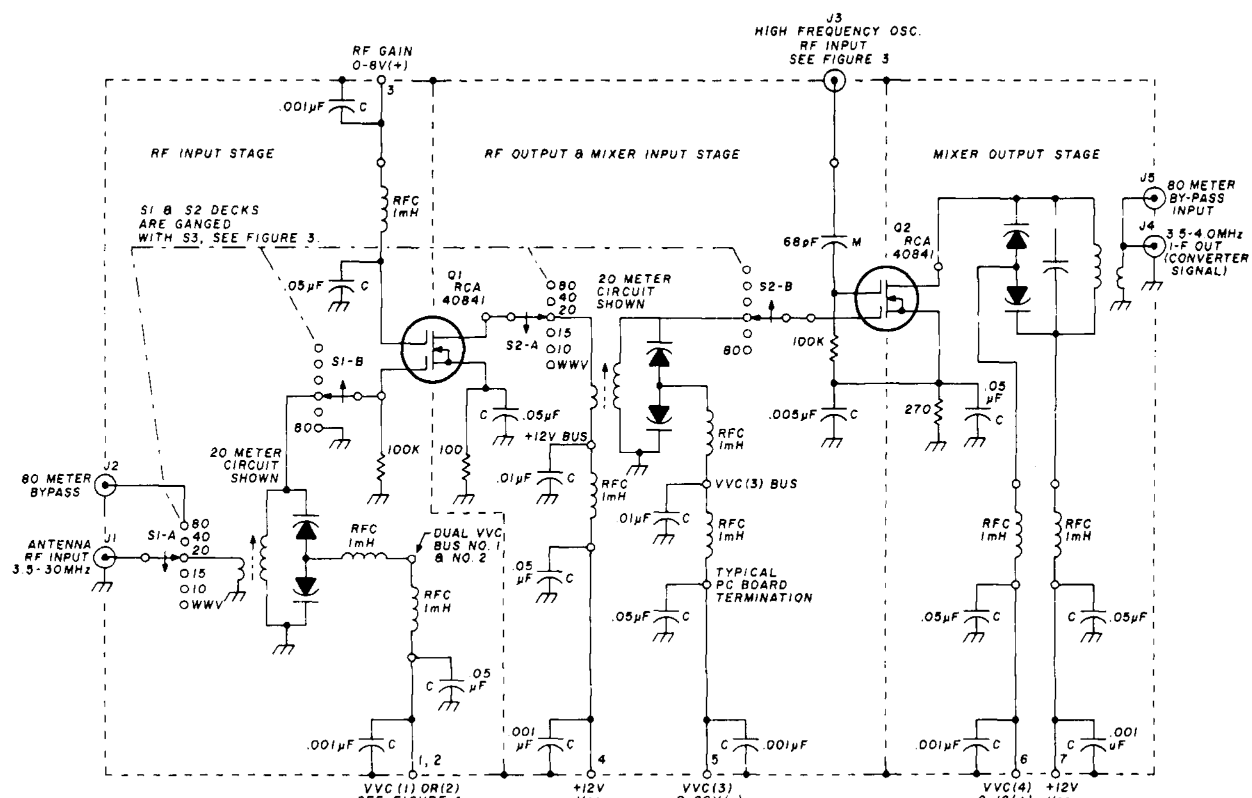


fig. 2. Rf amplifier and mixer schematic. Variable capacitance diode tuning is used in each stage.

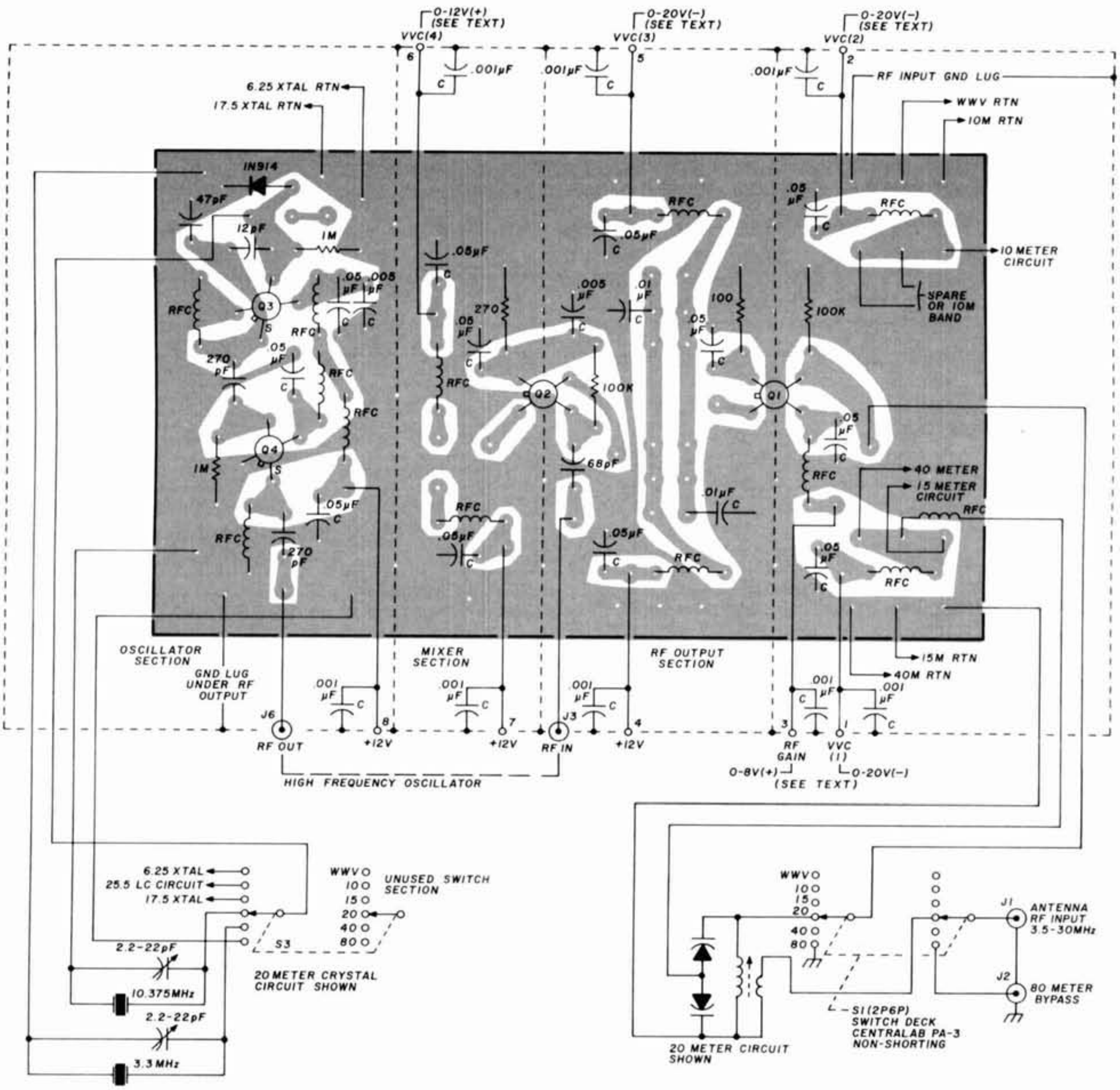


fig. 4. PC-board component layout for rf amplifier and oscillator.

determination, this frequency error is continuously known. If you depend on a slide rule dial visual display of the tuned frequency, then the oscillator crystal frequencies should be set precisely using some type of capacitor trimming similar to that indicated in the diagram. Both oscillator and buffer use a source inductive load. Considering the wide range of oscillator frequencies, this method of buffering the oscillator produces a stable, moderately uniform output level and minimizes spurious frequencies, which could fall within the tuned frequency range. In combination, the rf mixer and oscillator stages are joined to provide a comfortable 300-500 kHz segment from each of the hf bands to be converted to the 3.5-4.0 MHz 80-meter band. Because of the wide tuning capability in the rf and mixer stages, MARS and general-purpose tuning can also be added if sufficient i-f tuning range is available.

construction

From the photos it's seen that the converter is built around a simple PC board* with attached aluminum plates acting as interstage shields and a convenient mount for the switch decks. This PC board with the attached shield plates is mounted in a standard LMB 136 aluminum chassis box. The shield plates are aligned by drilling the switch shaft clearance hole through all three plates at the same time, then centering one plate on the

*Undrilled PC boards are available from the author for $3.00 including postage.

end of the box to locate and align the box and shaft clearance hole. The PC board is attached to the bottom edge of the plates using self-tapping screws. The shield plates are fastened to the sides of the box in the same manner. This construction follows the same general approach described in reference 3.

To reduce spurious noise and birdies in other portions of the receiver, generous use is made of bypass capacitors and rf chokes both at the PC board and at the point on the chassis side where control and supply voltages are applied. Each stage is individually separated and control lines decoupled from each other to minimize feedback. An examination of **figs. 4** and **5** shows potential bandswitching of six individual bands; however, only five are used in this design because of physical limitations of the box and coils. By using a slightly larger box, an additional portion of the 10-meter band could be included or perhaps a 6-meter segment.

Not everyone will be interested in the conversion of all the bands shown. For instance, if only two or three bands are desired, considerably more room would be available for access. If only one band conversion is desired, adequate room is available to mount the coil components directly to the PC-board groundplane areas. For those who may wish to experiment with a less complex and expensive system of fixed-band conversion, the capacitor resonant values of **fig. 6** may be used. Padding the coils with a 10k, ½-watt resistor will decrease coil Q sufficiently so that no rf tuning would be required over most of the 500-kHz band segment. The VVC control wouldn't be required, resulting in a further reduction in power-supply voltages.

The PC layout of **fig. 4** also indicates an unused pad next to Q3 gate. The purpose of this pad is for the installation of a coupling capacitor between the oscillator switch deck common and Q3 gate to allow for conventional LC networks to be used for all oscillator frequencies instead of crystals, similar to the method for generating high-frequency oscillator signals for 10 meters as in **fig. 3.**

oscillator

Construction should be accomplished one stage at a time. The easiest place to start is with the crystal oscillator. It's not necessary to have the shields or switch decks in place to verify crystal-oscillator operation. Mount the oscillator components to the PC board as indicated in **fig. 4.** Temporarily install any fundamental frequency crystal between 3-30 MHz between the PC pad from Q3 gate and the ground plane where holes are provided. Install a 100 ohm current-limiting resistor from the +12V PC pad and apply power. The crystal signal should appear at the source of both Q3 and at the output coupling capacitor (270 pF) from the Q4 source.

mixer

The mixer output stage should be wired next. Estimate the service lead length required from the output coil to the board, and allow the coil to hang free from its leads. The bandswitch and shield between the input and output aren't required for initial testing. Install 100-ohm current-limiting resistors to the +12V mixer coil PC pad and the VVC (4) control PC bus (0-12V). By applying a low-level 3.5-4.0 MHz signal to Q2 gate 1 and varying

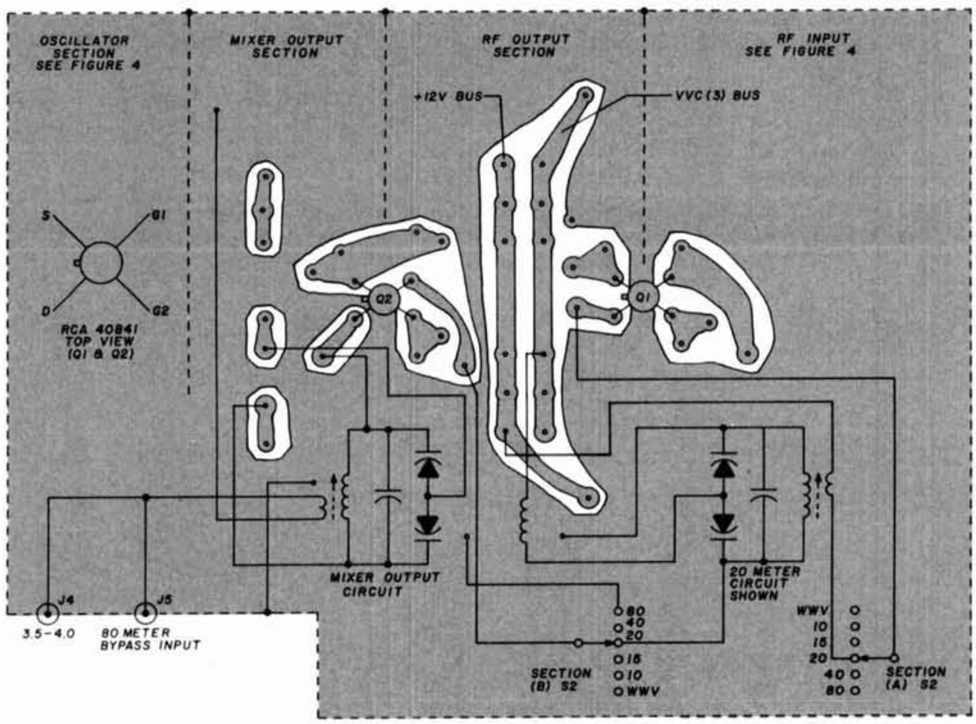


fig. 5. Rf and mixer tuned-circuit interconnections.

the VVC voltage on the output coil, signal gain and selection should be apparent. The network will be detuned because of oscilloscope internal shunt capacitance if a scope probe is not placed on the coil link output. In this mode we're using Q2 as a simple single-channel fet amplifier, tuning the drain circuit and monitoring the link. Additional verification can be obtained by temporarily grounding Q2 gate 2; the output signal should fall off immediately. To simplify these observations the signal applied to gate 1 should be modulated so that a low-frequency audio envelope can be monitored by an oscilloscope. To verify Q2 mixing characteristics and to optimize the circuit, substitute a 500-ohm potentiometer in place of Q2 source resistor. Add a jumper wire from the oscillator output PC pad (Q4 source capacitor) to the input coupling capacitor on Q2 gate 2.

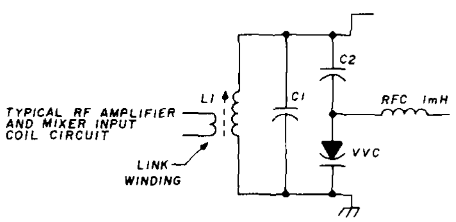
Any of the indicated crystals may be used in the oscillator; however, let's assume that the 10.375-MHz crystal is used and temporarily mounted in the PC board oscillator section as previously discussed. By applying a low-level modulated 14-MHz signal to Q2 gate 1 with the beat oscillator signal applied to gate 2, and by adjusting the VVC (4) control voltage on the mixer coil, a signal envelope similar to that shown in **fig. 7** should be seen while monitoring the mixer output coil link. Mixing linearity may be optimized by adjusting the temporarily installed 500-ohm potentiometer for both mixer gain and peak signal-to-local oscillator feedthrough ratios. Next, remove the potentiometer and measure its resistance. Select a fixed resistor of this value, which will be between 200 and 300 ohms. This resistor establishes the gate 1-to-source bias on Q2 for linearity of the fet transfer characteristics. For ideal mixing the transconductance curve should approach a straight line when the drain current is plotted against the gate 1 voltage. Since the zero gate voltage versus drain current varies over a wide margin, selection of this source resistor will optimize the device operation.

rf amplifier

The rf amplifier PC components should be installed next. As in the previous steps, no interstage shield is necessary for initial testing. From the pad on the PC board for Q1 drain circuit, temporarily install a 0.01μF capacitor to the gate 1 pad of Q2 and add a 100-ohm resistor between the Q1 drain PC pad and the +12V bus. Apply power to the rf and mixer drain circuits and oscillator, and apply a low-level modulated 14-MHz signal to Q1 gate 1. The mixer output should be similar to the previously described mixer test signal, with some high-frequency feedthrough apparent. Q1 gain control can be verified by temporarily grounding Q1 gate 2; the output signal should fall off immediately then slowly increase when gate 2 is left open. The rf and agc gain control features of a dual-gate mosfet can be observed if a variable positive voltage is applied to Q1 gate 2. The output signal should be nearly zero when gate 2 is at or near ground potential; then the output signal will rise in amplitude as the gate 2 voltage is increased to about 8 Vdc. Maximum gain is achieved when the gate 2-to-source voltage is about 6 Vdc.

You may want to adjust the Q1 source resistor. Because Q1 is a depletion device, the source could be operated at ground potential with maximum gain from the stage; however, some biasing of gate 1 is desirable for device stability, and 100 ohms is a common value for a mosfet in this type of rf application. For higher frequencies, such as 144 or 220 MHz, a 220-ohm resistor would

table 2. Rf tuned-circuit component complement



rf amplifier coils

band/ freq	L1 form(4)	Q(2)	winding(1) turns	winding(1) AWG	winding(1) (mm)	link(1) turns	link(1) AWG	link(1) (mm)	nominal cap (pF)	C1 (pF)	C2(3) (pF)	VVC type
40m(3)	4500-2	65	25	28	(0.3)	5	30	(0.25)	175	22	—	MV1666(2)
20m(3)	4500-3	80	20	28	(0.3)	5	30	(0.25)	65	—	—	MV1652(2)
15m	4500-3	65	13	26	(0.3)	4	28	(0.3)	58	—	68	MV1660
10m	4500-6	60	10	26	(0.3)	3.5	28	(0.3)	45	—	43	MV1660
10MHz	4500-2	60	23	28	(0.3)	4	30	(0.25)	82	82	—	—

Notes:
1. Turns are close wound, slightly loose over form.
2. Unloaded value.
3. C2 is a VVC, mounted anode-to-anode. Q dope all components after soldering.
4. D.W. Miller part numbers.

mixer coil

freq (MHz)	L1 form	Q(2)	winding(1) turns	winding(1) AWG	winding(1) (mm)	link(1) turns	link(1) AWG	link(1) (mm)	nominal cap (pF)	C1 (pF)	C2 (pF)
3.5/4.0	plex rod 3/8 in. (9.5mm)dia	90	48	28	(0.3)	10	30	(0.25)	185-245	150	MV1403(2) VVC mounted anode-to-anode

Notes:
1. Winding is 3/4 in. (2cm) long located along center of rod. Link is on bottom end near chassis.
2. Unloaded value.

provide some improvement in the noise figure with a sacrifice in gain. If you need the extra gain, the source resistor could be reduced to 47 ohms. If plenty of selectivity and gain are available in the i-f amplifier, my suggestion would be to increase Q1 source bias to several hundred ohms.

final adjustment

At this point we're confident of our component operation, and all temporary connections should be removed. **Figs. 4** and **5** illustrate the inductor and switch deck interconnection. The switches select the crystal beat oscillator frequency and appropriate tuning network for the desired band. In the 80-meter position the signal bypasses the rf mixer stages and is presented directly to the output connector. Using 24 AWG (0.5mm) solid hookup wire, adequate strength is available to support the coil components until you're ready to install the PC board into the box and fasten the coil to the chassis sides.

Before installing the PC board and coil assemblies into the box, the interface connections should be temporarily installed as in **fig. 7** and a low-level signal applied to each band. This will allow you to de-bug any wiring errors and initially adjust the coil slugs. A signal level of one or two millivolts may be required, depending on your oscilloscope sensitivity, for alignment. A typical modulated 14-MHz signal is shown in **fig. 6** and indicates the envelope response with proper tuning and normal gain settings. The signal amplitudes and features may differ considerably if the mixer output is left unloaded. During my initial testing, a 51-ohm, 1-watt resistor was used across the oscilloscope input terminals, and the normal input capacitance was 40 pF.

additional suggestions

VVC selection was based on what was readily available. Nominal resonance capacitance for middle of the band tuning of each coil is shown in **table 2**. An infinite number of VVC and capacitor values will tune the circuits. The total tuning capacitance of these VVC devices is not fully used for the bands indicated. The tuning ratio for the rf and mixer input VVC devices is approximately 3, and the mixer output VVC device has a tuning ratio of 10. The reader is referred to Motorola's application note[4] for design options.

Resistors for the oscillator are ¼-watt; those for the rf and mixer are ½-watt. Low-voltage ceramic or mica capacitors are used. You might experiment using 100-ohm resistors in place of rf chokes for cost reduction; however, the tradeoff here is added spurious noise.

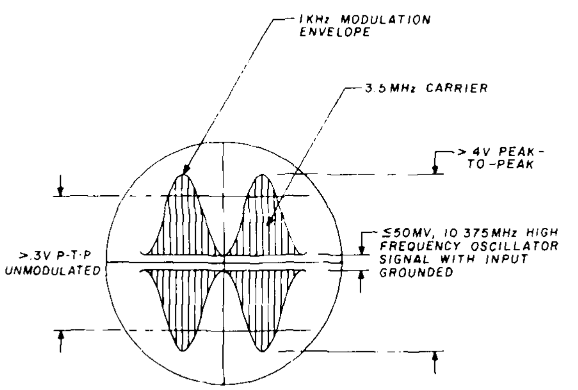


fig. 6. Typical oscilloscope patterns from mixer output with a 1000-microvolt input signal at 14 MHz and 50-ohm resistive load across the scope terminals. Gain is about 40 dB.

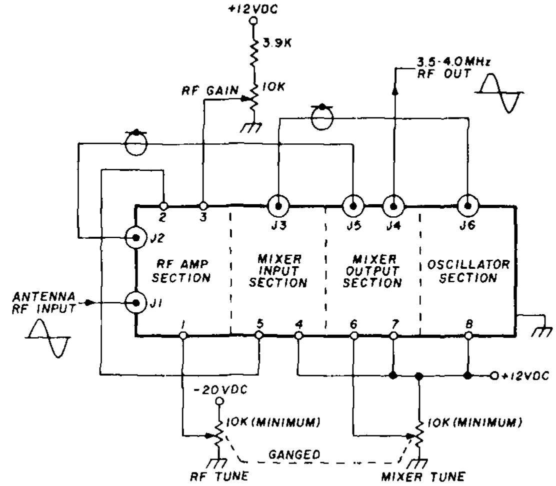


fig. 7. Application diagram. Agc control may be applied to pin 3 of the rf amplifier instead of the rf gain control (see reference 3).

application

Fig. 7 shows the typical interface for VVC control and power bus lines. The rf and mixer tuning controls may be ganged as indicated by adjusting the mixer tuning control for maximum signal output at 3.75 MHz then readjusting the coil slug to the center positions. Tracking between rf and mixer stages is not perfect because of device nonlinearity and frequency tuning ranges; however, for nominal 500 kHz bandwidths, these variations are minimal.

Reference 3 includes an agc circuit that may be used instead of the rf-amplifier gain control. Because of the high input impedance of Q1 control gate, several agc stages may be ganged in parallel; however, the device types in the rf and i-f stages should be similar because of variations in gate 2 voltage and gain between device types. For optimum receiver performance, the VVC controls should remain independent so that signal peaking will be maximum.

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2. D.L. Stoner, W6TNS, "High-Stability Crystal Oscillator," *ham radio*, October, 1974, page 36.
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ham radio

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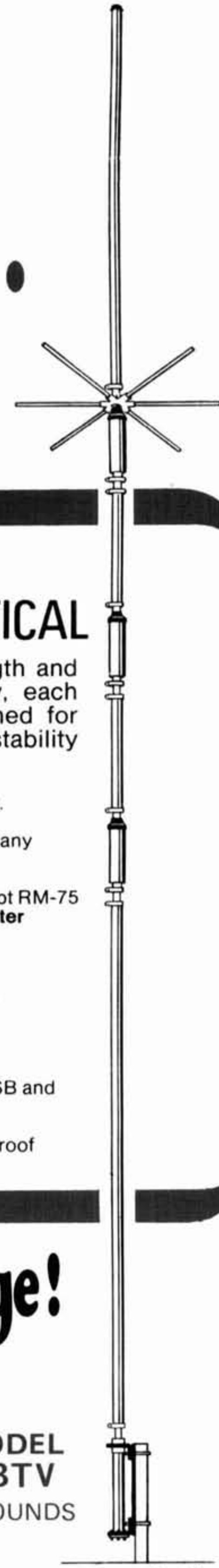

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solid-state microwave amplifier design

A complete design approach for uhf and microwave amplifiers, and the performance tradeoffs which result with simplified design methods

In two recent articles I introduced a rather simplistic design method for matching microwave transistors in microstripline preamplifiers for 1296 MHz.[1,2] The designs proved entirely acceptable for amateur applications, exhibiting input and output vswr well below 2:1, and yielding gains within 0.5 dB of those theoretically obtainable from the transistors used in the circuits. In fact, these designs were so successful that they ultimately formed the basis for a commercial product line for the 1296 MHz band. There is, however, no reason why full maximum available gain, and true 1:1 vswr, should not be achievable in microstripline amplifiers if simplistic design methods are abandoned in favor of a more complex, rigorous approach. In military and aerospace electronics, where the ultimate in performance is essential, such perfectionism has become a watchword. Small wonder, then, that a few discerning radio amateurs are attempting to achieve state-of-the-art performance in their microwave transistor designs.

This article is an exposition of state-of-the-art design techniques and is *intended to be instructional, rather* than constructional. By briefly reviewing the design approach applied in my previous articles, I shall attempt to identify the omissions which cause the resulting amplifier performance to depart from optimum. Following a brief discussion of transistor parameter characterization, I will show the mathematical procedures necessary to overcome the performance shortcomings of my previous designs. This is followed by a design example based on the MRF-901 transistor. The final section of the article outlines the minor performance differences between the rigorous design approach and more casual matching computations. For those readers who are interested, an appendix is provided which outlines some of the rules of vector arithmetic that are required for proper manipulation of semiconductor scattering parameters.

simple matching scheme

For any two-port device, optimum power transfer occurs when both input and output are terminated in their complex conjugate impedances. That is, for any port impedance consisting of a resistance and a series reactive component, the termination should appear as a like value of resistance in series with the *opposite* reactive component. For example, a port with an impedance of 35 + j100 ohms (35 ohms of resistance, 100 ohms of series *inductive* reactance), should be terminated in 35 − j100 ohms (35 ohms of resistance, 100 ohms of series *capacitive* reactance).

Recognizing the above, amplifier design consists of causing the ports (transistor input and output imped-

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ances) to be terminated in their complex conjugates. The problem is to identify the actual transistor input and output impedances occurring at a given frequency, and under a given set of bias conditions.

It is convenient to assume that the impedances related to the transistor's input and output reflection coefficients (s_{11} and s_{22}) approximate the device's input and output impedances. This is the approach I used in my previous articles. The fallacy (and one source of minor errors) is the fact that s_{11}, the *input reflection coefficient,* directly relates to input impedance *only* when the output is terminated in a pure resistance of 50 ohms. Similarly, s_{22}, *output reflection coefficient,* is directly related to the output impedance *only* when the input is terminated in a pure resistance of 50 ohms. In other words, varying the impedance match at one port will affect the impedance seen at the other port.

To understand the reason for this interaction, it's important to realize that no transistor is strictly a unilateral device. Any time a signal is injected into the output of a transistor amplifier, some signal will be discernible at the input. The physics of semiconductor construction allow for a feedback path which, though it may appear minimal, nontheless allows output matching to have an impact on input impedance, and vice versa.

If the input and output reflection coefficients of a transistor were both zero, this feedback path would have no effect on transistor matching. A reflection coefficient of zero indicates a corresponding port impedance of 50 ohms, nonreactive. Since s_{11} is measured with the output terminated in 50 ohms, and s_{22} is measured with the input similarly loaded, this is the only case in which terminating one port will not disrupt matching to the other. Of course, if both the input and output impedances of a transistor were 50 ohms, pure resistive, matching to a 50-ohm source and load would be considerably simplified. Unfortunately, we are not blessed with such transistors. Hence, to properly determine input and output impedances, the device's *transfer coefficients* must be considered.

S-parameters

In addition to s_{11} and s_{22}, the reflection coefficients discussed previously, a microwave transistor is characterized by a *forward transfer coefficient,* s_{21} (mathematically related to gain), and a *reverse transfer coefficient,* s_{12} (which describes the internal feedback path). Together, these four scattering parameters fully characterize the operation of the device. From them can be calculated the transistor's stability factor (tendency to oscillate under various conditions of source and load termination), maximum available gain, maximum stable gain, and equivalent input and output impedances. The s-parameters can be further manipulated to determine the device's maximum linear power output capability[3] although such an analysis is beyond the scope of this article.

It should be remembered that each of the four s-parameters varies with frequency, as well as with varying conditions of bias current and operating potential. The term "scattering" is derived from the fact that the parameters describe a set of variables, based on traveling waves incident on a port and reflected (or scattered) from it, which are evaluated with a mathematical tool called a scattering matrix.[4]

It should be further pointed out that s-parameters are *vectors.* That is, they appear as points on a Smith chart or polar plot which can be defined by both *magnitude* and *angle.* For example, at a frequency of 1.3 GHz, with

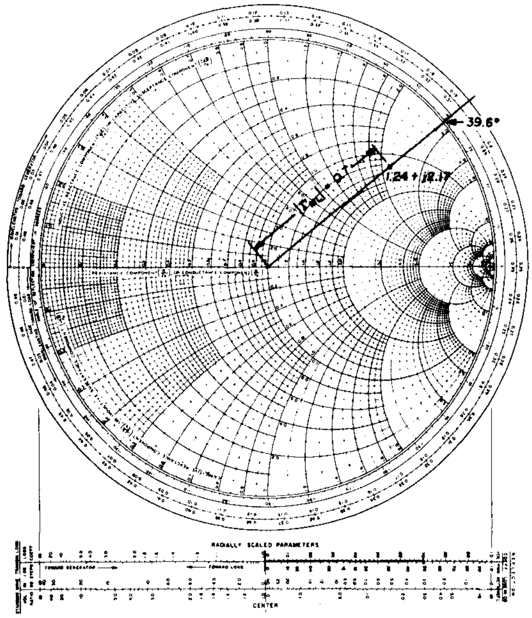

fig. 1. Using a Smith chart to plot the load impedance which exhibits the specified load reflection coefficient, 0.7 ∠ 39.6°. On this normalized Smith chart this yields 1.24 + j2.17. In a 50-ohm system the required load impedance is 62 + j108.5. To plot this point first locate the angle of the reflection coefficient on the peripheral scale and draw a line from 39.6° on this scale through the center of the chart. Referring to the radially scaled voltage reflection coefficient below the chart, measure the distance to 0.7, and transfer this length to the previously plotted line on the Smith chart. The crossover point marks the required complex load impedance.

a collector current of 10 mA and a collector-to-emitter potential of 10 volts, the common-emitter s-parameters for a Motorola MRF-901 microwave transistor are:

$$s_{11} = 0.47 \angle +161^\circ$$
$$s_{22} = 0.43 \angle -41^\circ$$
$$s_{12} = 0.08 \angle +64^\circ$$
$$s_{21} = 3.1 \angle +63^\circ$$

A complete discussion of the derivation and usefulness of the four s-parameters is available in an application note published by Hewlett-Packard.[5] Tabulations of s-parameters corresponding to various frequencies and bias conditions are available from the manufacturers of most microwave transistors.

gain and stability analysis

Before attempting to determine input and output impedances and design matching networks, it is desirable to approximate the gain capabilities of the transistor under the chosen operating conditions, and to determine whether the resulting amplifier will be stable. Three parameters which aid in such analysis are Maximum Available Gain *(MAG)*, Maximum Stable Gain *(MSG)*, and Rollett's stability factor *(K)*. K indicates the amplifier's tendency to oscillate. If K is greater than 1, the amplifier will be stable under any combination of input and output impedances or phase angles. Such an amplifier is said to be *unconditionally* stable. Conservative

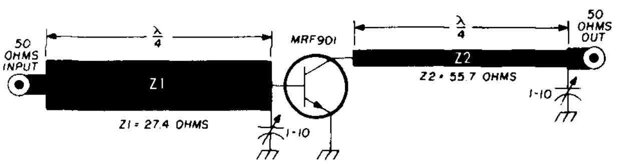

fig. 2. Basic circuit for a 1296-MHz amplifier which uses a Motorola MRF-901 transistor. Input and output matching is provided by microstriplines and 10 pF trimmer capacitors. This amplifier is unconditionally stable and gain is about 13 dB (bias networks are not shown).

design philosophy suggests that if K calculates to less than unity a different transistor or bias condition should be selected.

Maximum stable gain is, to quote WA6RDZ, ". . . the most important figure of merit. Transistors with high *MSG* are easy to match, easy to tune, and give high performance, trouble-free amplifiers."[6] Maximum available gain, also easily calculated, is a fairly accurate approximation of the gain you will observe in the actual circuit if it is carefully designed and built. If *MAG* is on the order of 2 or 3 dB less than *MSG*, the amplifier is likely to be both stable and reliable.

Of the above parameters, *MSG* is the most readily computed because it involves only the absolute values (magnitudes) of s_{21} and s_{12}:

$$MSG\,(dB) = 10 \log \frac{|s_{21}|}{|s_{12}|} \tag{1}$$

In order to perform the remaining calculations, the vector quantity *Δ and the scalar values B_2 are required:

$$\Delta = s_{11}\, s_{22} - s_{12}\, s_{21} \tag{2}$$

$$B_2 = 1 + |s_{22}|^2 - |s_{11}|^2 - |\Delta|^2 \tag{3}$$

It is now possible to calculate Rollett's stability factor, K

$$K = \frac{1 + |\Delta|^2 - |s_{11}|^2 - |s_{22}|^2}{2 \cdot |s_{21}| \cdot |s_{12}|} \tag{4}$$

If K proves greater than unity, go ahead and calculate maximum available gain:

$$MAG\,(dB) = MSG + 10 \log \left|K \pm \sqrt{K^2 - 1}\right| \tag{5}$$

where if B_2 is greater than zero (i.e., positive), the sign preceding $\sqrt{K^2 - 1}$ is negative, and if B_2 is less than zero (i.e., negative), the sign is positive. At this point in the circuit design, it is possible to determine whether the performance of the amplifier is acceptable for the intended application. If the amplifier proves only conditionally stable ($K \leq 1$), or if *MAG* is insufficient, select another transistor or bias point, and go through the calculations again with the new s-parameters.

*Rules for vector arithmetic are discussed in *appendix 1*.

output conjugate matching

Assuming that the gain and stability analysis indicate that the amplifier design is workable, the output circuit is designed to terminate the transistor in the complex conjugate of its *actual* output impedance. To determine the true output impedance requires a manipulation involving not only s_{22}, but also Δ and B_2 (**eqs. 2 and 3**) as well as s_{11}. To find the desired load reflection coefficient, first compute the intermediate vector quantity C_2:

$$C_2 = s_{22} - (\Delta \cdot s_{11}{}^*) \tag{6}$$

where the asterisk indicates that the complex conjugate of the immediately preceding vector is used (that is, same magnitude, angle has opposite sign).

The angle of the desired load reflection coefficient, $\Gamma_{ML\theta}$, is simply $C_{2\theta}{}^*$. The desired magnitude is found from:

$$|\Gamma_{ML}| = \frac{B_2 \pm \sqrt{B_2{}^2 - 4\,|C_2|^2}}{2\,|C_2|} \tag{7}$$

The sign preceding the radical sign is, once again, opposite to the sign on B_2. The desired load reflection coefficient may now be converted on a Smith chart into a complex impedance value, then matched to 50 ohms, as discussed in my previous articles.

input conjugate matching

Once the output load has been specified, the source reflection coefficient which will properly terminate the transistor's input is found from

$$\Gamma_{MS} = \left[s_{11} + \left(\frac{s_{12} \cdot s_{21} \cdot \Gamma_{ML}}{1 - (\Gamma_{ML} \cdot s_{22})}\right)\right]^* \tag{8}$$

where the asterisk indicates the complex conjugate (same magnitude, angle has opposite sign). This reflection coefficient may be plotted on a Smith chart to determine equivalent impedance, and the result transformed to 50 ohms.

To those readers who have Hewlett-Packard HP-45 engineering calculators, I highly recommend an article by Martin[7] which reduces formulas similar to the above to straightforward keystroke sequences. I have also published a HP-35 algorithm for series-to-parallel complex impedance conversion which may prove useful in designing matching networks.[8] Additionally, I recently derived a family of programs for the new HP-25 programmable calculator which greatly simplify all of the above calculations.*

*A complete set of HP-25 amplifier design programs is available from the author for $2.00 plus a stamped, self-addressed envelope.

design example

As noted previously, the common-emitter s-parameters for the Motorola MRF-901 transistor at 1.3 GHz, when biased at 10 volts and 10 mA, are as follows:

$$s_{11} = 0.47 \angle{+161^\circ}$$
$$s_{22} = 0.43 \angle{-41^\circ}$$
$$s_{12} = 0.08 \angle{+64^\circ}$$
$$s_{21} = 3.10 \angle{+63^\circ}$$

Using these parameters, the maximum stable gain, MSG, is calculated from **eq. 1**:

$$MSG\,(dB) = 10 \log \frac{s_{21}}{s_{12}}$$
$$= 10 \log (3.1/0.08) = 15.9\ dB$$

Before performing the remaining calculations, it's necessary to compute the vector quantity Δ (**eq. 2**) and the scalar quantity B_2 (**eq. 3**).

$$\Delta = (s_{11} \cdot s_{22}) - (s_{12} \cdot s_{21})$$
$$= [|s_{11}| \cdot |s_{22}| \angle(s_{11\theta} + s_{22\theta})] - [|s_{12}| \cdot |s_{21}| \angle(s_{21\theta} + s_{12\theta})]$$
$$= [|0.47| \cdot |0.43| \angle 161^\circ + (-41^\circ)] - [|0.08| \cdot |3.1| \angle(63^\circ + 64^\circ)]$$
$$= (0.202 \angle 120^\circ) - (0.25 \angle 127^\circ)$$

Converting to rectangular notation and subtracting the x and y components,

$$\Delta_x = 0.049 \qquad \Delta_y = -0.025$$

Returning to polar notation

$$\Delta_{tR} = \sqrt{\Delta_x^2 + \Delta_y^2} = \sqrt{0.00306} = 0.055$$
$$\Delta_{t\theta} = \arctan \Delta_y/\Delta_x = \arctan -0.500 = -26.56^\circ$$
$$\Delta = 0.055 \angle{-26.56^\circ}$$

The scalar quantity B_2 is calculated from the relationship

$$B_2 = 1 + |s_{22}|^2 - |s_{11}|^2 - |\Delta|^2$$
$$= 1 + (0.43)^2 - (0.47)^2 - (0.055)^2 = 0.96$$

With the vector quantity Δ and scalar quantity B_2 now known, it's possible to calculate Rollett's stability factor, K, from **eq. 4**.

$$K = \frac{1 + |\Delta|^2 - |s_{11}|^2 - |s_{22}|^2}{2\,|s_{21}| \cdot |s_{12}|}$$
$$= \frac{1 + (0.055)^2 - (0.47)^2 - (0.43)^2}{2(3.1)\,(0.08)} = 1.20$$

Since the stability factor is greater than 1, the amplifier will be unconditionally stable.

The maximum available gain, MAG, of the amplifier is calculated with **eq. 5**.

$$MAG\,(dB) = MSG\,(dB) + 10 \log |K \pm \sqrt{K^2 - 1}|$$

Since B_2 is greater than zero (i.e., positive) the sign of the radical is minus.

$$MAG\,(dB) = 15.9\ dB + 10 \log |1.20 - \sqrt{1.20^2 - 1}|$$
$$= 15.9\ dB + 10 \log 0.54$$
$$= 15.9 + (-2.7) = 13.2\ dB$$

Since MAG is approximately 3 dB lower than MSG, the amplifier can be expected to tune easily.

output matching. To terminate the transistor in the complex conjugate of its output impedance, first compute the intermediate vector quantity C_2 from **eq. 6**.

$$C_2 = s_{22} - (\Delta \cdot s_{11}{}^*)$$

remembering that the angle of $s_{11}{}^*$ has a sign opposite to that of s_{11} (in this case, $s_{11}{}^* = 0.47 \angle{-161^\circ}$

$$C_2 = (0.43 \angle{-41^\circ}) - [(0.054 \angle{-25.61^\circ}) \cdot (0.47 \angle{-161^\circ})]$$
$$= (0.43 \angle{-41^\circ}) - (0.02 \angle 173.4^\circ)$$

Converting to rectangular notation and subtracting the x and y components,

$$C_{2x} = 0.344 \qquad C_{2y} = -0.284$$

Returning to polar notation

$$C_{2R} = \sqrt{C_{2x}{}^2 + C_{2y}{}^2} = \sqrt{0.199} = 0.45$$
$$C_{2\theta} = \arctan C_{2y}/C_{2x} = \arctan -0.826 = -39.6^\circ$$
$$C_2 = 0.45 \angle{-39.6^\circ}$$

The angle of the load reflection coefficient, $\Gamma_{ML\theta}$ is $C_{2\theta}{}^*$ where the asterisk indicates that the sign of the angle is changed. In this case, $C_{2\theta}{}^* = +39.6^\circ$. The magnitude of the load reflection coefficient is found from **eq. 7**:

$$|\Gamma_{ML}| = \frac{B_2 \pm \sqrt{B_2{}^2 - 4|C_2|^2}}{2|C_2|}$$

where the sign ahead of the radical is opposite to the sign on B_2.

$$|\Gamma_{ML}| = \frac{0.96 - \sqrt{(0.96)^2 - 4(0.45)^2}}{2(0.45)}$$
$$= 0.70$$
$$\Gamma_{ML} = 0.70 \angle 39.6^\circ$$

This quantity may be plotted on a Smith chart to determine the desired load impedance as shown in **fig. 1.*** This yields

$$Z_L = 62.0 + j108.5$$

*The load impedance can also be calculated from the relationship

$$Z_L = Z_o \frac{1 + \Gamma_{ML}}{1 - \Gamma_{ML}}$$

where Z_L is the complex load impedance $(R \pm jX)$, $Z_o = 50 + j0$, and Γ_{ML} is the complex load reflection coefficient. Since all quantities are complex numbers, vector arithmetic is required.

Input matching. Now that the output load impedance has been specified, the source reflection coefficient which will properly terminate the input to the transistor can be calculated from **eq. 8**:

$$\Gamma_{MS} = \left[s_{11} + \left(\frac{s_{12} \cdot s_{21} \cdot \Gamma_{ML}}{1 - (\Gamma_{ML} \cdot s_{22})} \right) \right]^*$$

$$= (0.47 \angle 161^\circ) + \left(\frac{(0.08 \angle 64^\circ)(3.1 \angle 63^\circ)(0.7 \angle 39.6^\circ)}{1 - [(0.7 \angle 39.6^\circ)(0.43 \angle -41^\circ)]} \right)^*$$

$$= (0.47 \angle 161^\circ) + \left(\frac{(0.08 \cdot 3.1 \cdot 0.7) \angle 64^\circ + 63^\circ + 39.6^\circ}{1 - [(0.7 \cdot 0.43) \angle (39.6^\circ - 41^\circ)]} \right)^*$$

$$= (0.47 \angle 161^\circ) + \left(\frac{0.17 \angle 166.6^\circ}{0.70 \angle 0.60^\circ} \right)^*$$

$$= [(0.47 \angle 161^\circ) + (0.24 \angle 166^\circ)]^*$$

Converting to rectangular notation and adding the x and y components,

$$\Gamma_{MSx} = -0.68 \qquad I_{MSy} = 0.21$$

Returning to polar notation

$$\Gamma_{MSR} = \sqrt{\Gamma_{MSx}^2 + \Gamma_{MSy}^2} = \sqrt{0.51} = 0.71$$

$$\Gamma_{MS\theta} = \arctan \Gamma_{MSy} / \Gamma_{MSx} = \arctan -0.31$$

$$= 162.7^\circ$$

$$\Gamma_{MS} = 0.71 \angle 162.7^{\circ *} = 0.71 \angle -162.7^\circ$$

The source reflection coefficient for a complex conjugate input match may be plotted on a Smith chart to determine the corresponding source impedance. This yields the series complex impedance, $Z_s = 8.7 - j7.4$ ohms (parallel complex impedance, $Z_p = 15 \| -j17.6$ *ohms*).

Matching networks. An input conjugate match can be obtained by shunting the transistor base with a capacitive reactance equal to the desired parallel equivalent reactance value (-j17.6 ohms) and transforming the source impedance to the required parallel resistance value (15 ohms) through a quarter-wavelength transmission line. The required capacitance value is found from the familiar reactance equation

$$C = \frac{1}{2\pi f X_c}$$

At 1296 MHz:

$$C = \frac{1}{2\pi(1296 \cdot 10^6)\, 17.6} = 6.98\ pF$$

A 10 pF trimmer capacitor will assure a proper reactive termination.

Transformation of the resistive component (to a 50-ohm source in this case) is accomplished with a quarter-wavelength transmission line which has a characteristic impedance Z_o, equal to the geometric mean of the source impedance Z_f, and the parallel input resistance, R_p.

$$Z_o = \sqrt{R_p \cdot Z_f}$$

With a 50-ohm source and a parallel input resistance of 15 ohms

$$Z_o = \sqrt{15 \cdot 50} = 27.4\ ohms$$

At 1296 MHz this is easily provided by a microstrip transmission line 0.26 inches (6.5mm) wide and 1.16 inches (29.5mm) long on a 1/16-inch (1.5mm) double-clad, fiberglass printed-circuit board.

A similar quarter-wavelength transformer can be designed to match the resistive component, R_s, of the complex series output impedance (62 ohms) to a 50-ohm termination, Z_t. As before, the characteristic impedance of the transmission line is given by

$$Z_o = \sqrt{R_s \cdot Z_t} = \sqrt{62 \cdot 50} = 55.7\ ohms$$

At 1296 MHz this is provided by a microstrip transmission line 0.09 inches (2.3mm) wide and 1.21 inches (30.7mm) long on 1/16-inch (1.5mm) double-clad, fiberglass printed-circuit board.

The required inductive reactance in series with the collector (+j108.5 ohms) is provided by shunting a capacitive reactance across the output end of the quarter-wavelength transformer. The required capacitive reactance is given by

$$X_c = \frac{Z_o^2}{X_s} = \frac{55.7^2}{108.5} = 28.6\ ohms$$

At 1296 MHz:

$$C = \frac{1}{2\pi(1296 \cdot 10^6)\, 28.6} = 4.3\ pF$$

Again, a 10 pF trimmer capacitor will suffice. The circuit in **fig. 2** shows the complete matching layout.

performance comparison

A simplified amplifier design, in which source and load impedances appear as the complex conjugate of the impedances related to s_{11} and s_{22}, yields the circuit shown in **fig. 3.** This circuit was derived by matching to the following assumed shunt-equivalent impedances:

Parallel: Z_{in} *(derived from* s_{11}*)*

$$= 21 \| j56.5\ ohms$$

Series: Z_{out} *(derived from* s_{22}*)*

$$= 75 - j52.5\ ohms$$

A more rigorous analysis shows the actual device impedances to be:

Parallel: Z_{in} *(actual)* $= 15 \| +j17.6$ *ohms*

Series: Z_{out} *(actual* $= 62.0 - j108.5$ *ohms*

Note that the reactive components of the shunt input impedance and the series output impedance differ significantly. Thus some degree of mismatch can be anticipated if the circuit of **fig. 3** is built as shown. Since the actual device impedances are now known, this mismatch can be accurately predicted.

As it happens, only the resistive component of the transistor's input or output complex impedance sees a mismatch. This is because the tuning range of the trimmer capacitors in **fig. 3** is sufficiently wide to properly terminate the reactive components. The input and output mismatches are determined by transforming the actual resistive components through the existing quarter-wave length transformers and comparing the resulting impedance to 50 ohms. Referring to **fig. 3**,

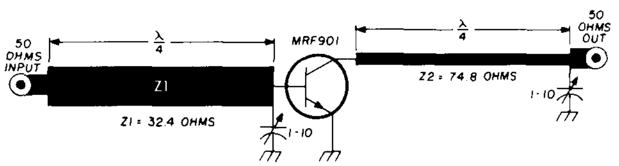

fig. 3. 1296-MHz amplifier which was designed with a simplified method. Input and output matching is provided by microstriplines and 10 pF trimmer capacitors. This amplifier is unconditionally stable and has about 0.5 dB less gain than the circuit shown in fig. 2 (bias networks are not shown).

$$Z_{in}\ (amplifier) = \frac{Z_1^2}{R_{p\ in}} = \frac{32.4^2}{15.0} = 70.0\ ohms$$

$$Z_{out}\ (amplifier) = \frac{Z_1^2}{R_{s\ out}} = \frac{74.8^2}{62.0} = 90.2\ ohms$$

Thus, the input vswr is 1.4:1 and the output vswr is 1.8:1, calculated values which correlate quite closely with those values observed in the actual amplifiers.

These input and output mismatches will result in somewhat lower stage gain than available from a properly terminated device Actual stage gain is found from

$$A_p\ (dB) = MAG + G_1 + G_2$$

where G_1 and G_2 are both negative and represent the mismatch losses at the input and output, respectively. Since G_1 (for a 1.4:1 vswr) is about -0.1 dB, and G_2 (for a 1.8:1 vswr) is about -0.4 dB,

$$A_p = 13.2 + (-0.1) + (-0.4)$$
$$= 12.7\ dB$$

This closely represents the measured gain of the amplifier shown in **fig. 3.**

summary

A method has been outlined for using device s-parameters to analyze the gain and stability of a microwave amplifier, and to determine appropriate source and load impedances for a complex conjugate match. It has been shown that designing around the reflection coefficients of a particular transistor (while ignoring the transfer coefficients) resulted in input and output mismatches of 1.4:1 and 1.8:1, respectively, while degrading overall amplifier gain by approximately 0.5 dB. This is a modest penalty for enjoying the convenience of a simplistic design approach. Whether the additional performance available from the more rigorous design method is justified depends largely upon the goals of the designer, and the intended application of the amplifier.

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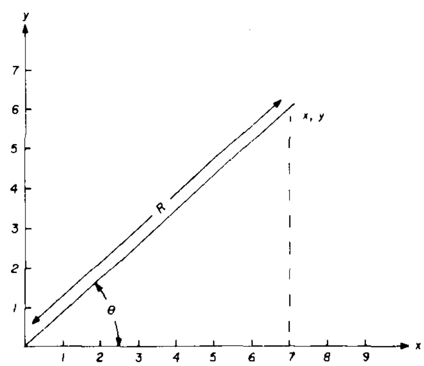
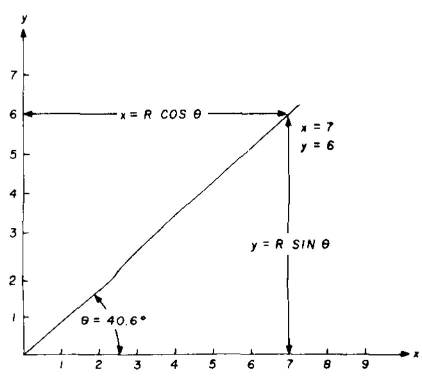
appendix 1

vector arithmetic

In the application of s-parameter design equations, it's necessary to perform numerous computations involving vector quantities. Vectors may be expressed either in conventional polar notation (magnitude R and associated angle θ) or may be resolved into their rectangular



components (x and y displacement on Cartesian coordinates) as shown below. Since the vector is part of a right triangle, manipulation between the two forms of notation involves the application of trigonometric functions. These manipulations may be accomplished on a slide rule, manually with the aid of trig tables, or on a hand-held digital calculator.



Readers who have advanced scientific calculators which include polar-rectangular conversion and summation keys will find the process considerably simplified. The following review is for the benefit of those not so fortunate.

1. Resolving vectors. Vector arithmetic often requires that the x and y components of the vector be known. Any vector, V, described by magnitude, R, and angle θ, can be resolved into its x and y components with the following formulas

$$x = R \cos \theta$$
$$y = R \sin \theta$$

Example: What are the x and y components of the vector $9.22 \angle 40.6^\circ$ $(R = 9.22, \theta = 40.6^\circ)$?

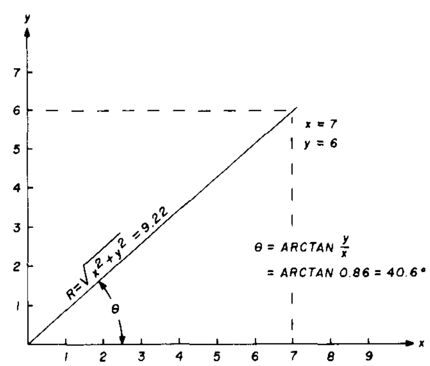
$$x = R \cos \theta = 9.22 \cos 40.6^\circ$$
$$= 9.22 \cdot 0.76 = 7.00$$
$$y = R \sin \theta = 9.22 \sin 40.6^\circ$$
$$= 9.22 \cdot 0.65 = 6.00$$

2. Constructing vectors. The results of vector addition and subtraction (reviewed later) generally appear as x and y components of the resultant vector. These coordinates can be converted to a polar vector of magnitude, R, and angle, θ, with measurements on a graphical plot, or by using a trigonometric solution. Since the vector is represented by the hypotenuse of a right triangle formed by dimensions x and y, the Pythagorean theorem may be used to find the magnitude R:

$$R = \sqrt{x^2 + y^2}$$

Trigonometry is used to calculate the angle θ

$$\theta = \arctan y/x$$



Example: What is the magnitude, R, and angle θ, of the vector described by the Cartesian coordinates, $x = 7$, $y = 6$?

$$R = \sqrt{x^2 + y^2} = \sqrt{49 + 36}$$
$$= \sqrt{85} = 9.22$$
$$\theta = \arctan y/x = \arctan 6/7$$
$$= \arctan 0.86 = 40.6^\circ$$

3. Vector addition. Any two vectors V_1 and V_2, which have been resolved into x and y components V_{1x}, V_{1y}, V_{2x}, and V_{2y}, may be added by summing the respective x and y components. The summed components may then be constructed into a resultant vector V_t, of magnitude V_{tR} and angle $V_{t\theta}$.

Example: What is the sum of vectors V_1 and V_2 when

$$V_1 = 1.5 \angle +40^\circ$$
$$V_2 = 2.0 \angle -60^\circ$$
$$V_{1x} = V_{1R} \cos V_{1\theta} = 1.5 \cos 40^\circ = 1.15$$
$$V_{1y} = V_{1R} \sin V_{1\theta} = 1.5 \sin 40^\circ = 0.96$$
$$V_{2x} = V_{2R} \cos V_{2\theta} = 2.0 \cos -60^\circ = 1.00$$
$$V_{2y} = V_{2R} \sin V_{2\theta} = 2.0 \sin -60^\circ = -1.73$$
$$\Sigma_x = V_{1x} + V_{2x} = 1.15 + 1.00 = 2.15$$
$$\Sigma_y = V_{1y} + V_{2y} = 0.96 + (-1.73) = -0.77$$
$$V_{tR} = \sqrt{\Sigma_x{}^2 + \Sigma_y{}^2} = \sqrt{2.15^2 + (-0.77)^2}$$
$$= \sqrt{5.22} = 2.28$$
$$V_{t\theta} = \arctan (\Sigma_y / \Sigma x) = \arctan (-0.77/2.15)$$
$$= -19.70^\circ$$
$$V_{t\theta} = V_1 + V_2 = 2.28 \angle -19.70^\circ$$

4. Vector subtraction. Any two vectors V_1 and V_2, which have been resolved into x and y components V_{1x}, V_{1y}, V_{2x}, and V_{2y}, may be subtracted one from the other by subtracting their respective x and y components. The results of such subtraction comprise the x and y components of the resulting vector V_t, which may be constructed to yield magnitude V_{tR} and angle $V_{t\theta}$.

Example: What is the difference when vector V_2 is subtracted from vector V_1?

$$V_1 = 0.8 \angle -40^\circ$$
$$V_2 = 0.4 \angle +120^\circ$$
$$V_{1x} = V_{1R} \cos V_{1\theta} = 0.8 \cos -40^\circ = +0.61$$
$$V_{1y} = V_{1R} \sin V_{1\theta} = 0.8 \sin -40^\circ = -0.51$$
$$V_{2x} = V_{2R} \cos V_{2\theta} = 0.4 \cos 120^\circ = -0.20$$
$$V_{2y} = V_{2R} \sin V_{2\theta} = 0.4 \sin 120^\circ = +0.35$$
$$(\Sigma_{-x}) = V_{1x} - V_{2x} = +0.61 - (-0.20) = +0.81$$
$$(\Sigma_{-y}) = V_{1y} - V_{2y} = -0.51 - (+0.35) = -0.86$$
$$V_{tR} = \sqrt{(\Sigma_{-x})^2 + (\Sigma_{-y})^2} = \sqrt{(0.81)^2 + (-0.86)^2}$$
$$= \sqrt{1.40} = 1.18$$
$$V_{t\theta} = \arctan (\Sigma_{-y})/(\Sigma_{-x}) = \arctan (-0.86/+0.81)$$
$$= -46.7^\circ$$
$$V_t = V_1 - V_2 = 1.18 \angle -46.7^\circ$$

5. Vector multiplication. For any two vectors, V_1 and V_2, each described by a magnitude, R, and an angle, θ, the vector product is found by multiplying the the magnitudes and adding the angles:

$$V_{tR} = V_{1R} \times V_{2R}$$
$$V_{t\theta} = V_{1\theta} + V_{2\theta}$$

Example: What is the vector product of V_1 and V_2 when

$$V_1 = 0.8 \angle 45^\circ$$
$$V_2 = 0.65 \angle -118^\circ$$
$$V_{tR} = V_{1R} \times V_{2R} = 0.8 \times 0.65 = 0.52$$
$$V_{t\theta} = V_{1\theta} + V_{2\theta} = +45^\circ + (-118^\circ) = -73^\circ$$
$$V_t = V_1 \cdot V_2 = 0.52 \angle -73^\circ$$

6. Vector division. For any two vectors, V_1 and V_2, each described by a magnitude, R, and an angle, θ, the vector quotient is found by dividing the magnitudes and subtracting the angles:

$$V_{tR} = V_{1R} \div V_{2R}$$
$$V_{t\theta} = V_{1\theta} - V_{2\theta}$$

Example: What is the vector quotient when V_1 is divided by vector V_2?

$$V_1 = 0.96 \angle +64^\circ$$
$$V_2 = 0.42 \angle -102^\circ$$
$$V_{tR} = V_{1R} \div V_{2R} = 0.96 \div 0.42 = 2.29$$
$$V_{t\theta} = V_{1\theta} - V_{2\theta} = +64^\circ - (-102^\circ) = +166^\circ$$
$$V_t = 2.29 \angle +166^\circ$$

7. Maximum angle. Whenever a vector manipulation yields an expression whose angle exceeds ±180°, subtract the absolute value of the angle from 360°, and assign to the resulting angle a sign opposite to that of the original angle.

Examples: $-196^\circ = +360 - |-196|$

$$= +360 - 196 = +164^\circ$$

$$265^\circ = -360 - |+265|$$

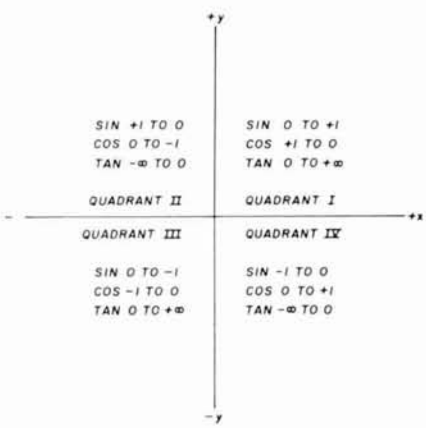
$$= -360 - 265 = -95^\circ$$

8. **Compound expressions.** In expressions involving both vector and scalar quantities, treat the scalar quantity as though it were a vector of angle 0°.

Example: In the expression for the source reflection coefficient which will properly terminate the transistor's input (**eq. 8**), the product of the load reflection coefficient, Γ_{ML}, and s_{22} are subtracted from one. If $(\Gamma_{ML} \cdot s_{22}) = 0.65 \angle -118^\circ$, what is the value of the expression, $1 - (\Gamma_{ML} \cdot s_{22})$?

$$V_1 = 1 \angle 0^\circ$$
$$V_2 = 0.65 \angle -118^\circ$$
$$V_{1x} = V_{1R} \cos V_{1\theta} = 1 \cos 0^\circ = 1.00$$
$$V_{1y} = V_{1R} \sin V_{1\theta} = 1 \sin 0^\circ = 0$$
$$V_{2x} = V_{2R} \cos V_{2\theta} = 0.65 \cos -118^\circ = -0.31$$
$$V_{2y} = V_{2R} \sin V_{2\theta} = 0.65 \sin -118^\circ = -0.57$$
$$\Sigma_{-x} = V_{1x} - V_{2x} = 1.00 - (-0.31) = 1.31$$
$$\Sigma_{-y} = V_{1y} - V_{2y} = 0 - (-0.57) = 0.57$$
$$V_{tR} = \sqrt{(\Sigma_{-x})^2 + (\Sigma_{-y})^2} = \sqrt{1.31^2 + 0.57^2}$$
$$= \sqrt{2.04 = 1.43}$$
$$V_{t\theta} = \arctan(\Sigma_{-y}/\Sigma_{-x}) = \arctan 0.57/1.31$$
$$= 23.51^\circ$$
$$1 - (\Gamma_{ML} \cdot s_{22}) = 1.43 \angle 23.51^\circ$$

9. **Angular functions.** Since most trigonometry tables show only the functions to +90°, when working with vectors which may fall in any of the four quadrants below (0 through 360 degrees) this can lead to ambiguities in specifying the angle θ of a resultant vector. Note that the



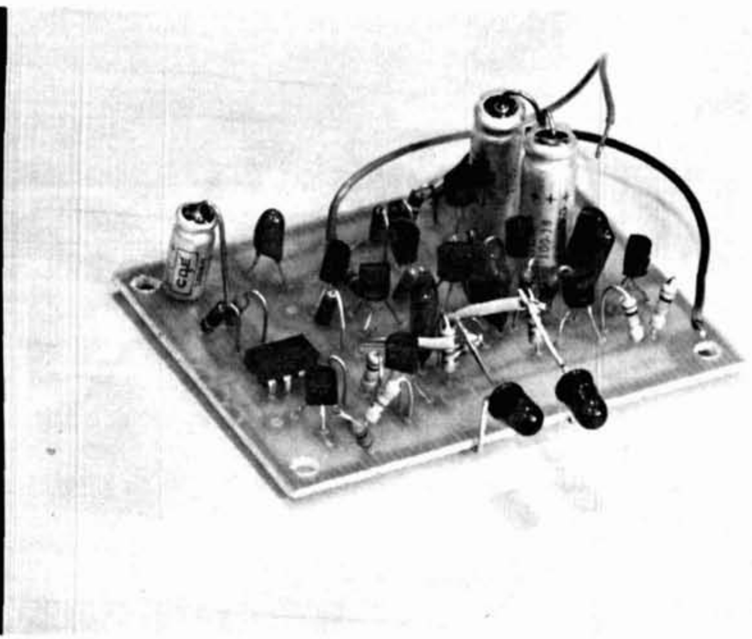
tangent function varies from zero to +∞, from zero to 90 degrees, from −∞ to zero in the second quadrant (90 to 180 degrees), from zero to +∞ in the third quadrant (180 to 270 degrees), and from −∞ to zero in the fourth quadrant (270 through 360 degrees). The sine and cosine functions are also ambiguous, as shown, but this doesn't create a problem in vector arithmetic.

In the expression for the angle of the source reflection coefficient, $\Gamma_{MS\theta}$, in the design example $\Gamma_{MSx} = -0.68$ and $\Gamma_{MSy} = 0.21$. Therefore,

$$\Gamma_{MS\theta} = \arctan -0.68/0.21 = \arctan 0.31$$
$$= -17.3^\circ$$

This is in the fourth quadrant whereas the x and y values place the vector in the second quadrant. Therefore, the correct value for the angle is 180° + (−17.3°) = 162.7°. The same sort of ambiguity exists for the first and third quadrants, and can only be resolved by inspection.

ham radio



two-channel scanner for repeater monitoring

Simple modifications to WA2GCF's popular channel scanner for use with the Icom IC-2F and other vhf-fm transceivers

The two-channel scanner described in this article was never really designed – it evolved. I wanted a simple scanner to monitor two repeaters, using a small solid-state transceiver. The simplest circuit I found was one described by WA2GCF in an article in *ham radio*.[1] The unit proved to have all the advantages the author claimed; it was inexpensive, small, easy to assemble, and operates directly from the 13-volt transceiver power supply or an auto battery. Only minor modifications were necessary to monitor the two club repeaters with my Icom IC-2F.

The first problem was how to use diode switching with the IC-2F receiver oscillator. WA2GCF described circuitry for switching crystals in oscillators requiring +5 volts to turn on, and oscillators requiring a ground for turn-on. Both are designed for circuits where one side of the crystal is grounded. The IC-2F, however, like many other Japanese receivers, has a receiver oscillator in which the crystal is connected between the collector and base of the oscillator transistor (**fig. 1A**). Neither side of the crystal can be grounded. At first it appeared it would be necessary to redesign the oscillator circuit, but the switching circuit modification shown in **fig. 1B**, which required rewiring only the switch, proved useable.

squelch recognition

Scanning action is stopped when a carrier opens the receiver squelch. The original scanner could be stopped by a squelch circuit that goes to ground when an incoming signal is received, or with alternate wiring by a squelch circuit that is at ground potential when the squelch is closed and goes high when it is opened. Unfortunately, the IC-2F does neither. Its squelch circuit is high when the squelch is closed, and goes to a lower voltage when it opens, but not all the way to ground. The scanner stopped for strong signals but not for ones that were weak but still quite readable.

The original circuit used to stop the scanner multivibrator is modified by changing Q1 to a pnp transistor with its emitter connected to the positive supply voltage, and adding an npn transistor, Q1A, to control it as

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shown in **fig. 2.** The base of Q1A is biased so that it will be turned off when the squelch circuit connected to its emitter is high; it is turned on by a voltage less than the bias but still above ground potential, which occurs when the squelch opens. When Q1A is off, Q1 and Q2 are also off, and the multivibrator can oscillate. When the three transistors turn on, the multivibrator is locked and scanning action stops.

The scanner still had one disadvantage for monitoring our two repeaters, which all our control operators and many club members want to do. If the repeater carrier does not drop between users' transmissions, the scanner locks on that repeater and will stay there for the entire conversation. Without a periodic search-back feature there is no way for the listener to know what may be happening on the other repeater.

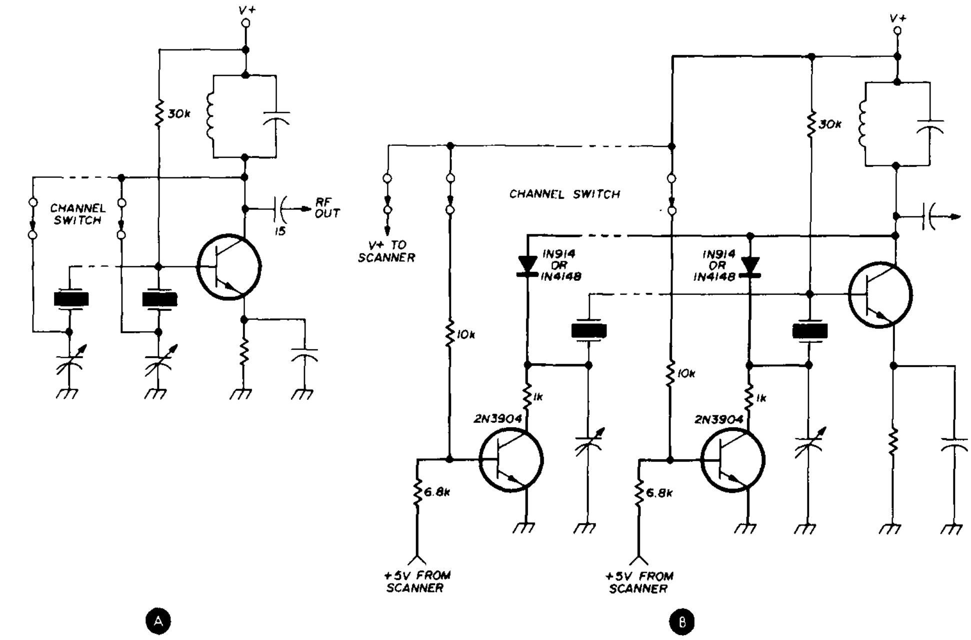


fig. 1. IC-2F receiver oscillator circuit. (A). As originally wired. (B). Modified for diode switching. Heavy lines show wiring changes.

To prevent lock-up on any channel for long periods a lock release, rather than a priority channel circuit, is needed. This can be done by adding a timer to the squelch recognition circuit described above. Instead of

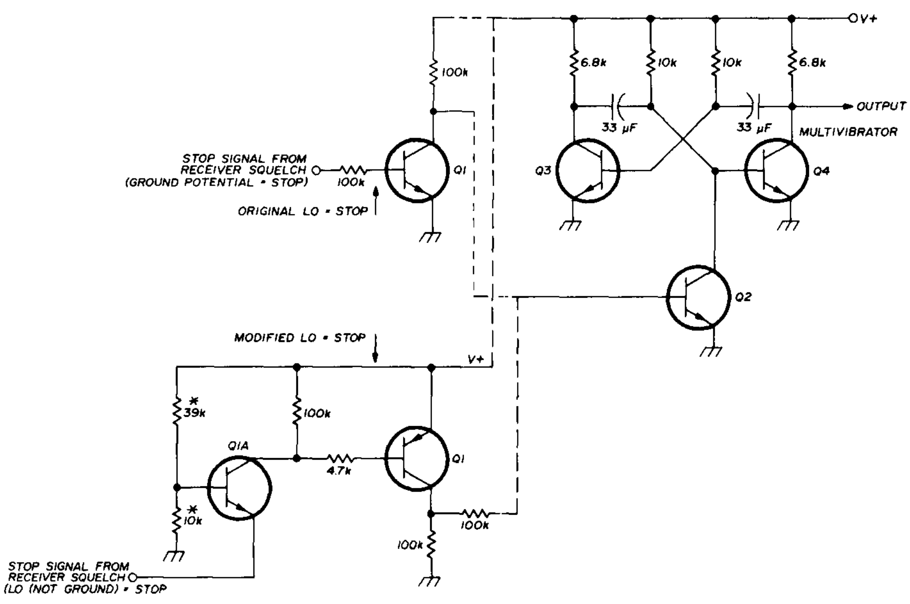


fig. 2. Multivibrator circuit modification to stop scanning action with low but above-ground stop signal.

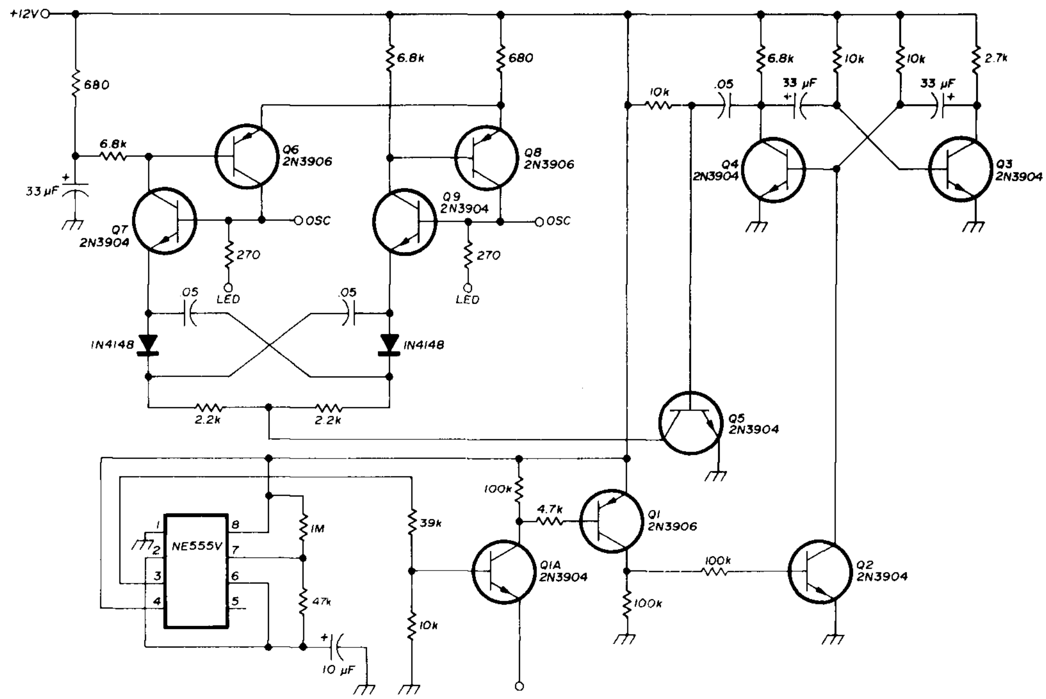


fig. 3. Two-channel scanner with search-back feature.

providing bias for Q1A from the V+, it is supplied from the output of an NE555 timer IC which is programmed to be high for about 15 seconds and low for a fraction of a second. Each time the IC cycles the scanner will move to the next frequency. With a two-channel scanner both frequencies are checked at least every 15 seconds. Even with four channels none is unguarded for more than a minute.

control switching

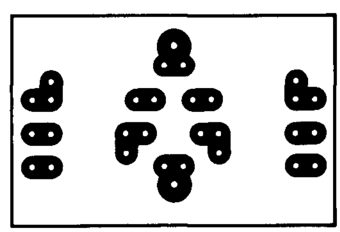
The scanner does not switch transmitter crystals, so manual switching to a selected frequency is necessary when the operator wants to talk. My IC-2F is a six-channel transceiver, with a six-position, double-pole, channel selector switch. As originally wired in **fig. 1A**, the common terminal on the receive side was connected to the collector of the oscillator. Rewired as in **fig. 1B**, this half of the switch controls V+ to the crystal switching circuits of four crystals. The other two positions, in which there are no crystals, control V+ to the scanner and the synthesizer, so that the scanner is disabled when the switch is turned to any transmitting position.

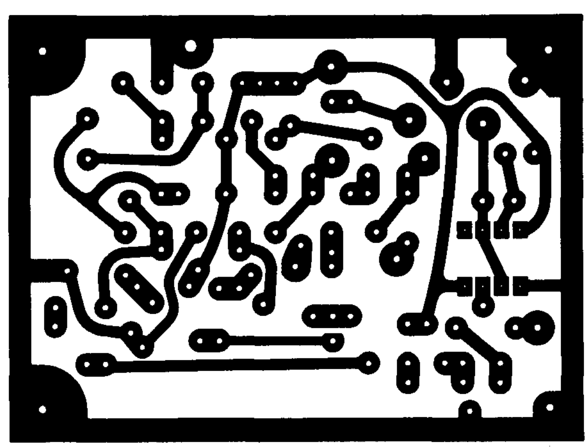
construction

The scanner modifications shown here can be easily applied to the scanner described in WA2GCF's article. For those who want only two channels, the complete circuit is reproduced in **fig. 3**. Full-size circuit board patterns for the scanner and the crystal-switching circuit are shown in **fig. 4**.* Component placement is shown in **figs. 5** and **6**. All of the components except the circuit boards can be purchased from *ham radio* advertisers for less than $5, or obtained locally. Tolerances are not critical, and almost any switching diodes and general-purpose silicon transistors can be substituted for those shown without affecting operation.

The 680-ohm resistor and 33-μF capacitor in the bias supply of Q6 unbalance the circuit sufficiently to insure

fig. 4. Full-size printed-circuit layouts for the switching circuit (top) and the scanner (below).





*A set of drilled, glass-epoxy, printed-circuit boards is available for $5, postpaid, from D.L. McClaren, 19721 Maplewood Ave., Cleveland, Ohio 44135

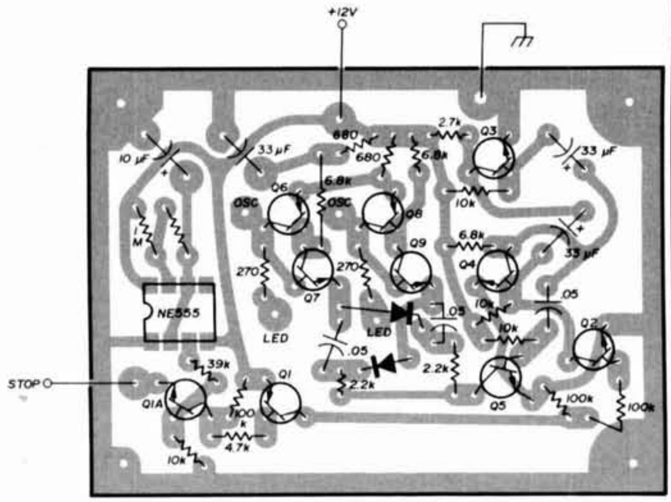


fig. 5. Printed-circuit component layout for the scanner viewed from the component side of board. Circles indicate external connections.

the start of scanning action. With the scan rate provided by the multivibrator circuit components shown in **fig. 3**, and the NE555 timing values, the scanner should stop every time it senses an active frequency.

adjustment

The scan rate can be varied by changing the values of the multivibrator capacitors. The length of time the scanner will stay locked on a single frequency can be

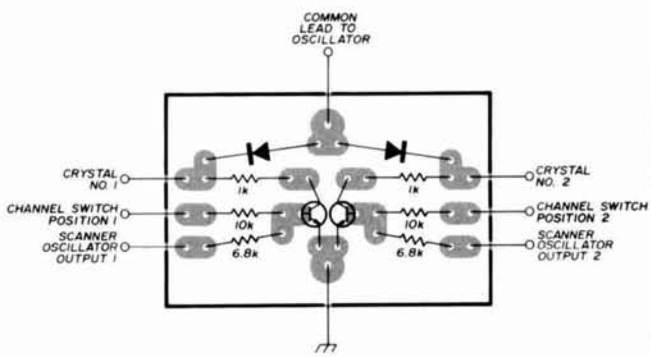


fig. 6. Printed-circuit component layout for oscillator switching circuit shown in fig. 1B.

changed by replacing the 1-megohm NE555 charging resistor with a different value. If these timing components are changed, it may be necessary to change the value of the 47k discharge resistor, also, to limit the multivibrator to only one cycle each time the NE555 output goes low. Excessive low time will cause the scanner to miss a signal on the next channel.

reference

1. Jerry Vogt, WA2GCF, "Improved Channel Scanner for VHF FM," *ham radio*, November, 1974, page 26.

ham radio

how to improve receiver performance of vacuum-tube vhf-fm equipment

Simple circuit modification for replacing 6AK5 receiver rf amplifier stages with a low noise, dual-gate, mosfet

Vhf fm operation is becoming increasingly popular with radio amateurs, with tens of thousands of fm transmitters and receivers now in use, and more being put into service daily. Not an insignificant number of these rigs are converted from commercial service – equipment that was originally manufactured by firms such as Motorola, RCA, and GE. However, most of the older receivers use vacuum-tube front ends so they don't provide the sensitivity that is possible with modern solid-state devices. This article describes a simple, proven mosfet circuit that can be easily substituted for those noisy tube-type front ends.

The most popular receiver rf amplifier used by communications equipment designers back in the late 1950s and early 1960s was based on the 6AK5 pentode – a workhorse used heavily during the war in vhf radar and communications systems. Typical receiver sensitivities provided by this tube are on the order of 1.5 to 3.0 microvolts for 20 dB quieting. Considering the state of the art for those days, that was not shabby performance.

In recent years the communications designer has been provided with a proliferation of very reasonably priced, high performance, dual-gate mosfet devices from several different manufacturers. One example is the 3N204 by Texas Instruments which has optimum spot noise figures of 2 dB at 200 MHz, 3 dB at 400 MHz, degrading to 7 dB at 900 MHz, all for $1.25 in small quantities. Anyone who has worked with a 417A/5842 vacuum-tube converter from the 1950s can certainly appreciate how rapidly technology has marched forward during the past two decades.

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It's a relatively simple task to replace the 6AK5 rf amplifier in your present rig with a 3N204 or other dual-gate device. **Fig. 1** shows the circuit of a typical 6AK5 rf amplifier stage found in many commercial rigs, while **fig. 2** shows the base diagram of the 6AK5. The following steps describe the circuit modification, which can be accomplished in less than an hour.

1. Remove the center pin from the 6AK5 tube socket by bending it over and breaking it with a pair of pliers. Cut any wires which are soldered to the center grounding pin.

2. Install a 33k, 1/2-watt resistor from pin 6 to ground. If there is already another resistor from pin 6 to ground, remove it. Make sure that there is a 0.001 μF disc bypass capacitor between pin 6 and ground.

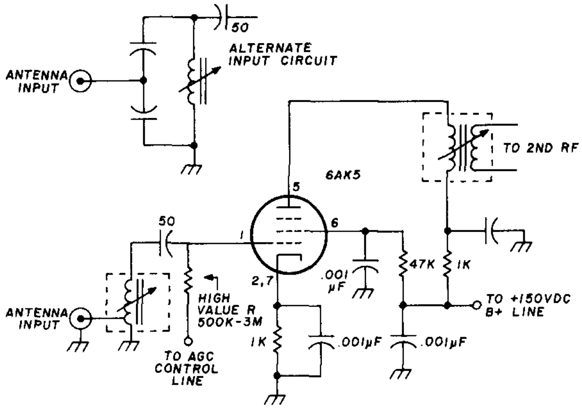


fig. 1. Typical 6AK5 receiver rf amplifier stage found in many older commercial vhf-fm rigs. Performance of this circuit is typically 1.5 to 3.0 microvolts for 20 dB quieting. Replacing the 6AK5 with a vhf dual-gate mosfet, as described here, increases sensitivity to 0.25 to 0.30 microvolt for 20 dB quieting.

3. Remove the original resistor from the B+ line to pin 6 and replace it with a 120k, 1/2-watt resistor.

4. Remove the voltage-dropping resistor(s) from the B+ line to the B+ point on the 6AK5 rf amplifier plate output coil and replace it with a 300-ohm, 1/2-watt resistor.

5. Break the B+ line from the high-voltage feedpoint to the 6AK5. Now install a 3-lug terminal strip (two insulated lugs, one grounded lug) by soldering the ground lug to a convenient point on the chassis. Install a 6.8k, 2-watt resistor between the two insulated tie points and connect one end of the resistor to the previously removed B+ line. Connect an 18-volt zener diode (400 mW to 1 watt rating) from the other end of the 6.8k resistor to ground. From this same point connect a wire to the former B+ feedpoint for the 6AK5.

6. Insert the 3N204 mosfet into the center hole of the 6AK5 tube socket.

7. Using the 3N204 basing diagram in **fig. 3**, make the following connections to the 6AK5 tube socket:

Drain to pin 5
Source to ground (pin 3 or 4)
Gate 1 to pin 1
Gate 2 to pin 6

8. Disconnect the antenna input lead and reconnect it to the top of the rf input coil as shown in **fig. 3**.

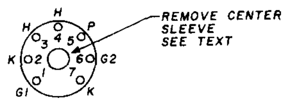


fig. 2. Base diagram (bottom view) of the 6AK5 tube socket.

9. Using a signal generator or on-the-air signal, repeak the tuned input and output circuits for maximum output.

10. Remove the 6-volt filament wiring from the socket, but before you do, check to see if two or more tubes are wired in series. If they are, install a 36-ohm, 2-watt resistor in series with the filament line to compensate for the current drain of the 6AK5 filament. This completes the conversion.

I have converted several 6AK5 rf amplifier stages to dual-gate mosfets using this simple procedure, and all have provided outstanding results. Sensitivity measurements using a Hewlett-Packard 608D signal generator indicate a sensitivity of about 0.25 to 0.30 microvolts for 20 dB quieting – a marked improvement in performance.

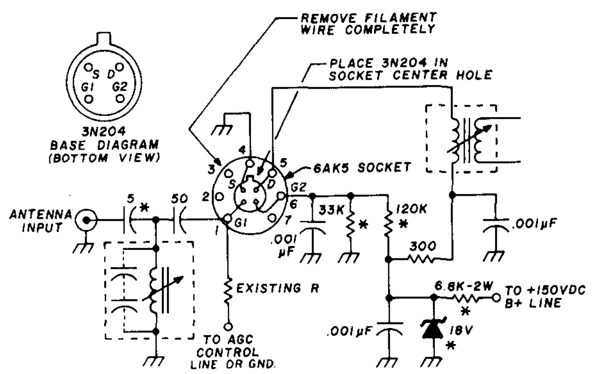


fig. 3. Modified receiver rf amplifier using a 3N204 dual-gate mosfet. Components marked with an asterisk are new.

In addition to the 3N204 I have tried several other, similar dual-gate devices, including the 3N201 and 3N200, all with good success. All of these devices are in the same price class, although the 3N204 has a bit better performance. If you're using an older tube-type commercial rig, this simple modification can significantly extend your receiving range and operating pleasure.

ham radio

RC active filters

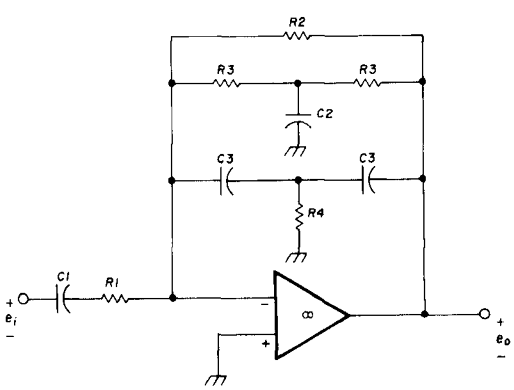
using op amps

An overview of three of the most popular filter circuits,

A narrow bandpass filter of small size and low power consumption is often needed for increased receiver selectivity or other communications applications, such as RTTY. Active filters are also presently found in IC form for applications in modems, function generators, and level detectors.[1] This article presents information that has been omitted from the many articles on the subject in the amateur literature. Three of the most popular RC active filters are described together with their design equations and response characteristics. Also included is a discussion of breadboard testing and results of filter performance both singly and in cascaded form.

choice of circuit

The parallel-T network is a popular narrowband filter, which is used in the feedback loop of an op amp in various ways. In this case the network was inserted between the inverting input and op-amp output: In this



arrangement R1, C1 form part of the network, allowing a symmetrical response about center frequency. This added input network increases the lower-frequency skirt selectivity, which otherwise would not fall off as fast as the upper-frequency skirt characteristic. The frequency-response equation including phase relationships is:

$$\frac{e_o}{e_i}(\omega) = -\left\{\frac{1}{R1+\frac{1}{j\omega C1}}\right\}\left\{\frac{R2[2+j\omega C2R3]}{\left[1-\omega^2\left(\frac{C2R3}{2}\right)^2\right]\frac{R2}{R3}+\left[2+j\omega C2R3\right]}\right\}$$

The response equation remains in this form since it becomes rather cumbersome to deal with if expanded further. The design equations for the filter are:

$$f_o = \frac{1}{\pi C2R3}\sqrt{1+\frac{2R3}{R2}} \approx \frac{1}{\pi C2R3} \qquad (2)$$

$$C1 = \frac{\alpha}{2\pi f_o R1} \qquad (3)$$

where $\alpha = \sqrt{2}$

$$R2 = R1 = \frac{1}{2R3(\pi f_o C2)^2} \qquad (4)$$

for $\frac{e_o}{e_i} = 1$

$$R3 = 2R4 \qquad (5)$$

Typical values chosen for this filter were:

R1 39,000 ohms	R4 1727 ohms	C3 0.05 μF
R2 39,000 ohms	C1 0.006 μF	f_o 1000 Hz
R3 3454 ohms	C2 0.1 μF	3-dB BW,
Q 14 (four stages); shape factor 13		(four Stages 71 Hz)

The Q can be made very high if desired, but the skirt

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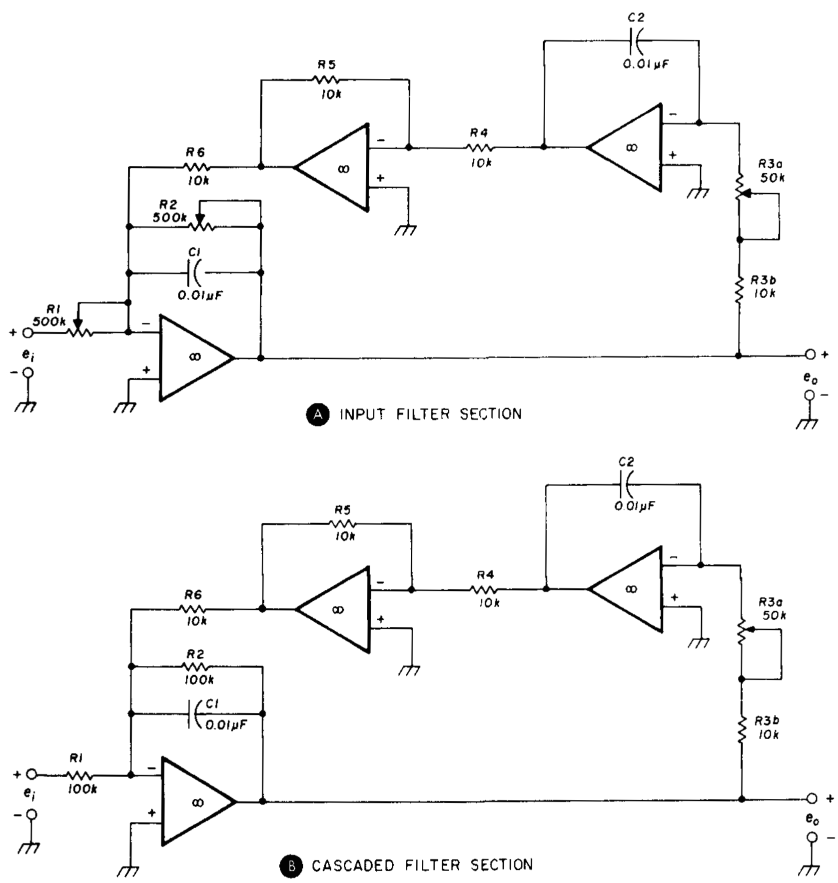
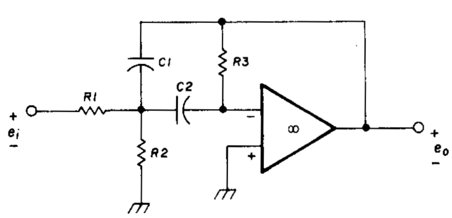


fig. 1. Input bandpass filter section, A, and cascaded sections, B, used for analysis. The rationale for component values is discussed in the text.

selectivity or bandpass-filter shape factor with only one stage remains poor. With three or four stages, the skirt selectivity and shape factor become fairly good for a narrow-bandpass audio filter; however, the component tolerance sensitivity is high.

Another circuit employing feedback in a slightly different manner (the same approach used by the popular MFJ-CWF2 filter[2]) to acquire the bandpass characteristic is given below (for one stage):



The response and design equations can be derived in the usual manner as in the previous case using flow-graph analysis. They are given below for reference:

$$\frac{e_o}{e_i}(\omega) = -\frac{j\omega[1/R1C1]}{\left[\frac{R1+R2}{R1R2R3C1C2} - \omega^2\right] + j\omega\left[\frac{C1/C2+1}{R3C1}\right]} \tag{6}$$

$$f_o = \frac{1}{2\pi\sqrt{\frac{R1+R2}{R1R2R3C1C2}}} \tag{7}$$

$$\frac{e_o}{e_i}(\omega_o) = -\frac{R3}{2R1} \tag{8}$$

$R1 = \frac{1}{2}f_oC$, where $C = C1 = C2$

$R3 = \alpha/f_oC$

$R2 = 1(2\pi)^2 f_oC\alpha$

$Q \approx \pi\alpha$, letting $\alpha = \sqrt{2}$

Again, this filter will not have good skirt selectivity and cascaded stages must be used to obtain a good shape factor. Otherwise, regardless of Q, signals will still be heard with respect to sideband frequencies far removed from the center frequency, f_o. Values chosen from the design equations for this filter are given below for four stages:

R1 975,000 ohms	C1 0.001 µF	f_o 1000 Hz
R2 13,165 ohms	C2 0.001 µF	3-dB BW 71 Hz
R3 1.95 megohms		Q = 14
Shape factor 13		

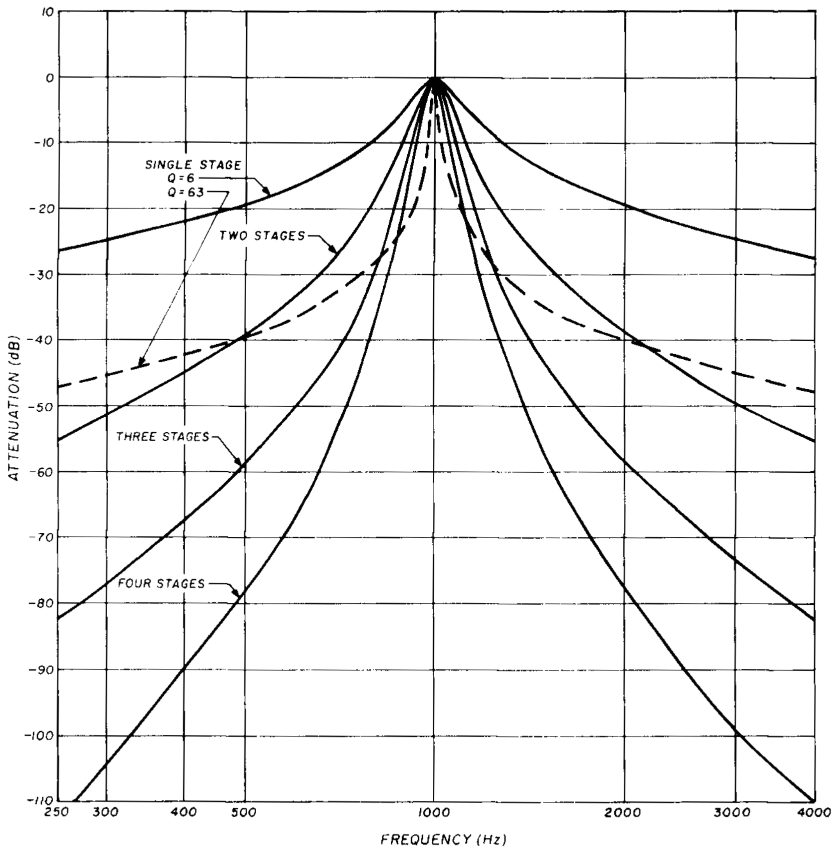


fig. 2. Response characteristics of single and cascaded filter sections. Cascading offers a more practical approach for R1 and R2 values, which are used to vary filter gain and Q (see fig. 1).

Still another circuit that is probably a little more complex is shown in **fig. 1**. The response and design equations for this filter are:

$$\frac{e_o}{e_i}(\omega) = -\frac{j\omega[1/R1C1]}{\left[\frac{1}{R3R6C1C2}\left(\frac{R5}{R4}\right) - \omega^2\right] + j\omega[1/R2C1]}$$

$$f_o = \frac{1}{2\pi}\sqrt{\left(\frac{1}{R3R6C1C2}\right) \cdot \left(\frac{R5}{R4}\right)}$$

$$\frac{e_o}{e_i}(\omega_o) = -\frac{R2}{R1} \qquad \textbf{(9)}$$

Choose $C1 = C2$
$R4 = R5$
$R3 = R6$

$$Q = 2\pi f_o C1R2$$

Typical component values chosen for this circuit were:

R1 98,000 ohms	R4 10,000 ohms	C1 0.01 μF
R2 98,000 ohms	R5 10,000 ohms	C2 0.01 μF
R3 15,915 ohms	R6 15,915 ohms	f_o 01000 Hz

3-dB BW 71 Hz (four stages); shape factor 13
(See **fig. 1** for final values).

The values for R1, R2, R3 were made variable for the first stage, while a potentiometer was used in the following stages for R3. (See **fig. 1**.) R1 varies the overall filter gain, R2 the Q, and R3 tunes each stage to the same desired center frequency. A value of 50k was more than adequate to cover the tolerance variations and range of interest. The variable pots, R1 and R2, were 500k maximum for the input stage. The remaining stages included the values given above except for R3, which was chosen to be 50k in series with a 10k resistor.

This approach shows no improvement in skirt selectivity over the others and, in the same manner, requires cascaded stages or building blocks as they are sometimes called.[3] However, this circuit has the advantage of not being so component-tolerance sensitive. In fact, the Q can be varied without affecting the gain or center frequency by varying R2; the gain can be varied independently by varying R1; and the center frequency can be varied independently by varying R6 or R3. The filter can be easily tuned using components with 10-percent tolerance. Also, you can adjust the tone or output level without varying other filter parameters, such as frequency or Q.

This completes the basic description of the three more popular bandpass filter circuits. The following discussion addresses the breadboard testing of each circuit, with a final filter design using the circuit having independent characteristic control.

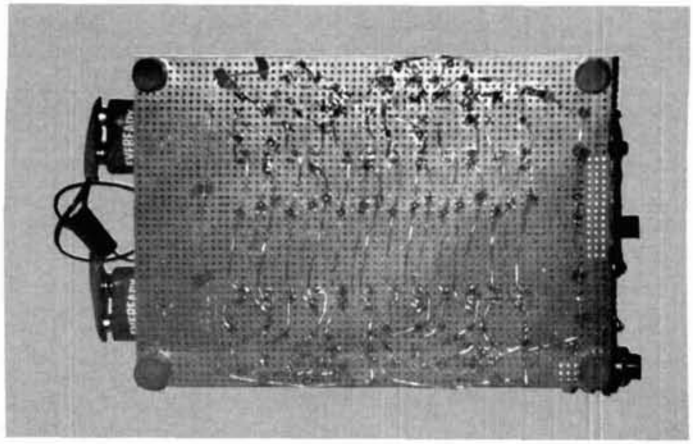
filter-circuit breadboards

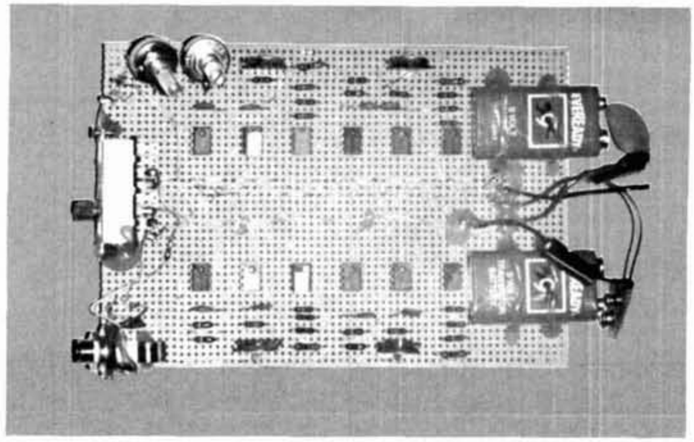
The parallel-T network was evaluated first. The feedback network using a parallel-T network was very sensitive to component variation. The measured filter characteristics agreed with theoretical results, using a programmable hand calculator, almost to the point where one would be satisfied without evaluating a breadboard circuit (if 1-percent or better component tolerances were used).

The second circuit evaluated was that used in the popular MWJ-CWF2, which uses matched components for each section – difficult to implement unless you have an impedance bridge. The measured results again agreed with theoretically derived results when 1-percent component tolerances were used.

When reviewing the references on this subject, I found that all attained the same success between theoretical and measured results at audio frequencies when using close tolerance values.[4] As in the parallel-T circuit, the second circuit was sensitive to component-value variation and only slight improvement was obtained regarding characteristic control.

The third circuit was by far the most superior in terms of varying filter characteristics independently of each other. Again, theoretical results agreed with measured results after tuning each stage to the desired center frequency. In all cases, the measured results departed from those of the theoretical case only when the operational amplifier characteristics no longer were allowed to assume a very high input impedance, a near-zero output impedance (power limited of course), and a very high open-loop gain (with respect to the Q and skirt selectivity plus stage-to-stage isolation).





Top and bottom (above and below) of the Vector board construction used for the experimental active bandpass filter using a single μ741 op amp.

construction and cost

Cost was kept to a minimum by using available barrel-kit components. In this case, 14 good μ741 ICs were found among fifty purchased, for a little less than two dollars. The quad operational amplifier, LM324, although a little more expensive, provided a much neater layout and also showed a slightly higher open-loop gain at 1000 Hz, where the μ741 drops of quite a bit.

Construction and layout were arranged so that either the quad operational amplifier chips or the single μ741 could be evaluated. The photos show top and bottom component layout. Printed circuit boards could have been used, but construction time and cost would have increased. Batteries were used; however, a separate power supply might be more desirable. The filter draws only about 17 mA.

concluding remarks

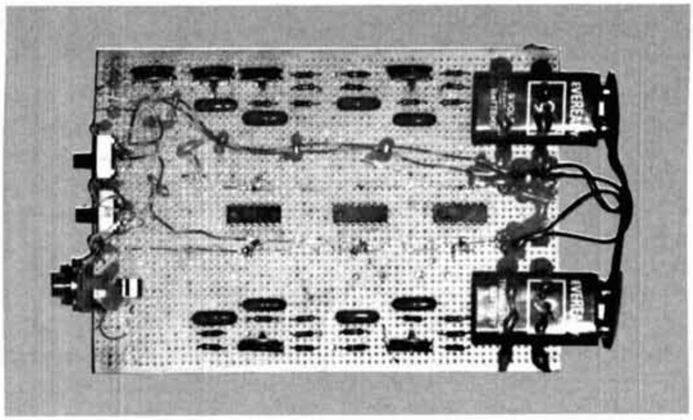
Fig. 2 illustrates the response characteristics as the stages are cascaded. At first I thought that one stage with a higher Q would suffice (the dashed curve shows the response characteristic of a single stage with a Q of 63). This arrangement didn't provide the desired skirt attenuation for rejecting strong signals close to center frequency, so additional stages were cascaded to improve the shape factor (60-dB bandwidth divided by the 3-dB bandwidth).

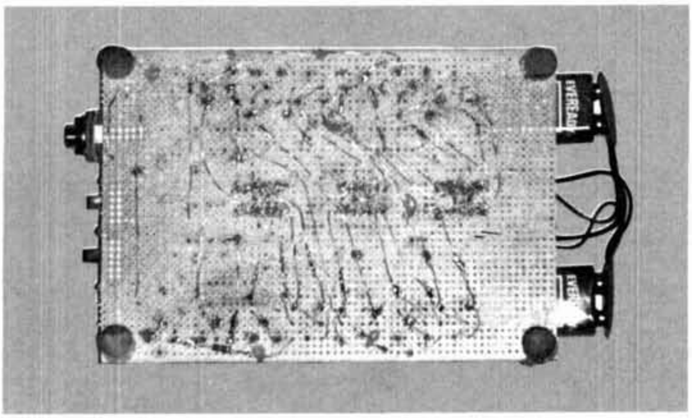
Cascading stages offered more practical and less component-sensitive values for R1 and R2. An important observation noted from the response curves is that only one stage needs a variable pot for varying Q, at least within the ranges normally desired. If a greater Q range is desired, pots can be used in all four sections.

The more practical approach for wide Q variation is to use a four-stage pot with a common shaft coupling. Similarly, the frequency adjustment requires only one pot per stage unless a wide frequency range is desired. Normally R5 is chosen to equal R4; C1 to equal C2; and R3 to equal R6 for maximum dynamic range. However, varying only one of the resistors, R3 or R6, seems to allow ample dynamic range, especially if the filter is installed immediately after the detector in a receiver. A center frequency between 700 and 1500 Hz is usually chosen.

It will also be noted from **fig. 2** that the response is not affected nearly as much near the center frequency when cascading stages as it is far removed from the center frequency. This further illustrates the need for cascading stages to obtain the desired skirt selectivity.

The two circuits shown in **fig. 1** illustrate the input stage, **A**, which includes the pots for Q and gain adjustment and the following additional cascaded stages, **B**, each of which uses a pot for tuning to center frequency. The op amps, with their near-infinite input impedance





The experimental quad op amp active bandpass filter showing component layout, above, and wiring, below, using Vector board construction for flexibility and low cost.

and near-zero output impedance, provided excellent isolation between filter stages in the feedback loop. In this case, the noninverting inputs were connected to ground along with the center common point of the two bias sources of plus and minus 9 volts.

Very weak signals could be pulled out of the noise by increasing the first-stage filter gain and Q. The filter exhibits an increase in weak-signal level when the gain and Q are adjusted carefully by presenting energy near the center frequency of the filter, which causes it to ring slightly. Filter ringing increases the desired signal output level and provides an apparent signal-processing gain, which strengthens the lower signal voltage levels. This does not mean that energy is created, but that energy is added to the amount detected.

It was not the intent of this article to describe each filter in great detail but rather to present an introduction and the supporting design equations. Many other configurations are possible, but I feel that those discussed here are of primary interest. At any rate, you can use the design equations for your own audio bandpass filter and tailor it to your requirements.

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2. J.G. Graeme, G.E. Tobey, and L.P. Huelsman, *Operational Amplifier Design and Applications,* 1973 Burr-Brown Electronic Series, McGraw-Hill, New York.
3. P.E. Fleischer and J. Tow, "Design Formulas for Biquad Active Filters Using Three Operational Amplifiers," *Proceedings of the IEEE* (letter), May, 1973, page 662.
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ham radio

ten commandments for technicians

I Beware the lightning that lurketh in the undischarged capacitor, lest it cause thee to bounce upon thy buttocks in a most untechnician-like manner.

II Cause thou the switch that supplieth large quantities of juice to be opened and thusly tagged, that thy days in this earthly veil of tears may be long.

III Prove to thyself that all circuits that radiateth and upon which thou worketh are grounded and thusly tagged lest they lift thee to radio frequency potential and causeth thee also to make like a radiator.

IV Tarry not amongst those fools who engageth in intentional shocks for they are surely nonbelievers and are not long for this world.

V Take care that thou useth the proper method when thou takest the measure of a high-voltage circuit lest thou incinerate both thyself and thy meter, for verily, though thou hast no account number and can easily be surveyed, the test meter doth have one and, as a consequence, bringeth much woe unto the supply department.

VI Take care that thou tampereth not with safety devices and interlocks, for this incurreth the wrath of thy supervisor and bringeth the fury of thy safety inspector down upon thy head.

VII Work thou not on energized equipment, for if thou dost, thy fellow workers will surely buy beers for thy widow and console her in other ways.

VIII Service thou not equipment for electrical cooking. It is a slothful process and thou might sizzle in thine own fat for hours upon a hot circuit before thy Maker sees fit to end thy misery.

IX Trifle thou not with radioactive tubes and substances lest thou commence to glow in the dark like a lightning bug and thy wife have no further use for thee except thy wages.

X Thou shalt not make unauthorized modifications to equipment, but causeth thou to record all field changes and authorized modifications made by thee, lest thy successor tear his hair out and go slowly mad in his attempt to decide what manner of creature hath made a nest in the wiring of such equipment.

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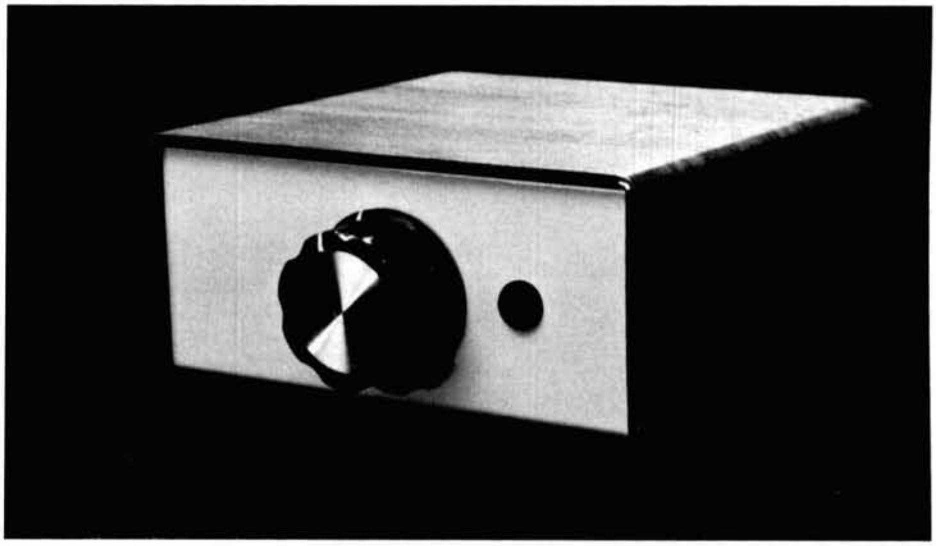


wideband preamp

Construction details for a low noise, wideband receiver preamp that covers the frequency range from 1000 kHz to 50 MHz

Here's an effective yet inexpensive wideband preamp that covers the amateur high-frequency bands. Use it to soup up old receivers of the pre-1970 era or improve performance of more modern equipment. The circuit is easy to build, requires no exotic parts, and should go together in one or two evenings. Total parts cost is less than ten dollars.

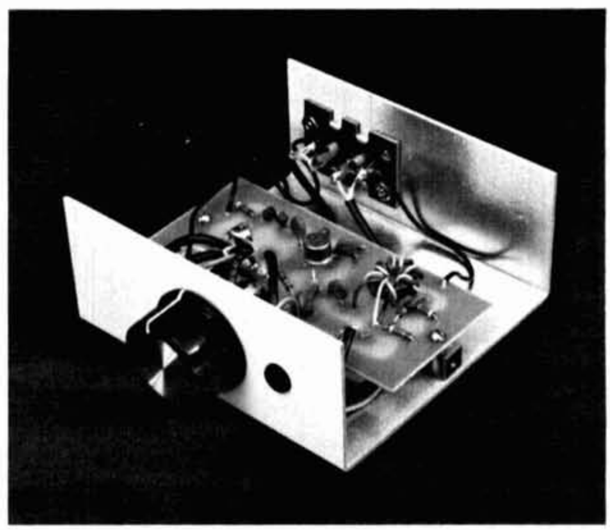
The circuit (**fig. 1**) features a 2N5109 npn silicon high-frequency transistor, which is used in CATV amplifiers requiring low cross-modulation distortion and low-noise input. The 2N5109 has a 3-dB noise figure at 200 MHz, current gain-bandwidth product of 1200 MHz, cross-modulation distortion of −70dB, and collector-emitter amplifier voltage gain of 11 dB (50-216 MHz). It sells for about $3.00 in small quantities. A PC-board layout is shown in **fig. 2**.

While the circuit was intended to be used in the frequency range from 1 to 50 MHz, the 2N5109 is a hot performer well into the vhf range. However, the coupling and bypass capacitors, balun transformer, and circuit layout may need to be changed to reach the higher frequencies. In certain cases, as when using a wideband antenna or equipment with poor intermodulation or spurious-response characteristics, it may be necessary to use bandpass filters[1] on the amplifier input or output.

By Ed Pacyna, W1AAZ, Danbury Circle, Amherst, New Hampshire 03031

acknowledgement

I would like to acknowledge the contributions of Jim Jenkins, K1JXK, who's early ideas led to my writing this article.



The printed-circuit board for the wideband rf preamplifier is built into a 4x4x1½ inch (10x10x3.8cm) aluminum enclosure. LED is mounted inside rubber grommet to the right of the in/out switch knob.

reference

1. Wes Hayward, W7ZOI, "Bandpass Filters for Receiver Preselectors," *ham radio,* February, 1975, page 18.

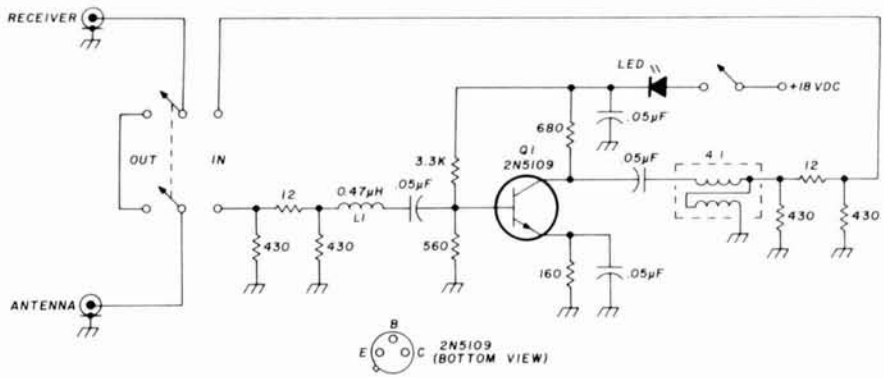


fig. 1. Wideband preamp schematic. The .47 µH inductor can be a couple of ferrite beads placed over a short loop of no. 20 to no. 30 AWG (0.8-0.25mm) insulated wire. The 4:1 wideband transformer in the transistor collector circuit is made by twisting together two pieces of insulated 20 to 30 AWG (0.8-0.25mm) wire, then winding 8 to 10 bifilar turns on a ¼ to ½ inch (6-13mm) toroid core.

fig. 2. Foil side of the printed-circuit layout for the wideband amplifier.

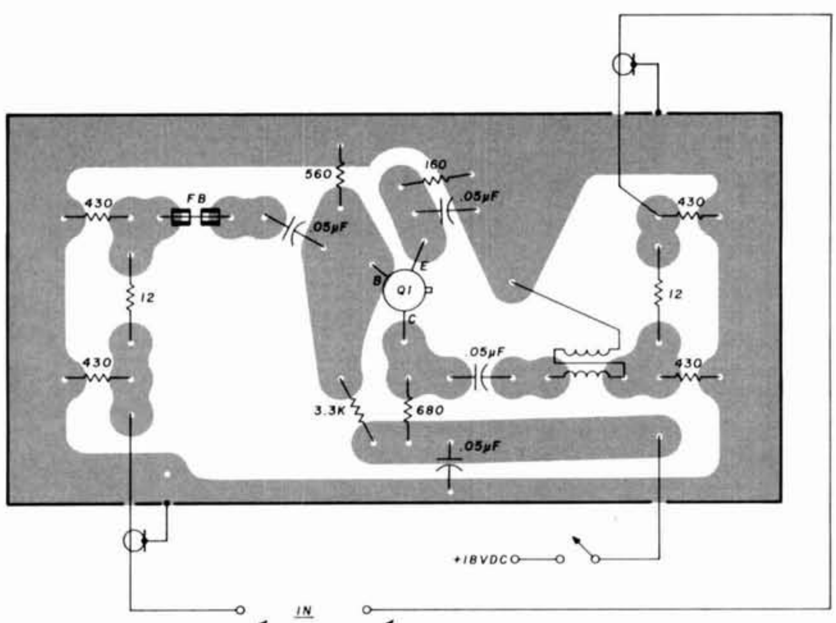


fig. 3. Component layout for the wideband high-frequency preamplifier.

More Details? CHECK-OFF Page 134

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four-band vhf receiving converter

One 46-MHz crystal and an ingenious mixing scheme provide coverage between six meters and 420 MHz

How would you like an inexpensive way to monitor your local two-meter repeater? Or how about a 220-MHz receiver? In fact, how about a receiver covering 6 meters through 420 MHz? These are a few of the questions that were buzzing around in my head, so I decided to build a general-purpose receiver capable of covering all vhf bands. The result is the front end of an all-band vhf converter.

design

At a recent test-equipment seminar, a nifty but little-used fact concerning mixers surfaced once again: Not only is the local oscillator's fundamental frequency available; but the mixer itself, being a nonlinear device, produces harmonics of the oscillator frequency that can be used in the mixing action.

Example: Use a 46-MHz local oscillator and an input of 51 MHz to generate an i-f at 5 MHz. The local oscillator's 138-MHz third harmonic can produce the same i-f from an input at 143 MHz. In fact, this i-f will be produced from any of the following input frequencies: 87 MHz, 97 MHz (2 x 46 = 92); 133 MHz, 143 MHz (3 x 46 = 138); 179 MHz (4 x 46 = 184); 179 MHz, 189 MHz (4 x 46 = 184); and so on.

Now, 46 MHz is not just a number plucked out of thin air for this example. It was developed after weeks of doodling, and is a key ingredient in the design. To understand why, consider this:

1 x 46 = 46	(used on 6 meters)
3 x 46 = 138	(used on 2 meters)
5 x 46 = 230	(used for 220 MHz)
7 x 46 = 322	(not used)
9 x 46 = 414	(used for 420 MHz)

Therefore, it should be possible to use only one crystal and some sort of odd-harmonic generator to meet the design objective of all-vhf-band coverage. 420 MHz is picked up as a bonus (vhf ends at 300 MHz).

circuit description

Fig. 1 is the complete converter schematic with parts list. Ignoring the broadband rf preamplifier (Q1 through Q3) for the moment, let us continue our study of the harmonic mixer. Crystal Y1, a plated, overtone unit, oscillates at 46 MHz in a series-resonant mode. Transistors Q5 and Q6 form the differential amplifier oscillator: rf voltage at Q5's collector is fed back to Q6's base by capacitor C5 and the crystal.

Harmonic emphasis is supplied by tuned line L1 connected from Q6's base to rf ground. L1 is a shorted, quarter-wave transmission line cut to resonance at the crystal frequency. It is 3/4 wavelength long at the third harmonic, 5/4 wavelength long at the fifth harmonic, etc., and acts as a parallel-tuned circuit at all odd harmonics. Therefore, the fundamental and odd harmonics of the crystal are present at the base of mixer transistor Q6, as desired. Construction details for a miniaturized quarter-wave line suitable for PC-board mounting are given in **fig. 2**, which is self-explanatory.

The remainder of the converter consists of a broad-

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band, cascode preamplifier (transistors Q1, Q2, and Q3) that is R-C coupled to mixer driver Q4. No tuning is incorporated purposely so that the converter may be conveniently used on any of the four bands. Of course, external tuning is required, or the device will work on all bands at once! It also converts local TV and fm frequencies, because even harmonics of the crystal are not eliminated in the design.

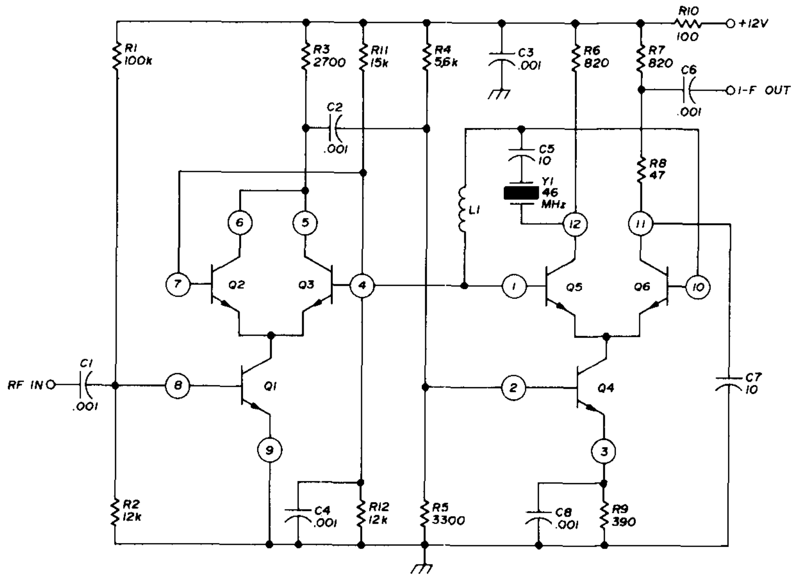
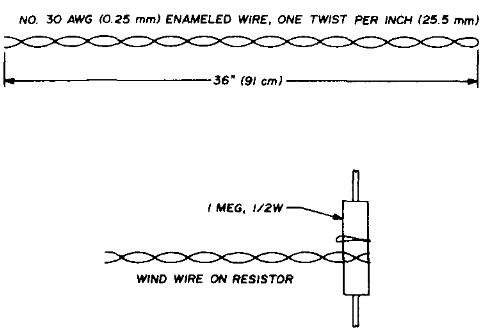


fig. 1. Schematic of the four-band vhf receiving converter. Resistors are ½ watt, 10% composition. All transistors are part of RCA CA3049T integrated circuit. Circled numbers are IC pin numbers. Capacitor voltage rating is 25 volts working or greater. Construction of transmission line resonator, L1, is shown in fig. 2.

construction

Fig. 3A is the foil pattern I used. A glass/epoxy board, 2 by 2 inches (51x51mm) is used. Fig. 3B is a parts layout diagram. Note that all components mount on top of the board; the foil is the bottom. By reference to figs. 2 and 3, you should have no difficulty building one of these converters.



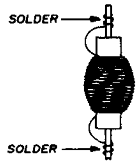


fig. 2. Construction details for a miniaturized ¼-wave transmission line suitable for PC-board mounting.

The six transistors, in two differential amplifier configurations, are packaged as RCA's CA3049T integrated circuit. This IC was reportedly designed for low-noise vhf amplifier service and was chosen for its small size and cost.

performance

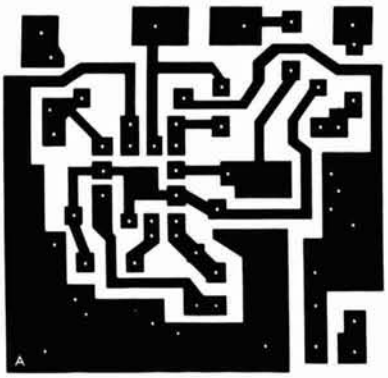
Several converters I built had sensitivities below 1 μV on 50, 144, and 220 MHz, and sensitivities of about 2 μV on 420 MHz. Image rejection and i-f feedthrough, being functions of the input tuning, are not meaningfully measured on such a broadband device. These parameters vary according to the tuning scheme used. Fig. 4 is a simple tuner that serves the purpose and matches the converter input to a 50-ohm antenna. Note that, with a bit of care, this input tuner could be bandswitched.

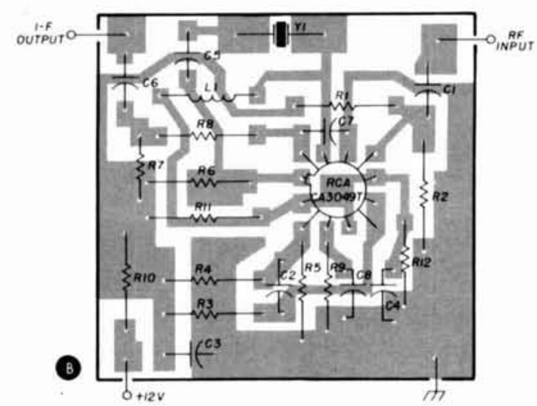
Conversion gain is a function of the band in use. The output level obtained during sensitivity checks, at 10 dB s/n ratio, is 10 μV. Power required is 12 Vdc at approximately 15 mA; the circuit works well on a 9-volt supply with slightly reduced sensitivity.

applications

I use one converter to monitor 146.76 and 146.94 MHz on a broadcast-band transistor radio. To do this, the crystal was changed to 48.5 MHz, achieving an injection frequency of 3 x 48.5 = 145.5 MHz. The resulting i-f is in the broadcast band, and the segment of two meters between 146 and 147 MHz can be tuned by tuning the broadcast receiver through the i-f range of 0.5 to 1.5 MHz. For 146.76 MHz, the i-f converts to 1260 kHz, while 146.94 MHz converts to 1440 kHz. Another common vhf fm frequency, 146.52, converts to 1020 kHz.

fig. 3. Foil side of the PC board, A, and parts layout, B. Note that all components are mounted on top of the board; foil is the bottom.





The i-f output can be inductively coupled to a pocket transistor radio or directly coupled to an auto broadcast receiver. See **fig. 5** for suggested coupling methods. In either case, fm can be recovered quite successfully by slope detection. When using a pocket transistor set, interference from broadcast stations can be removed by wrapping the radio in aluminum foil. Be sure to put the rf choke of **fig. 5** *inside* the foil and as close as possible to the radio's Loop-Stick antenna.

Another of the converters is used to monitor the local repeater output at 146.76 MHz on my Drake R-4A receiver. Again, a crystal other than 46 MHz is used, so that the i-f can be tuned with the receiver. In this case, a 46.5-MHz crystal yields an i-f of 7.460 MHz, which is in range of the Drake's 40 meter band. Once again, slope detection is used to recover fm; some distortion is experienced because of the receiver's selectivity, but no trouble is experienced copying traffic.

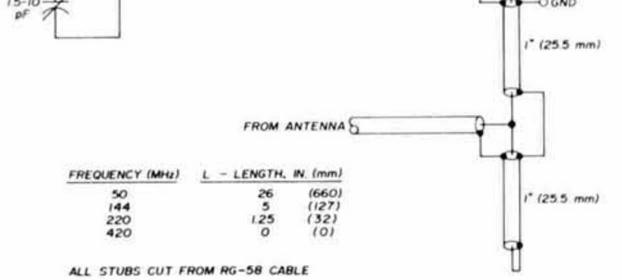
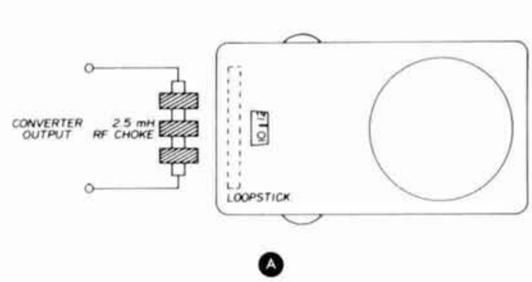


fig. 4. Simple tuner for reducing feedthrough and images.

Yet another unit is in use converting both 15- and 10-meter signals to an i-f in the range of a surplus BC-348 receiver. A crystal frequency of 25 MHz was chosen. It was necessary to replace the tuned line, L1, with a parallel-resonant circuit at the crystal frequency.

The converter may also be used as a general-coverage vhf/uhf receiver. Any i-f up to 35 MHz may be used without modification of the original circuit, and other intermediate frequencies could be adapted by modification of the output stage.

As all crystal harmonics are present in the mixer, a frequency in the 25 to 1000 MHz range can be converted to an i-f of less than 35 MHz. As an example, using



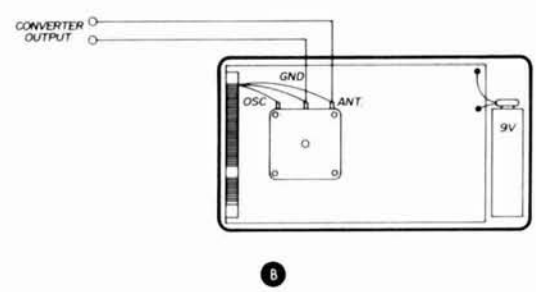


fig. 5. Methods for coupling the vhf converter to a pocket transistor broadcast-band radio, A, or an automobile broadcast-band set, B. Transistor radio and rf choke should be wrapped in aluminum foil to eliminate BC-band interference.

the second harmonic of 46 MHz (92 MHz), 121.5 MHz would be converted to 29.5 MHz. Using the third harmonic (138 MHz), this same signal would be found at 16.5 MHz. Thus, the converter can be used to monitor aviation, police, business, and other frequencies. Again, the tuned input circuit should be resonated to the desired frequency. Problems experienced when the signal frequency falls close to or on a crystal harmonic can be circumvented by using a second crystal differing in frequency from the 46-MHz standard, so that low i-fs are avoided.

Now there's little excuse for not monitoring the higher-frequency amateur bands. This project can be built in a few evenings, and for less than $20.00. It's inexpensive, simple, and it works. See you on vhf.

ham radio

Ham Radio Magazine

October 1976

Page numbers 67, 68, 69 and 70
were skipped in this issue.
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phasing-type single-signal detector

Phase-shift network design and construction data for improving receivers with marginal selectivity

Phasing-type detectors, despite their superficial complexity, are the only means of avoiding audio images in a direct-conversion receiver. Likewise if a superheterodyne receiver has inadequate selectivity or is plagued by broad-band i-f noise,[1,2] the cure lies in either additional i-f filters or in a phasing detector.

Although the circuits described here would make a good direct-conversion receiver, they were designed for use in a superhet with lots of i-f noise and inadequate stop-band rejection. While this system can't offer all the advantages of a second crystal filter, it does give substantial improvement at moderate cost. (Phase-shift networks are described in the appendix for unwanted sideband suppressions of at least 30, 37, and 60 dB, depending on the audio bandwidth and circuit complexity. A simple network for CW reception offering 50 dB rejection is also described).

The circuits were chosen for their ease of construction and adjustment. No special LCR precision bridges are needed to set the audio phase network capacitors. If you can tune a musical instrument the only test equipment needed is a grid-dip oscillator and a vtvm. A frequency counter and oscilloscope are useful but not absolutely required. The system costs between $5 and $25 more than a simple bfo-product detector combination depending on how good a scrounger you are. (Compare this price with the cost of crystal filters!) The phasing detector block diagram is shown in **fig. 1**.

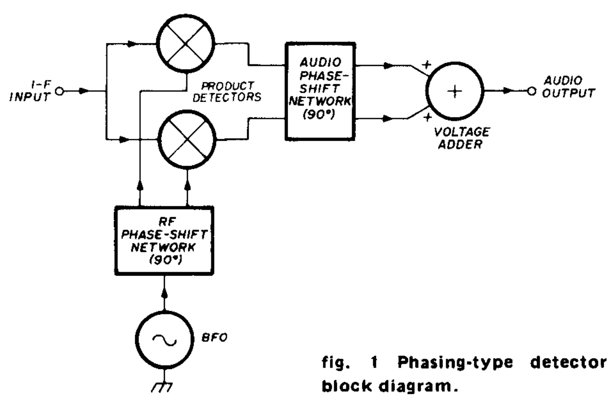

fig. 1 Phasing-type detector block diagram.

By Don Lawson, WB9CYY, 4601 Jay Drive, Madison, Wisconsin 53704

The audio phase-shift networks described here are also suitable for use in ssb phasing-type transmitters. For such applications wire the two inputs (C and D of **fig. 3**) together and run the outputs (TP1 and TP2) to the balanced modulators.

bfo and product detector

The bfo and product detector schematic is given in **fig. 2.** L2 and C2 resonate at the bfo frequency; R1 is should be chosen so that the combined drain current of both mosfets is equal to the average of their individual values of I_{Dss} so that about 9 volts are between the source and drain of each transistor (whichever condition gives the larger value of R4).

R2 and R3 can have values anywhere between 470 ohms and 1k and should be closely matched. C4 and C5 should also be matched to 5% or so (using the method described later). L3, L4, C6, and C7 form a lowpass

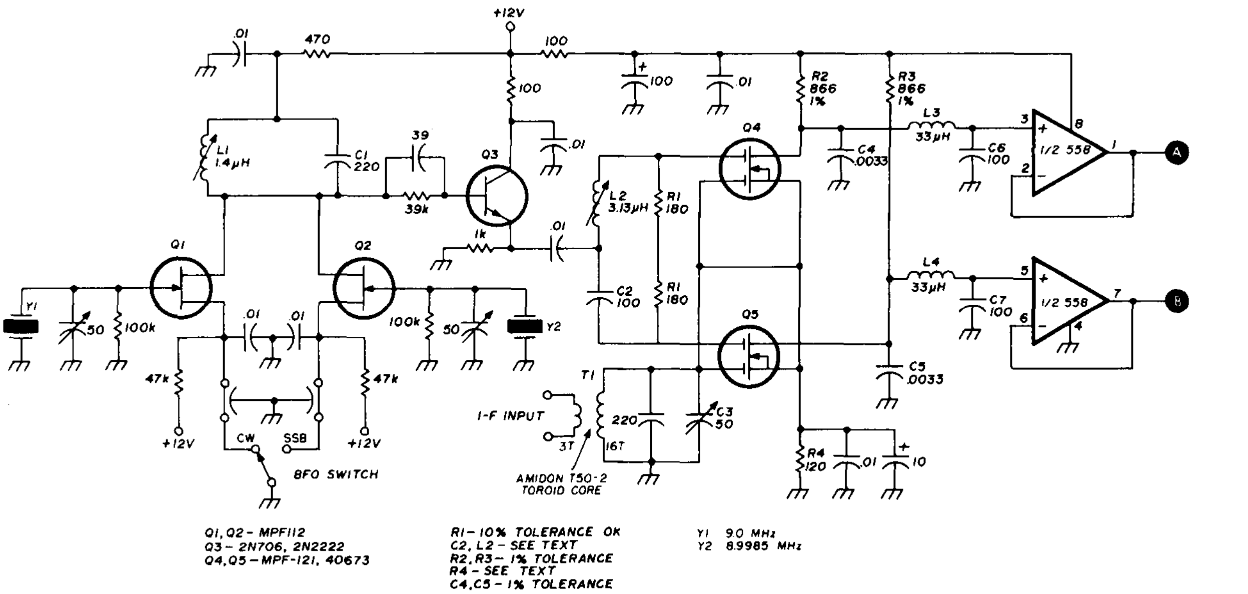

fig. 2. Bfo-product detector schematic. Components that must be closely matched are discussed in the text.

equal to the reactance of the capacitor or inductor at resonance. For an i-f of 9 MHz, representative values are R1 = 180, C2 = 100 pF, and L2 = 3.13 μH.

The bfo uses about the simplest method I know for switching parallel-resonant crystals, and with MPF-112s at 35 cents apiece, it's fairly inexpensive. The product detector's dual gate mosfets, Q4 and Q5, can be MPF-121s or 40673s. If you buy three of four mosfets (instead of just two), you can be fairly sure of finding two that have similar values of V_{gsoff} and I_{Dss}. R4 filter to prevent the bfo signal from getting into the op-amps. The bfo-product detector assembly should be built into a shielded enclosure to keep the bfo signal out of the i-f strip.

audio circuits

The audio phase-shift network is shown in **fig. 3.** There are simpler circuits in the literature but this one is by far the easiest to build. C1, C2, C3, and C4 should be set to equal values by padding with small capacitors in

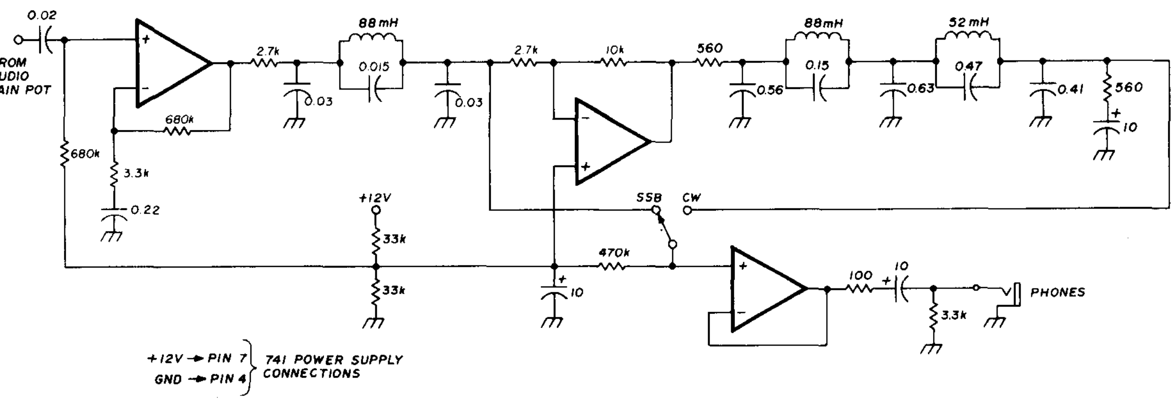

fig. 4. Audio amplifier and filter schematic. Cutoff frequencies are respectively 3 kHz and 800 Hz for the single- and double-section filters.

parallel. Several 50, 100, and 200 pF capacitors are handy for this. If you want to use some other value for the capacitors, multiply R1 through R4 by your capacitor value divided by 0.022; i.e., for 0.1 μF, R1 would be 23.03k x 0.1/0.022 or 104.7k. The phase-shift network is designed to work from 300 to 4000 Hz and theoretically provides at least 31 dB suppression. For this circuit, R5 through R8 form voltage dividers. In the circuit of **fig. 3** they have a ratio of resistances of 14.94 to 1.

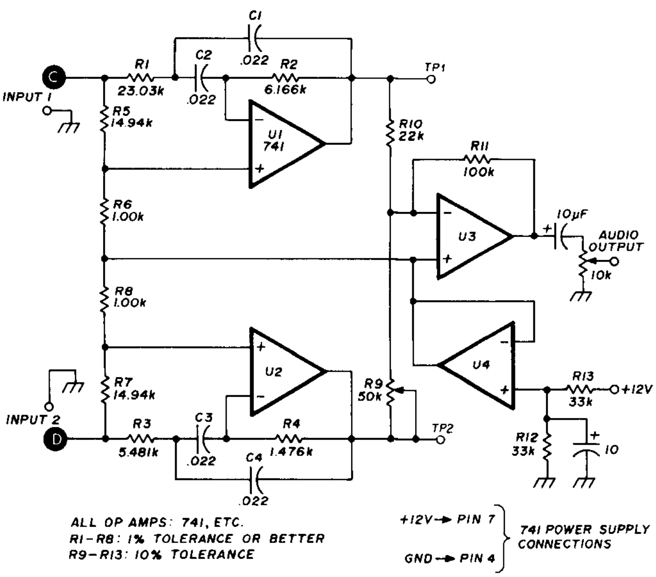

fig. 3. Audio phase-shift network schematic. C1-C4 should be matched by adding small capacitors in parallel.

R9, R10, R11, and U3 form the voltage adder of **fig. 1.** U4 gives a good ac ground for the networks. Trying to substitute a large bypass capacitor for U4 isn't good enough. To eliminate U4 you'd need dual power supplies for U1 through U3, so that dc ground is also a suitable ac ground. The audio output can be applied to an amplifier with sharp filters; a typical circuit is shown in **fig. 4.** The single-section filter is a 3-kHz elliptical lowpass; the two-section filter has a cutoff frequency of 800 Hz for CW reception.

construction and alignment

A grid-dip oscillator is all that's necessary for building and checking the bfo and product detector. If an oscilloscope is available, it might be a good idea at this point to set L2 initially for the point where the signals at A and B **(fig. 2)** are 90 degrees apart in phase. (Use the circuit of **fig. 5A** if a scope with a triggered sweep is available; use **fig. 5B** if one is not).

All capacitors in the audio phase-shift network should be matched as closely as possible and R1-R4 should be proportional to their calculated values. The ratios of R5/R6; R7/R8 should also be as close as possible to their computed ratios. For example, if all your capacitors were equal to 0.019 μF and R1-R4 were 0.9 times their computed values, the only bad effect would be that

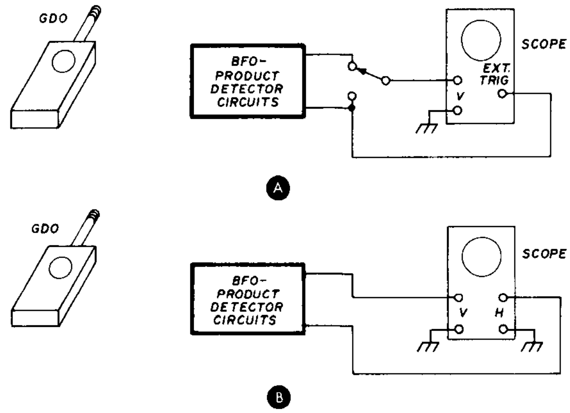

fig. 5. Setups for checking and presetting the rf phase-shift network. Methods (A) and (B) are for oscilloscopes with and without triggered sweep.

the network frequency range would be shifted slightly. The resistors can be made up from parts on hand, using an accurate volt-ohmmeter, or 1% wirewound resistors may be used.

The capacitors can be matched quite well by using the oscillator circuit of **fig. 6.** Be sure to use an uncom-

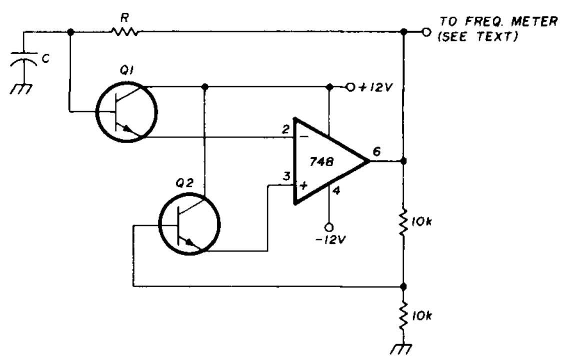

fig. 6. Oscillator circuit for measuring capacitance. An uncompensated op-amp should be used. Q1, Q2 are general-purpose audio transistors.

pensated op-amp, such as a 748 or a 709. The internal compensation in a 741 will slow down the oscillator and measurements will be unreliable. Try to keep the frequency between 1 and 10 kHz for best accuracy. If you're going for readings of four significant digits, power the oscillator from either batteries or a regulated supply to eliminate effects of line-voltage variations.

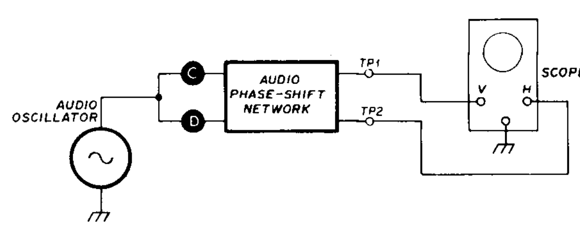

fig. 7. Test circuit for checking the audio frequency phase-shift network.

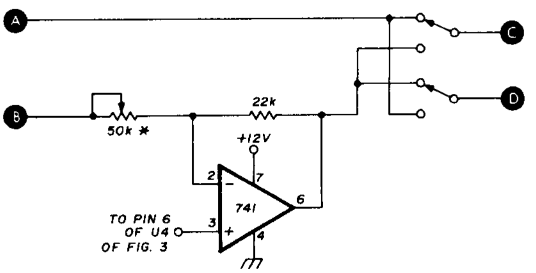

fig. 8. Method for sideband selection through a panel switch. Adjust the 50k pot so that the signal amplitude at pin 6 of the 741 is the same as at point A.

The "frequency meter" should be either a digital frequency counter or a tape recorder. A meter-movement type instrument won't allow sufficient precision. If you use a tape recorder, find which capacitor gives the lowest-pitch tone then pad the other capacitors to give this same frequency. By taping the lowest pitch, you'll save having to switch constantly between capacitors for comparison. If you use a counter, choose one capacitor as a reference and note its frequency. Assuming it equals its marked value exactly, the relative capacitance of the other capacitor is:

$$C_x = \frac{freq}{f_x} \times C_{ref}$$

Unless you're using a very good ohmmeter to measure the resistors, there's little point in matching the capacitors to better than 0.25%. If possible, use high-stability capacitors (polystyrene is best).* If you're using carbon-composition resistors, coat them with epoxy to reduce

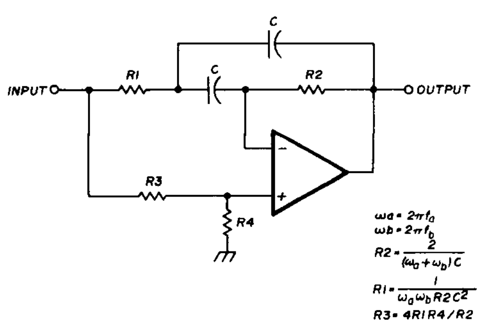

fig. 9. Portion of a phase-shift network with singular frequencies, f_a and f_b. Formulas are used with table 1 to compute component values.

variations of their values with humidity. Use **fig. 7** to check the operation of the phase-shift network. With the scope setup shown you should be able to see a circle on the scope screen.

Referring to **figs. 2** and **3**, in final adjustment first wire A to C and B to D. Sweep the grid-dip oscillator or signal generator past the bfo frequency and decide which sideband (upper or lower) is being rejected. If it's not the sideband you want, try wiring A to D and B to C. (If you want to select sidebands by a panel switch, try using the circuit of **fig. 8**). Now, alternatively adjust L2 of the bfo and R9 of the audio phase-shift network for the best unwanted sideband rejection.

results

If you use the circuit in **fig. 3**, you should be able to easily get about 30 dB of unwanted sideband rejection between 300-4000 Hz. For better rejection, use one of the other circuits in the appendix.

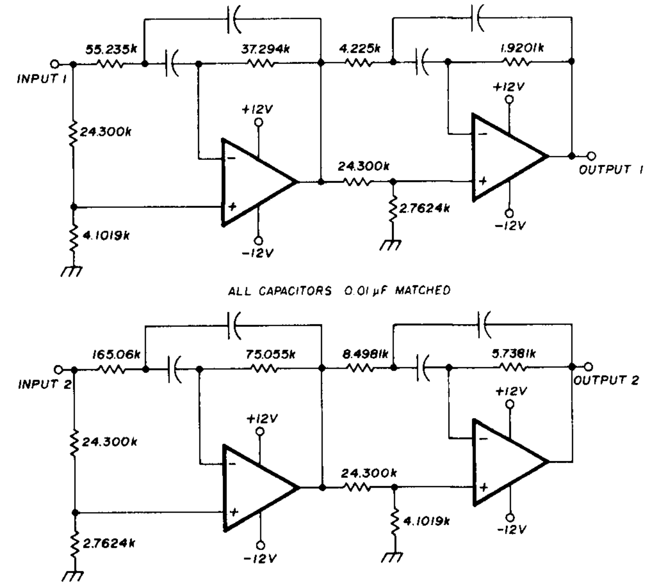

fig. 10. Cascaded filters to obtain an 8-pole network. Circuit provides at least 60 dB unwanted sideband rejection between 200-4000 Hz using 741 op-amps.

I hope this article provides some ideas for those who build or modify receivers. When it comes to improving the selectivity of a marginal receiver, this system is the best for the money.

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*Polystyrene capacitors can be obtained from Weinschenker, Box 353, Irwin, Pennsylvania 15642.

appendix

The design of any phase-shift network is a trade between complexity, bandwidth, and phase accuracy. Phase-shift networks are a special category of "all-pass filters," meaning that the network gain is constant regardless of frequency. Phase-shift networks are characterized by a set of "singular frequencies." The calculation of these frequencies is too involved to be covered here, but an algorithm is in the literature that requires nothing more than a calculator with a square-root key.[3,4] For those who don't like lots of arithmetic, I've compiled the results of several design computations in **table 1.** Network 1 is the general-purpose network shown in **fig. 3,** while network 2 is for more narrow-band ssb use. Network 3 is for use in direct-conversion CW receivers. These networks are all of the four-pole variety since each has four singular frequencies.

table 1. Phase-shift network design data.

network	freq range (Hz)	minimum unwanted sideband rejection (dB)	singular frequencies f_1	f_2	f_3	f_4
1	300-4000	31	148.9	625.8	1916.	8049.
2	300-2500	37	136.2	525.2	1428.	5507.
3	225-950	49	88.7	298.6	672.0	2265

As noted before, the circuit used here was chosen for its simplicity of construction as compared to other circuits.[5] (Unless you have a frequency counter or precision LCR bridge, it's much easier to set capacitors equal than it is to set them to arbitrary ratios). To compute component values from **table 1**, we take either odd- or even-numbered singular frequencies and plug them into the formulas of **fig. 9.**

As an example, let's try designing the network shown in **fig. 3.** As a start, we take odd frequencies of 148.9 and 1916 Hz. Converting to radians/sec, we get $\omega_1 = 935.6$ and $\omega_3 = 12{,}038$. Assuming $C = 0.025\ \mu F\ (2.5 \times 10^{-8}\ F)$, we compute

$$R2 = \frac{2}{(935.6 + 12{,}038) \times 2.5 \times 10^{-8}} = 6166\ ohms$$

Likewise,

$$R1 = \frac{1}{935.6 \times 12{,}038 \times (2.5 \times 10^{-8})^2 \times 6166}$$
$$= 23{,}040\ ohms = 23.04k$$

Finally, since we have a couple of 1k 1% resistors on hand, we choose R4 = 1k and

$$R3 = \frac{4 \times R1 \times R4}{R2} = \frac{4 \times 23{,}040 \times 1000}{6166}$$
$$= 14{,}946\ ohms = 14.95k$$

We now have half of our network designed. For the other half we follow the same procedure except that we work with f_2 and f_4.

Note that at dc the circuit gain is $R4/(R3 + R4)$. Since this is an all-pass network, all audio frequencies are attenuated by this same amount. Hence, this network attenuates the signal by 24 dB. Because the 741 op amps have internal noise it may be desirable, especially in direct-conversion receivers, to put low-noise amplifiers (such as an LM381) between the product detector and the phase-shift network.

A final note: It's possible to cascade phase-shift network sections to get an 8-pole network. A practical 8-pole network is shown in **fig. 10.** According to a computer simulation, this circuit provides at least 60 dB of unwanted sideband rejection between 200-4000 Hz. This simulation was with 741s. It may be possible to get an additional 5 dB suppression by using 556 op-amps since their input impedance is much higher than that of a 741. Of course, the components would have to be set very accurately to realize this theoretical figure and a preamplifier would be necessary to overcome the network signal attenuation. This circuit would be ideal for a high-quality, phasing-type ssb direct-conversion receiver.

ham radio

7400N TTL

SN7400N	16	SN7459A	25	SN74154N*	1.00
SN7401N	16	SN7460N	22	SN74155N	99
SN7402N	21	SN7470N	45	SN74156N	99
SN7403N	16	SN7472N	39	SN74157N	99
SN7404N*	18	SN7473N*	37	SN74160N	1.25
SN7405N	24	SN7474N*	32	SN74161N	99
SN7406N	20	SN7475N*	50	SN74163N*	99
SN7407N	29	SN7476N*	32	SN74164N	1.10
SN7408N	25	SN7479N*	5.00	SN74165N	1.10
SN7409N	25	SN7480N	50	SN74166N	1.25
SN7410N*	18	SN7482N	98	SN74167N	5.50
SN7411N	30	SN7483N	70	SN74170N	2.10
SN7412N	33	SN7485N	89	SN74172N	8.95
SN7413N	45	SN7486N	39	SN74173N	1.50
SN7414N	70	SN7488N	3.50	SN74174N	1.25
SN7416N	35	SN7489N	2.25	SN74175N	99
SN7417N	35	SN7490N*	45	SN74176N	90
SN7420N	21	SN7491N	75	SN74177N	90
SN7421N	33	SN7492N	49	SN74180N	99
SN7422N*	49	SN7493N*	49	SN74181N	2.49
SN7423N	37	SN7494N	79	SN74182N	95
SN7425N	29	SN7495N	79	SN74184N	1.95
SN7426N	29	SN7496N	89	SN74185N	2.20
SN7427N	37	SN7497N*	4.00	SN74186N*	5.00
SN7429N	42	SN74100N*	1.00	SN74187N	6.00
SN7430N	26	SN74107N	39	SN74188N	3.95
SN7432N	31	SN74121N*	39	SN74190N	1.19
SN7437N	27	SN74122N	39	SN74191N	1.25
SN7438N	27	SN74123N*	70	SN74192N*	89
SN7439N	25	SN74125N	60	SN74193N*	89
SN7440N	15	SN74126N	60	SN74194N	1.25
SN7441N	89	SN74132N	1.09	SN74195N	75
SN7442N	59	SN74136N	95	SN74196N	1.25
SN7443N	75	SN74141N	1.15	SN74197N	75
SN7444N	75	SN74142N*	4.00	SN74198N	1.75
SN7445N	75	SN74143N*	4.50	SN74199N	1.75
SN7446N	81	SN74144N*	4.50	SN74200N	5.59
SN7447N*	69	SN74145N	1.15	SN74279N*	90
SN7448N	79	SN74147N	2.35	SN74251N	1.79
SN7450N	26	SN74148N	2.00	SN74284N	6.00
SN7451N	27	SN74150N	1.00	SN74285N	6.00
SN7453N	27	SN74151N	79	SN74367N	75
SN7454N	20	SN74153N	89		

MANY OTHERS AVAILABLE ON REQUEST
20% Discount for 100 Combined 7400's

CMOS

CD4000	25	CD4035	1.85	74C04N	75
CD4001	25	CD4040	2.45	74C10N	65
CD4002	25	CD4042	1.90	74C20N	65
CD4006	2.50	CD4044	1.50	74C30N	65
CD4007	25	CD4046	2.51	74C42N	2.15
CD4009	59	CD4047	2.75	74C73N	1.50
CD4010	59	CD4049	79	74C74	1.15
CD4011	25	CD4050	79	74C90N	3.00
CD4012	25	CD4051	2.95	74C95N	2.00
CD4013	47	CD4053	2.95	74C107N	1.25
CD4016	56	CD4060	3.25	74C151	2.90
CD4017	1.35	CD4066	1.75	74C154	4.00
CD4019	55	CD4069	45	74C157	2.15
CD4020	1.49	CD4071	45	74C160	3.25
CD4022	1.25	CD4081	45	74C161	3.25
CD4023	25	CD4511	2.50	74C163	3.00
CD4024	1.50	CD4518	2.50	74C164	3.25
CD4025	25	74C00N	39	74C173	2.60
CD4027	69	74C02N	55	74C193	2.75
CD4028	1.65			74C195	2.75
CD4029	2.90			MC4044*	4.50
CD4030	65			MC14016*	56

LINEAR

LM300H	80
LM301H	35
LM301CN	35
LM302H	75
LM304H	1.00
LM305H	95
LM307CN	35
LM308H	1.00
LM308CN	1.00
LM309H	1.10
LM309K	99
LM310CN	1.15
LM311H	90
LM311N	90
LM318CN	1.50
LM319N	1.30
LM320K-5	1.35
LM320K-5.2	1.35
LM320K-12	1.35
LM320K-15	1.35
LM320T-5*	1.75
LM320T-5.2	1.75
LM320T-8*	1.75
LM320T-12*	1.75
LM320T-15	1.75
LM320T-18*	1.75
LM320T-24	1.75
LM323K-5*	9.95
LM324N	1.80
LM339N	1.70
LM340K-5	1.95
LM340K-6	1.95
LM340K-8	1.95
LM340K-12	1.95
LM340K-15	1.95
LM340K-18	1.95
LM340K-24	1.95
LM340T-5	1.75
LM340T-6	1.75
LM340T-8*	1.75
LM340T-12	1.75
LM340T-15	1.75
LM340T-18*	1.75
LM340T-24	1.75
LM350N	1.00
LM351CN	65

LM370N	1.15
LM373N	3.25
LM377N	4.00
LM380N	1.39
LM390CN	1.05
LM381N	1.79
LM382N	1.79
NE501K	8.00
NE510A	6.00
NE531H	3.00
NE536T	6.00
NE540L	6.00
NE550N	79
NE555V*	45
NE560B*	5.00
NE561B*	5.00
NE562B*	5.00
NE565H*	1.25
NE565N*	1.75
NE566CN*	1.25
NE567H*	1.95
NE567V*	1.50
LM703CN	45
LM709H	29
LM709N	29
LM710N	79
LM711N	39
LM723H	55
LM723H	55
LM733N	1.00
LM739N	1.00
LM741CH	35
LM741CN	35
LM74114N	39
LM747H	79
LM747N	79
LM748H	39
LM748N	39
LM1303N	90
LM1304N	1.19
LM1305N	1.40
LM1307N	85

LM1310N	2.95
LM1351N	1.65
LM1414N	1.75
LM1458C	65
LM1496N	95
LM1556V	1.85
LM2111N	1.95
LM2901N	2.95
LM3065N	69
LM3900N	55
LM3905N	60
LM3909*	1.25
LM5556N	1.85
MC5558V	1.00
LM7525N	90
LM7535N	1.25
8038B*	4.95
LM75450	49
75451CN	39
75452CN	39
75453CN	39
75454CN	39
75491CN	79
75492CN	89
75494CN	89
RCA LINEAR	
CA3013	2.15
CA3023	2.56
CA3035	2.48
CA3039	1.35
CA3046	1.30
CA3059	3.25
CA3060	3.25
CA3080	85
CA3081	2.00
CA3082	2.00
CA3083	1.60
CA3086	85
CA3089	3.75
CA3091	10.20
CA3123	2.15
CA3130	1.39
CA3140	1.25
CA3600	1.75
RC4194	5.95
RC4195	3.25

74LS00 TTL

74LS00	39	74LS55	.39	74LS151	1.55
74LS02	39	74LS73	65	74LS153	1.89
74LS03	39	74LS74	65	74LS157	1.55
74LS04	45	74LS75	79	74LS162	2.25
74LS05	45	74LS76	65	74LS163	2.25
74LS08	39	74LS83	2.19	74LS164	2.25
74LS10	39	74LS86	65	74LS181	3.69
74LS13	79	74LS90	1.25	74LS190	2.85
74LS14	2.19	74LS92	1.25	74LS191	2.85
74LS20	39	74LS93	1.25	74LS192	2.85
74LS26	49	74LS95	2.19	74LS193	2.85
74LS27	45	74LS96	1.89	74LS194	2.25
74LS28	49	74LS107	65	74LS195	2.25
74LS30	39	74LS112	65	74LS257	1.89
74LS32	45	74LS132	1.55	74LS260	55
74LS40	49	74LS136	65	74LS279	79
74LS51	39	74LS138	1.89	74LS670	5.95

DATA HANDBOOKS

7400 Pin-out & Description of 5400/7400 ICS **$2.95**
CMOS Pin-out & Description of 4000 Series ICS **$2.95**
Linear Pin-out & Functional Description **$2.95**
ALL THREE HANDBOOKS $6.95

100 PER STRIP MOLEX PINS

Intended for use as an inexpensive substitute for IC sockets. Also perfect for use as board connectors and in subassemblies.

SPECIAL — 100/1.49 — 1000/12.00

CONSUMER ELECTRONICS

PONG — SINGLE GAME $59.95
SUPER PONG — 4 GAMES IN ONE $89.95

GAMES INCLUDED IN SUPER PONG ARE:
- PONG
- CATCH
- SUPER PONG
- HANDBALL

FEATURES OF PONG AND SUPER PONG

- Incremental speed on volleys increases excitement
- Playing field adjusts to any size screen
- Game appears in color or in black & white, depending on television set
- Unmistakable "PONG" sound accompanies each volley
- Digital scoring flashes on the screen between each point
- 2 player challenge or Solitaire
- Hooks up simply to any model television set: the screen actually becomes the playing field
- English and other techniques can be used to make any member of the family a Pong champion
- Battery operated by 4 size "D" flashlight batteries included with the Unit

AC Adaptor (Eliminates Batteries) **$9.95**

DISCRETE LEDS SPECIAL SALE!

.125" dia.
XC209Red 10/$1
XC209Green 4/$1
XC209Yellow 7/$1
XC209Orange 4/$1

.190" dia.
XC111Red 8/$1
XC111Green 4/$1
XC111Yellow 4/$1
XC111Orange 4/$1

.200" dia.
XC22Red 5/$1
XC22Green 4/$1
XC22Yellow 4/$1
XC22Orange 4/$1
SSL-22RT 8/$1

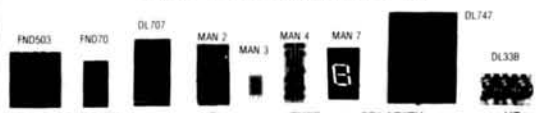
.185" dia.
XC526Red 10/$1
XC526Green 7/$1
XC526Yellow 4/$1
XC526Orange 4/$1
XC526Clear 7/$1

.200" dia.
XC556Red 10/$1
XC556Green 7/$1
XC556Yellow 7/$1
XC556Orange 7/$1
XC556Clear 7/$1

.085" dia.
MV50
.085" dia. Micro red LED
6/$1

DISPLAY LEDS SUPER SAVINGS!

MAN 7 $.99 — DL33B $.99



TYPE	POLARITY	HT		TYPE	POLARITY	HT	
MAN 1	Common Anode	.270	2.49	MAN 3620	Common Anode-orange	.300	.99
MAN 2	5 x 7 Dot Matrix	.300	4.95	MAN 3640	Common Cathode-orange	.300	.99
MAN 3	Common Cathode	.125	3/1.	MAN 4710	Common Anode-Red	.400	.99
MAN 4	Common Cathode	.187	1.95	DL701	Common Anode-red	.300	.99
MAN 7	Common Anode	.300	.99	DL707	Common Anode	.300	1.50
MAN 7G	Common Anode-green	.300	1.95	DL 728	Common Cathode	.500	1.99
MAN 7Y	Common Anode-yellow	.300	1.95	DL 747	Common Anode	.600	2.25
MAN 52	Common Anode-green	.300	.99	DL 750	Common Cathode	.600	2.49
MAN 64	Common Anode-red	.400	.99	DL 338	Common Cathode	.110	.99
MAN 72	Common Anode	.300	.99	FND70	Common Cathode	.250	.50
MAN 74	Common Cathode	.300	1.15	FND503	Common Cathode	.500	1.29
MAN 82	Common Anode-yellow	.300	.99	FND507	Common Anode	.500	1.29
MAN 84	Common Cathode-yellow	.300	.99				

IC SOLDERTAIL — LOW PROFILE (TIN) SOCKETS

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.17	16	15	24 pin	$.38	37	36
14 pin	20	19	18	28 pin	45	44	43
16 pin	22	21	20	36 pin	60	59	58
18 pin	29	28	27	40 pin	63	62	61
22 pin	37	36	35				

SOLDERTAIL STANDARD (TIN)

	1-24	25-49	50-100		1-24	25-49	50-100
14 pin	$.27	25	24	28 pin	$.99	90	81
16 pin	30	27	25	36 pin	1.39	1.26	1.15
18 pin	35	32	30	40 pin	1.59	1.45	1.30
24 pin	49	45	42				

SOLDERTAIL STANDARD (GOLD)

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.30	27	24	24 pin	$.70	63	57
14 pin	35	32	29	28 pin	1.10	1.00	90
16 pin	38	35	32	36 pin	1.75	1.40	1.26
18 pin	52	47	43	40 pin	1.75	1.59	1.45

WIRE WRAP SOCKETS (GOLD) LEVEL #3

	1-24	25-49	50-100		1-24	25-49	50-100
10 pin	$.45	41	37	24 pin	$1.05	95	85
14 pin	39	38	37	28 pin	1.40	1.25	1.10
16 pin	43	42	41	36 pin	1.59	1.45	1.30
18 pin	75	68	62	40 pin	1.75	1.55	1.40

50 PCS. RESISTOR ASSORTMENTS $1.75 PER ASST.

ASST. 1	5 ea.	10 OHM 27 OHM	12 OHM 33 OHM	15 OHM 39 OHM	18 OHM 47 OHM	22 OHM 56 OHM	1/4 WATT 5% — 50 PCS.
ASST. 2	5 ea.	68 OHM 180 OHM	82 OHM 220 OHM	100 OHM 270 OHM	120 OHM 330 OHM	150 OHM 390 OHM	1/4 WATT 5% — 50 PCS.
ASST. 3	5 ea.	470 OHM 1.2K	560 OHM 1.5K	680 OHM 1.8K	820 OHM 2.2K	1K 2.7K	1/4 WATT 5% — 50 PCS.
ASST. 4	5 ea.	3.3K 8.2K	3.9K 10K	4.7K 12K	5.6K 15K	6.8K 18K	1/4 WATT 5% — 50 PCS.
ASST. 5	5 ea.	22K 56K	27K 68K	33K 82K	39K 100K	47K 120K	1/4 WATT 5% — 50 PCS.
ASST. 6	5 ea.	150K 390K	180K 470K	220K 560K	270K 680K	330K 820K	1/4 WATT 5% — 50 PCS.
ASST. 7	5 ea.	1M 2.7M	1.2M 3.3M	1.5M 3.9M	1.8M 4.7M	2.2M 5.6M	1/4 WATT 5% — 50 PCS.

5 EACH MINIMUM PER VALUE ON ANY RESISTORS FROM 2.2 OHM TO 5.6M

5-25 PCS .05 ea. 30-95 PCS .04 ea. 100-495 PCS .03 ea. 500-995 .0275 ea.

SINGLE TURN ⅜" CERMET POTENTIOMETER

- Resistance Tolerance ± 20%
- High Power — 0.5 Watt at 70°C
- Wide Operating Temp. Range −55°C to +125°C

$.89

STANDARD RESISTANCE VALUES

50Ω 100Ω 500Ω 1K 2K 5K 10K 20K 50K 100K 200K 500K 1 MEG

Astrisk Denotes Items On Special For This Month

Satisfaction Guaranteed. $5.00 Min. Order. U.S. Funds.
California Residents — Add 6% Sales Tax — Data Sheets 25c each
Send a 13c Stamp (postage) for a FREE 1976 Catalog

JAMES

1021 HOWARD AVE., SAN CARLOS, CA. 94070
PHONE ORDERS — (415) 592-8097

MINIATURE TOGGLE

DPDT				ANT	1	2-9	10-29	30-99
	ON	OFF	ON	221	2.95	2.55	1.87	1.70
	ON	NONE	ON	223	2.95	2.15	1.58	1.43
	ON	OFF	ON	121	2.35	1.95	1.43	1.30
	ON	NONE	ON	123	2.05	1.65	1.21	1.10

PUSH BUTTON

SPDT — SERIES PB, PB1

Maintained Action Switch
Push To Make Contact And Push Again To Reserve Action

Momentary Action Switch
Normally Open or Normally Closed (For N.C. circuit use terminals 2 & 1; for N.O. 2 & 3; for N.C and N.O. 1, 2 & 3)

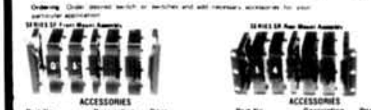
PB-123 $1.75
PB-126 $1.75

5 AMP RATINGS

THUMBWHEEL SWITCHES

THUMBWHEEL SWITCH ONLY

Ordering: Order desired switch or switches and add necessary accessories for your particular application



ACCESSORIES

Part No.	Description	Price
SF EP	End Plate (Pair)	$.50
SF DP	Divider Plate (each)	.40
SF BB	Blank Body (each)	.40
SF HB	Half Body (each)	.40

ACCESSORIES

Part No.	Description	Price
SR EP	End Plate (Pair)	$.50
SR DP	Divider Plate (each)	.40
SR BB	Blank Body (each)	.40
SR HB	Half Body (each)	.40

CLIPLITE!! 4/$1.00

NEW LED MOUNTING SYSTEM

- CLIPLITE mounts from the front of the Panel in a .250 hole on 3/8" centers. Panel thickness from 1/16" to 1/8".
- CLIPLITE equalizes and increases the brightness of commonly used wide beam LEDS.
- CLIPLITE is to be used with XC556 LEDS only. Specify red, green, amber or yellow CLIPLITES when ordering.

ZENERS—DIODES—RECTIFIERS

TYPE	VOLTS	W	PRICE	TYPE	VOLTS	W	PRICE
1N746	3.3	400m	4/1.00	1N4005*	600 PIV	1 AMP	10/1.00
1N751A	5.1	400m	4/1.00	1N4006*	800 PIV	1 AMP	10/1.00
1N752	5.6	400m	4/1.00	1N4007*	1000 PIV	1 AMP	10/1.00
1N753	6.2	400m	4/1.00	1N3600	50	200m	6/1.00
1N754	6.8	400m	4/1.00	1N4148	75	10m	15/1.00
1N959	8.2	400m	8/1.00	1N4154	35	10m	12/1.00
1N965B	15	400m	4/1.00	1N4734	5.6	1w	28
1N5232	5.6	500m	28	1N4735	6.2	1w	28
1N5234	6.2	500m	28	1N4736	6.8	1w	28
1N5235	6.8	500m	28	1N4738	8.2	1w	28
1N5236	7.5	500m	28	1N4742	12	1w	28
1N456	25	40m	6/1.00	1N4744	15	1w	28
1N458	150	7m	6/1.00	1N1183	50 PIV	35 AMP	1.60
1N485A	180	10m	5/1.00	1N1184	100 PIV	35 AMP	1.70
1N4001	50 PIV	1 AMP	12/1.00	1N1185	150 PIV	35 AMP	1.70
1N4002	100 PIV	1 AMP	12/1.00	1N1186	200 PIV	35 AMP	1.80
1N4003	200 PIV	1 AMP	12/1.00	1N1188	400 PIV	35 AMP	3.00
1N4004	400 PIV	1 AMP	12/1.00				

SCR AND FW BRIDGE RECTIFIERS

C36D	15A @ 400V	SCR	$1.95
C38M	35A @ 200V	SCR	1.95
2N2328	1.6A @ 200V	SCR	50
MDA 980-1	25A @ 50V	FW BRIDGE REC.	1.95
MDA 980-3	25A@ 200V	FW BRIDGE REC.	1.95

TRANSISTORS

MPS A05	5/$1.00	PN3567	3/$1.00	PN4249	4/$1.00
MPS A06	5/$1.00	PN3568	4/$1.00*	PN4250	4/$1.00
2N2219A	3/$1.00	PN3569	4/$1.00	2N4400	4/$1.00
2N2221	4/$1.00	2N3704	5/$1.00	2N4401	4/$1.00
2N2222A	5/$1.00	2N3705	5/$1.00	2N4402	4/$1.00
2N2369	5/$1.00	2N3706	5/$1.00	2N4403	4/$1.00
2N2369A	4/$1.00	2N3707	5/$1.00	2N4409	5/$1.00
FN2415	5/$1.00	2N3711	5/$1.00	2N5086	4/$1.00
2N2484	4/$1.00	2N3724	$.65	2N5087	4/$1.00
2N2906A	4/$1.00	2N3725	$1.00	2N5088	4/$1.00
2N2907A	5/$1.00	2N3903	5/$1.00	2N5089	4/$1.00
2N2925	5/$1.00	2N3904	4/$1.00	2N5129	5/$1.00
2N3053	2/$1.00	2N3905	4/$1.00	2N5138	5/$1.00
2N3055	$.89	2N3906	4/$1.00	2N5139	5/$1.00
MJE3055	$1.00	2N4013	3/$1.00	2N5209	5/$1.00
MJE2955	$1.25	2N4014	3/$1.00	2N5951	5/$1.00
2N3392	5/$1.00	2N4123	10/$1.00	C106B1SCR	2/$1.00
2N3398	5/$1.00			2N5432	$2.00

CAPACITOR CORNER

50 VOLT CERAMIC DISC CAPACITORS

	1-9	10-49	50-100		1-9	10-49	50-100
10 pf	05	04	03	001µF	05	04	035
22 pf	05	04	03	0047µF	05	04	035
47 pf	05	04	03	01µF	05	04	035
100 pf	05	04	03	022µF	06	05	04
220 pf	05	04	03	047µF	06	05	04
470 pf	05	04	035	1µF	12	09	075
100 VOLT MYLAR FILM CAPACITORS							
001mf	12	10	07	022mf	13	11	08
0022	12	10	07	047mf	21	17	13
0047mf	12	10	07	1mf	27	23	17
01mf	12	10	07	22mf	33	27	22
+20% DIPPED TANTALUMS (SOLID) CAPACITORS							
1/35V	28	23	17	1.5/35V	30	26	21
15/35V	28	23	17	2.2/25V	31	27	22
22/35V	28	23	17	3.3/25V	31	27	22
33/35V	28	23	17	4.7/25V	32	28	23
47/35V	28	23	17	6.8/25V	36	31	25
68/35V	28	23	17	10/25V	40	35	29
1.0/35V	28	23	17	15/25V	63	50	40
MINIATURE ALUMINUM ELECTROLYTIC CAPACITORS							
Axial Lead				**Radial Lead**			
47/50V	15	13	10	47/25V	15	13	10
1.0/50V	16	14	11	47/50V	16	14	11
3.3/50V	15	13	10	1.0/16V	15	13	10
4.7/25V	16	14	12	1.0/25V	16	14	11
10/25V	15	13	10	1.0/50V	16	14	11
10/50V	16	14	12	4.7/16V	15	13	10
22/25V	17	15	12	4.7/25V	15	13	10
22/50V	24	20	18	4.7/50V	16	14	11
47/25V	19	17	15	10/16V	14	12	09
47/50V	25	21	19	10/25V	15	13	10
100/25V	24	20	18	10/50V	16	14	12
100/50V	35	30	28	47/50V	24	21	19
220/25V	32	28	25	100/16V	19	15	14
220/50V	45	41	38	100/25V	24	20	18
470/25V	33	29	27	100/50V	35	30	28
1000/16V	55	50	45	220/16V	23	17	16
2200/16V	70	62	55	470/25V	31	28	26

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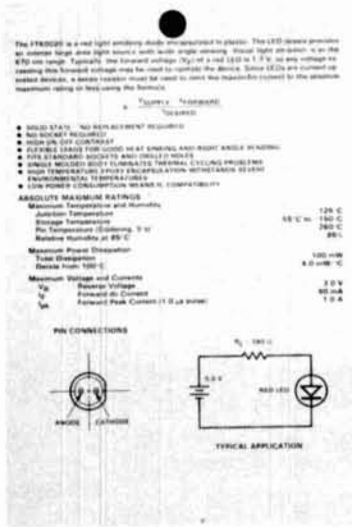

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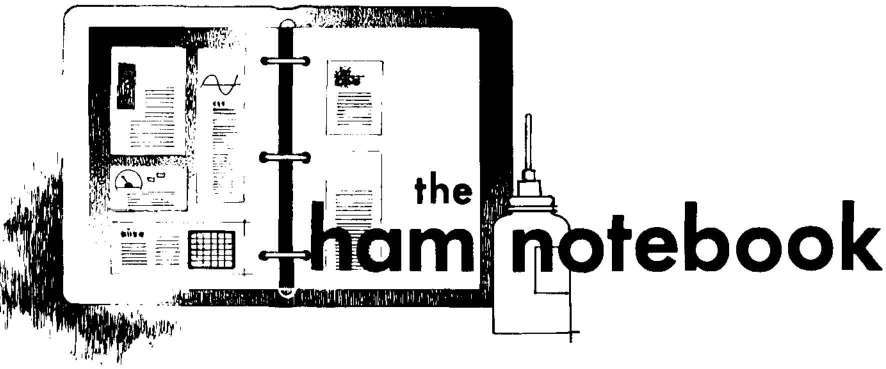


another squelch circuit

Interest has developed in the squelch circuit appearing in the September, 1974, issue of *ham radio*, page 68, which was published before the inexpensive and popular LM380 audio amplifier IC became readily available. **Fig. 1** shows a squelch circuit with excellent performance and greater simplicity.*

Transistor Q1 amplifies random noise which is greater in frequency than the normal spectrum occupied by voice communications during no-signal conditions. Diodes CR1 and CR2 rectify the noise, which is filtered by R6, C5. Q2 conducts and clamps U1 off during no-signal conditions. A carrier activates the LM380. R6 determines the length of the squelch tail, and R5 controls the promptness to which the circuit responds to carrier signals. Increasing the value of C1 increases the noise-amplifier gain, but the smaller the value of C1, the less susceptible the circuit will be in responding to heavy voice peaks, thus the more positive its operation. Cost of the parts at the time of this writing: Q1 $0.45, Q2 $0.77, U1 $1.65. Not bad!

Robert Harris, WB4WSU

*A complete kit, including circuit board, of the audio amplifier section only (less volume control and speaker) is available for $4.25 from Tekpro Design Systems, Box 6324, Virginia Beach, Virginia 23456.

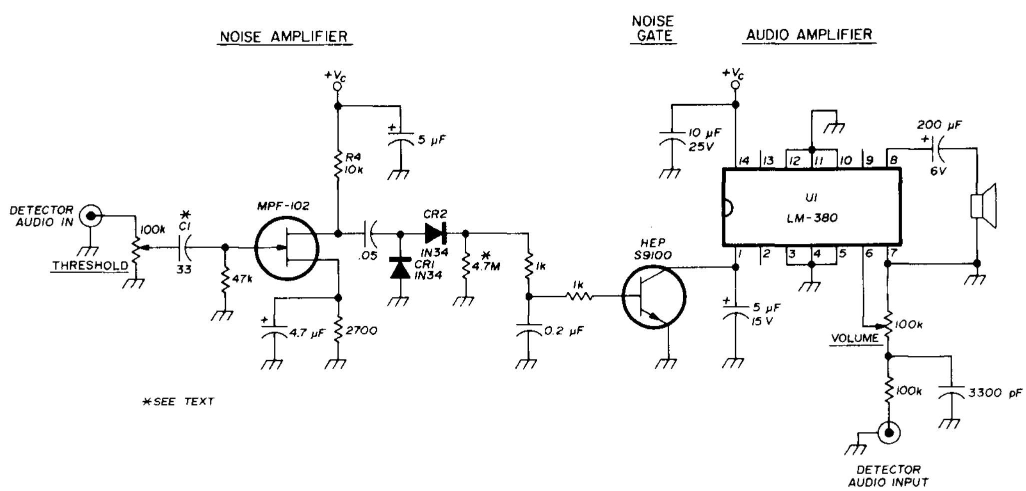


fig. 1. Squelch circuit using the M-380 audio amplifier IC.

transfer letters

In the past I have had great difficulty with dry transfer letters failing to adhere to various surfaces, especially unfinished aluminum. There are two techniques which I have found helpful in overcoming this problem. The first is to abrade the surface to be lettered gently with very fine (0000 or finer) sandpaper and water, and then to dry it thoroughly before lettering. Alternatively, swab the area to be lettered with acetone (most nail polish removers will do). Other solvents probably would work as well, but I have not experimented with them. Either of these methods should always be used before trying to letter a raw aluminum surface that has been handled, or etched in a sodium hydroxide bath (this process leaves a waxy film on the aluminum, and the letters will not adhere to it).

Michael Tortorella, WA2TGL

voltage safety valve

The safety circuit in **fig. 2** was developed after several accidents where a power transistor shorted in a low-voltage power supply causing the transformer output voltage to be supplied to a solid-state transceiver. This voltage could be up to three times the rectified voltage and could cause considerable damage to the transceiver.

A high wattage zener diode could be used across the output of the power supplied, but it takes a very large zener to blow a fuse. Furthermore, the zener could burn open if used to carry currents above its rating.

What was needed was a circuit that would respond when a set limiting voltage was reached and cut off the current flow. The circuit in **fig. 2** is simple and cheap, and uses a relay with contacts large enough to handle the required current (6 or 8 amperes should be ample). The relay coil should be about 200 ohms, the potentiometer 100 ohms or more, the zener diode is a 6 or 8 volt, 10 watt unit.

This circuit can be adjusted to trip at 14 to 24 volts. When the limiting voltage is reached the relay opens up, breaking the circuit to the transceiver. The residual charge in the circuit will hold the relay open until the voltage drops or the power is disconnected.

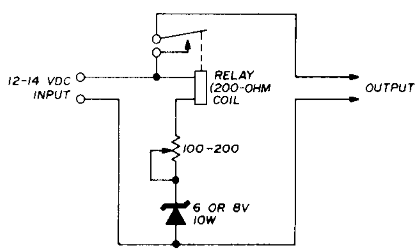


fig. 2. Over-voltage protection circuit. Potentiometer can be set so relay picks up at any preset voltage between 14 and 24 volts.

Harold C. Dressel, W2UVF

Swan 250 carrier suppression

Most Swan 250 transceivers have only about 30 to 35 dB carrier rejection. I have done some work on my unit and now have obtained more than 50 dB carrier rejection. All I did was put 1% resistors in the plate and screen circuits, remove the original 5k carrier balance pot, and install a ten-turn 2000-ohm pot with 1.5k resistors on each side of the pot to give finer control (see circuit in **fig. 3**).

After doing this I can still insert enough carrier to get full power output from the Swan 250. Before the modification I had about 150 mW of output power with the carrier completely nulled; now, with the carrier completely nulled, output is only 1.5 mW. All measurements were made with a Bird Thruline wattmeter.

Charles A. Beener, WB8LGA

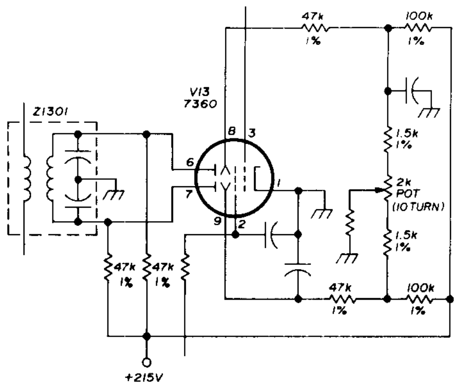


fig. 3. New carrier-balance circuit for the Swan 250 increases carrier suppression to 50 dB or more. The 2000-ohm pot is a miniature, precision 10-turn unit (Bournes 3707).

using your signal generator for absorption measurements

An absorption meter measures the frequency of an rf circuit to which it is coupled. Like a conventional signal generator, it uses an rf tank circuit to tune over the desired band. Therefore, it is logical to combine both functions in a single instrument. I modified my RCA model WR-50A signal generator to also operate as an absorption meter. Performance of the generator is not affected in any way.

The following are added: two RCA-type phono jacks on the front panel, an rf detecting diode, and a pickup (search) coil (see **fig. 4**). J1 is for the coil, J2 for a 50 μA meter. There is only a slight error in the dial calibration for absorption measurements (due to the pickup coil shunting the cathode portion of the generator tank).

The search coil does not have to be plugged in at the panel where it would be awkard to couple to anything, but may be plugged in at the far end of an extension cable. I use 4 feet (1.2m) of ordinary shielded audio cable and find that 8 turns around a ¼ inch (6.5mm) polystyrene coil form works well up to the frequency limit of the generator (40 MHz). An RCA phono plug is mounted on the coil form.

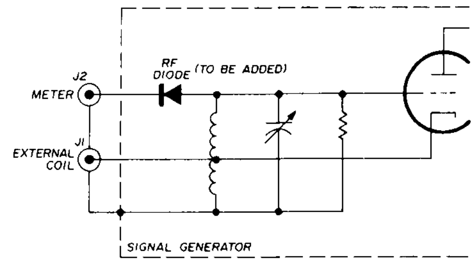


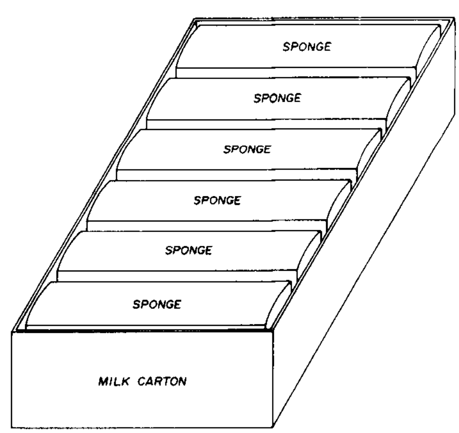
fig. 4. The rf pickup coil is rectified by the diode and indicated by the meter.

When you are ready to test the absorption device, couple its search coil to a grid-dip meter. At resonance, the meter will read about full-scale with the coils ½ inch (13mm) apart. Turn off the signal generator when making absorption measurements.

I. Queen, W2OUX

soldering tip cleaner

A clean soldering-iron tip helps make reliable solder connections. **Fig. 5** shows an inexpensive tip cleaner made from a half-pint (¼ liter) milk container and some sponges cut to fit the container.



SOLDERING IRON TIP CLEANER

fig. 5. Soldering iron tip cleaner.

Best results are obtained when the sponges are kept moist. To clean your soldering iron, just insert the tip between the sponges.

Gary Tater, W3HUC

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The Atlas 210x or 215x measures only 9½" wide x 9½" deep x only 3½" high, yet the above photograph shows how easily the Atlas transceiver fits into a compact car. And there's plenty of room to spare for VHF gear and other accessory equipment. With the exclusive Atlas plug-in design, you can slip your Atlas in and out of your car in a matter of seconds. All connections are made automatically.

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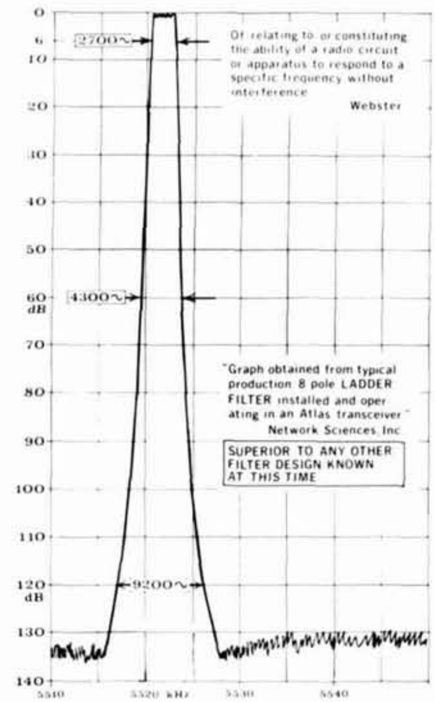

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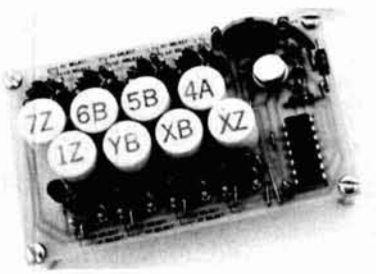
NEW products

sub-audible tone encoder



Communications Specialists announce their new ME-8, a sub-audible tone encoder for repeater access. Measuring only 2x3x9 inches (5x7.5x22.5cm) and operating on any voltage between 6 and 16 volts dc, the unit is capable of generating any eight of the 32 EIA tone codes between 67.0 and 203.5 Hz.

Frequency selection is made by supplying ground or positive voltage to the desired control lead, and plugging in the desired K-1 element. The unit is completely immune to rf and has a start-up time of only 10 milliseconds. Frequency accuracy and stability are excellent and the output is a low-distortion sinewave at 3 volts rms.

The ME-8 is priced at $79.95 and includes eight K-1 elements. Extra elements are available for $3.00 each. For additional information write Communications Specialists, Post Office Box 153, La Brea, California 92621 or use *check-off* on page 134.

high-current regulated power supplies

VHF Engineering, Binghamton, New York, has announced two new regulated 12-volt dc power supplies for use in amateur and commercial applications. The two power supplies have current ratings of 15 amps at 10 to 12 volts dc, and 25 amps at 10 to 12 volts dc, respectively. Voltage regulation is 2% from no load to maximum rated current.

These new power supplies feature over-voltage protection and fold-back current limiting. Over-voltage protection is provided by a crowbar circuit which will shut down the supply if its output voltage exceeds 14 volts, thereby protecting the equipment being powered. The foldback current-limiting circuit limits the output current to a maximum of 1 amp if the output is shorted. This protects the power supply from damage and prevents high current from flowing into a piece of equipment that may have developed an internal short. When the short is removed, the power supply output returns to normal.

These regulated power supplies have been designed to commercial standards and use tinned, glass-filled epoxy circuit boards and high quality components. Additional information may be obtained from VHF Engineering, 320 Water Street, P.O. Box 1921, Binghamton, New York 13902 or use *check-off* on page 134.

air-variable capacitors

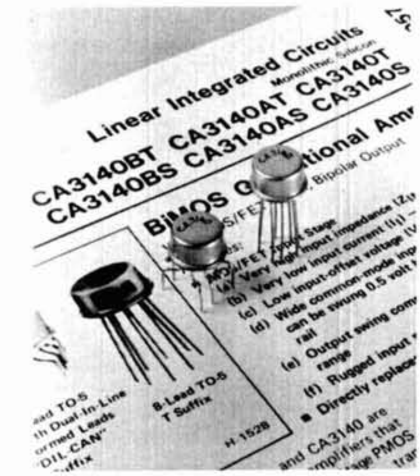
Among the components which are most difficult to find these days are air-variable capacitors, especially those rated for transmitting powers. Now, however, the Dentron Radio Company is offering a line of air variables which are suitable for most amateur applications in transmitters and antenna tuning units. Listed in the table above are the electrical specifications for these capacitors.

model	type	capacitance minimum	capacitance maximum	voltage rating	price
D-88-120	single	53 pF	208 pF	6000	$23.50
D-140-75	single	23 pF	140 pF	4500	16.50
D-232-45	single	23 pF	232 pF	3000	16.50
D-500-45	single	48 pF	500 pF	3000	19.50
DD-150	dual	33 pF	205 pF	3000	20.00

(per section)

These air-variable capacitors may be ordered directly from Dentron Radio Company, 2100 Enterprise Parkway, Twinsburg, Ohio 44087. For more information use *check-off* on page 134.

bipolar/mos op amp



A new bipolar/mos op amp, the CA3140, featuring a pmos input stage and a bipolar output stage with a wide output voltage range, has been introduced by RCA Solid State Division. Its versatility permits it to fill virtually all 741 sockets and, at the same time, most of the premium op amp sockets currently served. The pmos input stage is similar to the one used in the RCA CA3130 op amp but with internal compensation and the ability to operate from a supply voltage between 4 and 44 volts, dual or single supply. A special feature is the addition of bipolar diodes which protect the input to such an extent that preliminary tests, under simulated electrostatic conditions up to 1000 volts, show the CA3140 to be more rugged than any other device yet tested, including bipolar and fet input op amps.

Typical performance features for the CA3140 include: an input impedance of 1500 megohms; input current of 10 pA at ± 15 volts, low input-offset voltage of

5 mV, wide common-mode input voltage range, −0.5V below negative bus; output swing to within 0.2 volt of negative supply; high and low operating supply voltage of 4 to 44 volts, high slew rate of 9V/μs; high gain bandwidth product of 4.5 MHz; and fast settling time of 1.4 μs, to 10 mV with a 10 $volt_{p-p}$ signal. The new device is available in the basic TO-5 package or the dual-in-line package.

The breadth of applications permits the device to replace numerous op amp categories such as general purpose, fet-input, and wideband or high slew-rate types. The wide bandwidth feature reduces costs by permitting wideband video and audio circuits at lower cost than with current op amps, as well as lower-cost wideband TTL interfaces. The strobable output stage allows the output to be driven low, independent of the input signal. The fact that the output swings to within 0.2 volt of the negative supply permits power transistors to be driven directly, thus eliminating level shifting circuitry.

For additional information, including copies of the 20-page data sheet, File no. 957, and a descriptive applications brochure, Publication no. 2M1144, offering a free sample, write to RCA Solid State Division, Box 3200, Somerville, New Jersey 08876 or use *check-off* on page 134.

digital pulse generator

Continental Specialties Corporation has introduced the Design Mate-4, a versatile, laboratory-quality generator which lists for $124.95. The DM-4 may *be used as a clock source, delayed pulse* generator, synchronous clock source, manual system stepper, pulse stretcher, or clock burst generator.

The DM-4 is designed to fill the bill wherever a source of clean, crisp, fast output pulses compatible with virtually all logic families and descrete circuits is required. It is capable of generating symmetrical and asymmetrical pulses from 0.5 Hz - 5 MHz and has a positive output of 100 mV to 10 V, with a rise and fall time of less than 30 nanoseconds. Additionally, the DM-4 offers an independently controlled pulse width and spacing from 100 ns to 1 second in seven overlapping ranges, as well as independent variable-amplitude CMOS, and fixed-amplitude TTL outputs. The unit operates either continuously or in manual one-shot fashion. It also features external triggering from dc to 10 MHz and synchronous output gating.

For more information contact Continental Specialties Corporation, 44 Kendall Street, Box 1942, New Haven, Connecticut 06509 or use *check-off* on page 134.

electronic keyer

The *CW Sendin' Machine* was designed by an amateur and allows you to write in your log while it sends your routine information. Not only a fine keyer with iambic operation and dot and dash memories, it is also equipped with two random-access memories that can store short CW messages and send them at the push of a button.

The model 2048 is equipped with a cord and plug to be connected to your favorite paddle, The keyer contains no relays and can operate into open-key voltages of up to 150 volts and closed-key currents to 10 mA. Sidetone level can be controlled by a volume control; pitch can be varied by a pot on the PC board.

The *CW Sendin' Machine* model 2048 comes with two 1024-bit memories, and an automatic reset feature is available as a modification which can be added to your board. The PC board is on a connector for easy service. The board operates on 5 volts at 400 mA.

For more information write H.A. Harp, WA4SVH, 718 Magnolia Dr., Lake Park, Florida 33403 or use *check-off* on page 134.

More Details? CHECK-OFF Page 134

short circuits

frequency synthesizers

Several errors crept into DJ2LR's excellent article on frequency synthesizers in the July, 1976, issue of *ham radio*. In **fig. 13** (page 16), the wiper of the 4.7k potentiometer between the emitters of the two 2N3570 transistors should be grounded. In **fig. 17** the input labeled "10 kHz from divider" should show a pulse width of 200-500 nanoseconds (*not* microseconds). In **fig. 20** the CD4000 gate should be a 74C00, and in **fig. 23** the gate with three inputs which is not identified is a 74LS30. Finally, the caption to **fig. 24** on page 23 is incorrect — this circuit *adds* 41 MHz to the synthesizer reading.

DT-600 RTTY demodulator

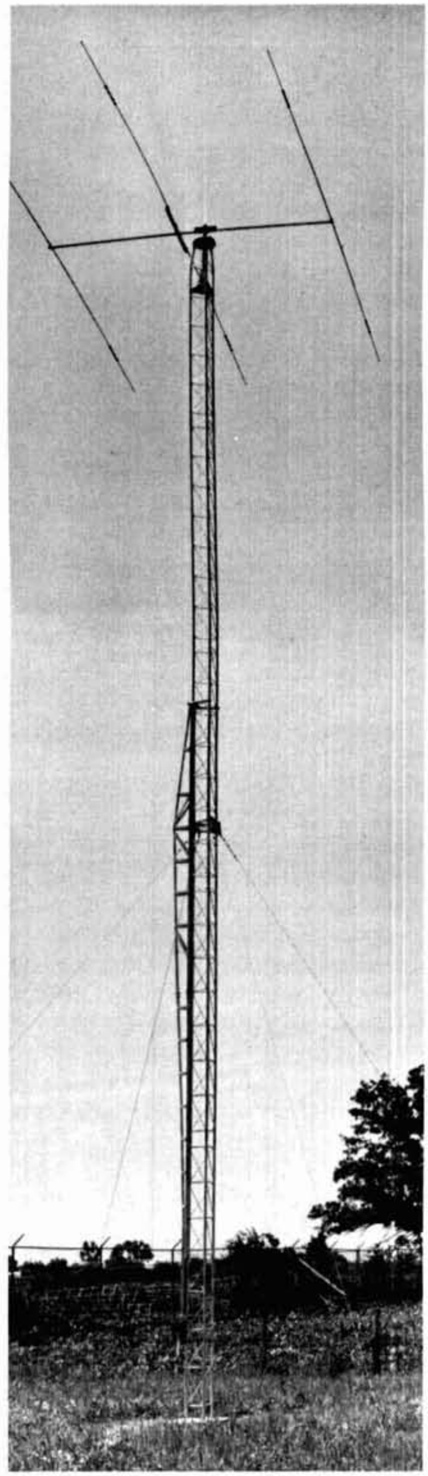
In the schematic of the DT-600 RTTY demodulator in the February, 1976, issue of *ham radio* one end of the 10k balance pot (at the input to U1, wiper connected to R9) should go to −12 Vdc. This pot should be labeled R8. In **fig. 5** the horizontal axis was measured in Hz but mistakenly labeled in baud; none of the standard RTTY speeds (60, 75, and 100 wpm) are attenuated. Potentiometers to fit the circuit board are available from Data Technology Associates, Inc., Box 431912, Miami, Florida 33143. A set of four pots is priced at $2.00.

DT-500 RTTY demodulator

Due to an editorial oversight, the names of co-authors Garey K. Barrell, K4OAH, and Archie C. Lamb, WB4KUR, were inadvertently deleted from the cover page of the DT-500 article in the March, 1976, issue. In addition, in **fig. 2** (page 26), the negative return from the +170 volt loop supply should be connected to ground through a 2500 ohm, 20 watt resistor. Printed-circuit boards for the DT-500 are priced at $10.50 from Data Technology Associates, Inc., Box 431912, Miami, Florida 33143. A set of four PC potentiometers is available from the same source for $2.00.

S-line frequency synthesizer

In the S-line frequency synthesizer article published in the December, 1975, issue, in **fig. 3** (page 12) R2 should be connected to pins 4 and 5 of U1B, *not* pin 6. The parts layout in **fig. 10** is correct.



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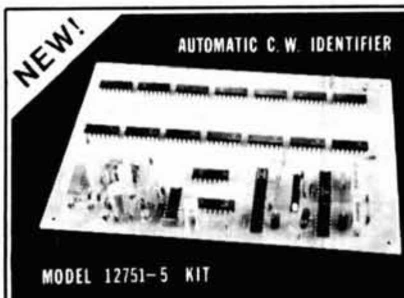

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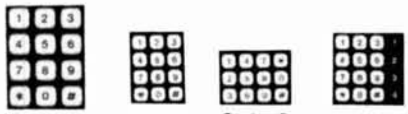

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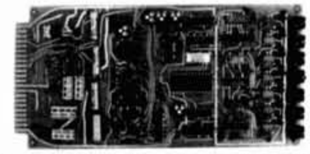

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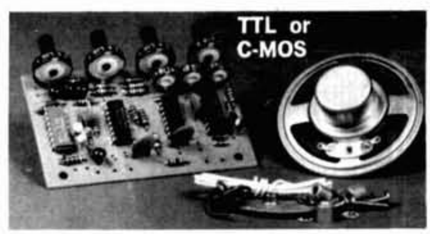

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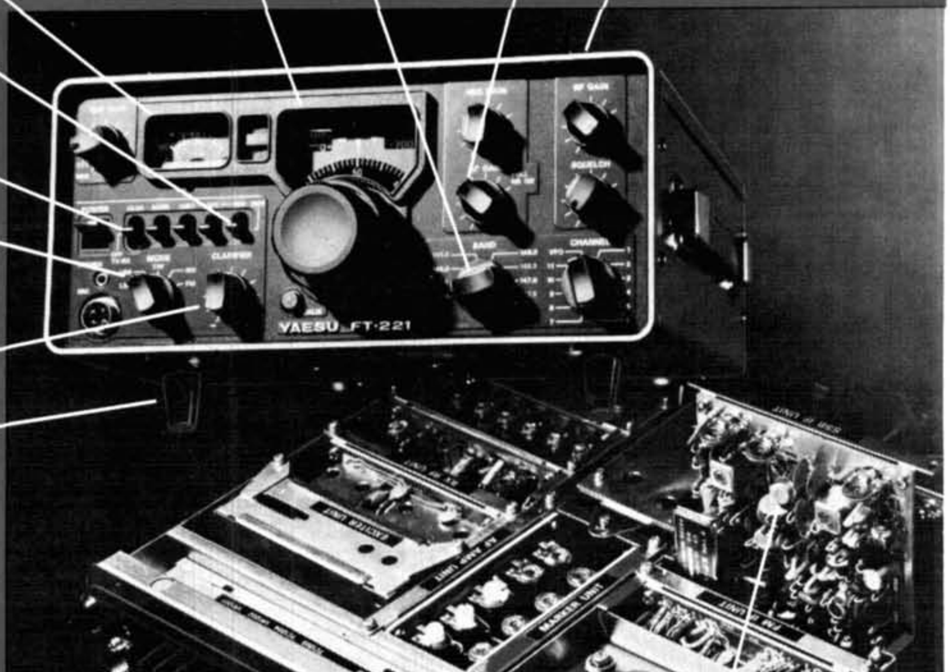

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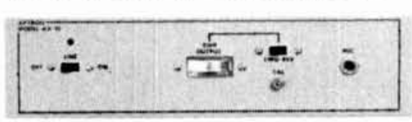

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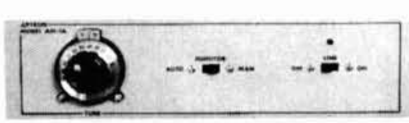

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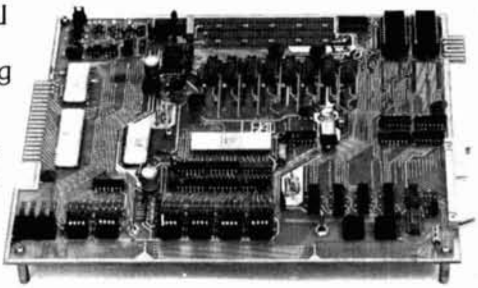

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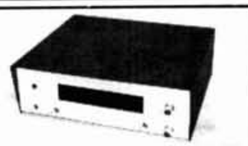

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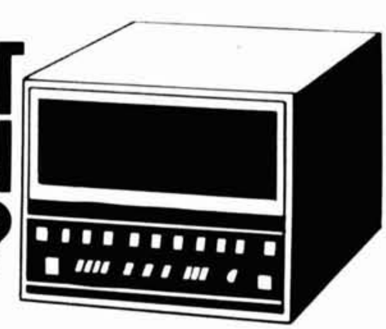

The last few years have seen a remarkable quantum jump in the cost effectiveness of computing power brought about by the advent of inexpensive microcomputer systems. You've been reading about these machines for some time in the technical press. You've seen stories of applications ranging from specialized numerical control systems for machine tools to intelligent traffic light controllers.

One of the most exciting possibilities which has come about as a result of the inexpensive LSI computer is the fact that a general purpose computing system is within the budget range of you, the individual with an interest in the technology and an appreciation of both the intellectual challenge and fun of building and using a personal computing system. There is now a whole micro industry of companies which package and market microcomputer related items as kits, finished products and related items for personal use. The products range from complete system kits with hardware and software to peripherals, software and literature on the subject. But where do you go to find out about this personal involvement aspect of microcomputer technology?

Turn to BYTE, The Small Systems Journal for a wealth of information on the personal use of computer systems. BYTE is a monthly compendium of information you need for personal use applications of microcomputers:

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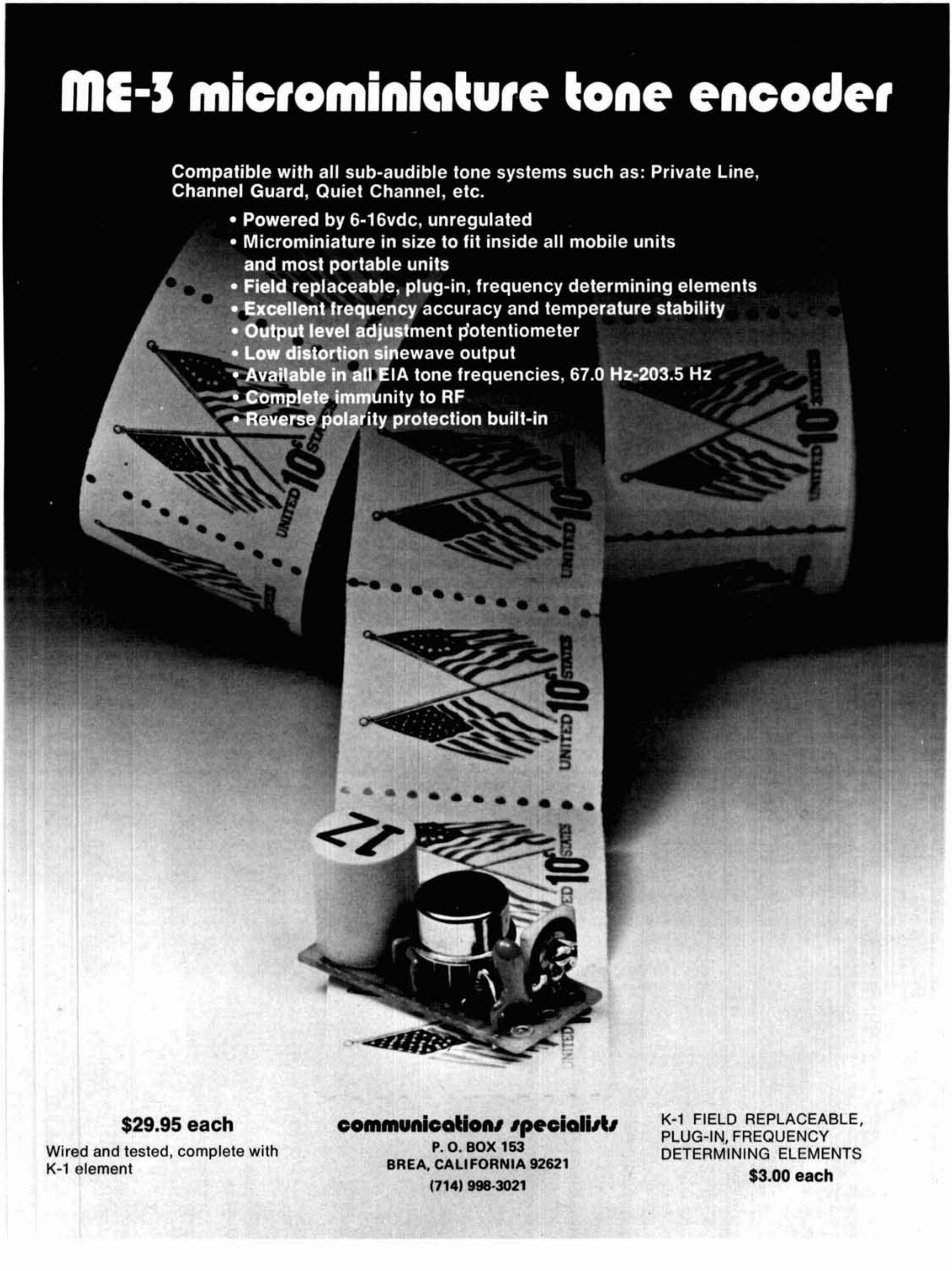
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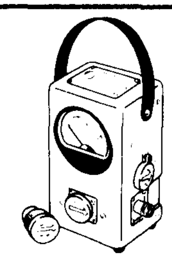

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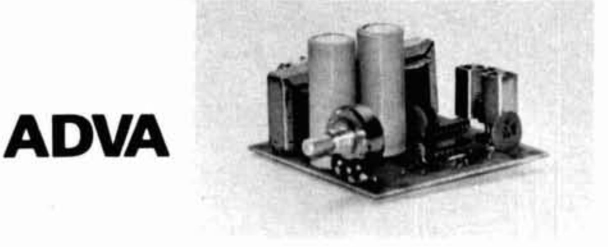

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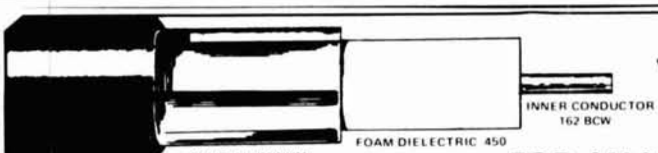

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HP616 Sig gen 1.8-4gHz FM-CW ... 385
HP686 Sweep Gen 8.2-12.4gHz Sweep range 4.4mHz-4.4gHz ... 595
HP803A VHF Ant bridge 50-500mHz ... 95
SINGER SSB4 Sideband spec anal 0-40mHz, res. to 10Hz ... 685
TEK 181 Time-mark scope calib. ... 45
TEK 190 Sig gen(const ampl) 50mHz ... 125
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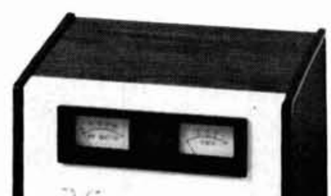

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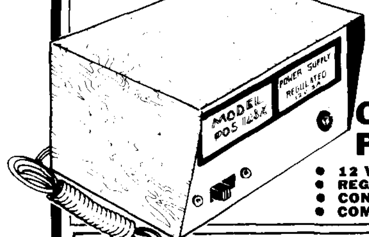

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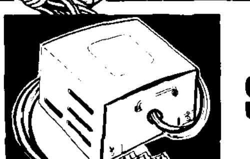

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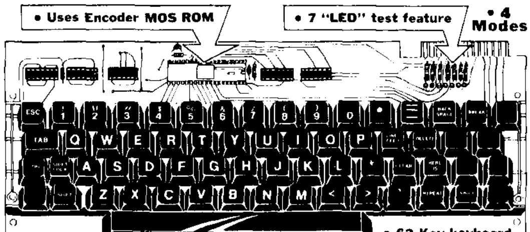

- 63-Key keyboard

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Outputs — standard ASCII 7 bits plus strobe

$59.95 ☐ KIT

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Electrical specs: voltage requirements plus 5 volts and minus 12 volts, 200mw, negative or positive logic output, jumper selectable. Output connector. *On the PC board we have* A NEW KEYBOARD TEST FEATURE . . . . 7 LEDS! These LEDs display visually the ASCII code for the depressed key. TTL compatible. Kit includes: 63 key keyboard, pc board, ROM, 7 LEDs, resistors, capacitors, TTL IC's and SOCKETS We include diagram and wiring instructons, ROM code. Wt. 7 lbs. Size overall: 13 x 5½ x 2".

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$19.95 ☐

Cat. No.	Description	Sale	BUY 'EM SEPARATELY		
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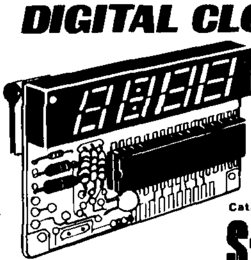

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Cat. No. 10H3411

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☐ SN7401	.22	.23	☐ SN7495	.86	.87
☐ SN7402	.22	.23	☐ SN7496	.86	.87
☐ SN7403	.22	.23	☐ SN74125	1.29	1.30
☐ SN7404	.24	.25	☐ SN74126	1.29	1.30
☐ SN7406	.36	.37	☐ SN74132	1.95	1.96
☐ SN7407	.46	.47	☐ SN74141	1.12	1.13
☐ SN7408	.22	.23	☐ SN74145	1.69	1.70
☐ SN7409	.52	.53	☐ SN74148	3.25	3.26
☐ SN7410	.22	.23	☐ SN74150	1.12	1.13
☐ SN7411	.27	.28	☐ SN74153	1.10	1.11
☐ SN7417	.62	.63	☐ SN74154	1.53	1.54
☐ SN7420	.22	.23	☐ SN74157	1.09	1.10
☐ SN7421	.55	.56	☐ SN74158	1.55	1.56
☐ SN7423	.47	.48	☐ SN74160	1.59	1.60
☐ SN7425	.37	.38	☐ SN74163	1.50	1.51
☐ SN7430	.22	.23	☐ SN74164	1.79	1.80
☐ SN7432	.36	.37	☐ SN74165	1.79	1.80
☐ SN7440	.22	.23	☐ SN74173	1.69	1.70
☐ SN7442	.97	.98	☐ SN74174	1.42	1.43
☐ SN7445	.97	.98	☐ SN74175	1.45	1.46
☐ SN7446	1.10	1.11	☐ SN74176	1.59	1.60
☐ SN7447	1.10	1.11	☐ SN74177	1.59	1.60
☐ SN7448	1.10	1.11	☐ SN74181	3.75	3.76
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☐ SN7451	.27	.28	☐ SN74190	2.75	2.76
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☐ SN7485	1.41	1.42	☐ SN74196	2.19	2.20
☐ SN7486	.47	.48	☐ SN74198	2.25	2.26
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FREQUENCY COUNTER BOARDS, Jan. 76 Ham Radio includes 500 MHz prescaler circuitry and LED board with instructions, $15.00. QST March 76 speech compressor, $10.00. Both projects come with parts source listing. Double sided glass epoxy, drilled aand plated. CSJ Electronics, 5201 Cameron Ct., Lincoln, NE. 68512.

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NOT C.B. decals, 1 x 2 inches, black over white, plastic, weatherproof, peal and stick. Great for load coils, radios, control heads, etc. Stop having your radio gear mistaken for C.B. radio. 10 for $1.00, with SASE 12 for $1.00, 100 for $7.00. WA4SGI, Ronald P. Pitts, 2113 Lester Lane, Birmingham, AL 35226.

SWAN, 1200X, linear amplifier, new, in sealed carton, $250. Russell, 19680 Mountville Dr., Maple Hts., Oh. 44137.

QRP TRANSMATCH for HW7, Ten-Tec, and others. Send stamp for details to Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

FOR SALE: AR-88, RCA WW-II comm. receiver. Top condition, with manual and complete spares. Tunes 540 kHz to 30 MHz in 5 bands. Art Fillebrown, 48 Hunton St., Warrenton, Va. 22186.

MAXI TUNER solves antenna problems! Matches coax, random wires, and balanced feedlines from 160 thru 10 meters. Features Johnson 229-203 rotary inductor, vernier dials, optional SWR metering, styled cabinet and color options to complement Collins, Heath, and Drake lines. Spec. sheet, pricing, and amateur market antenna tuner comparison chart from RF Power Components, P. O. Box 11, Ladysmith, WI 54848. (715) 532-3971.

MOTOROLA HT220, 4 freq., 2 mtr. $275. Motrac U43HHT-3100 vhf $210, Mocom D34DRT-3100 UHF $190, Mocom D33CMT-3100 vhf $190. Daniel M. Herlihy, K6KTP, 2338 Berry St., Lemon Grove, Cal. 92045. 714-466-7558.

SALE: Model 28 ASR's - KSR's repurfs - keyboards TD's - printers parts - all priced for hams. All in excellent condition. A.D.M. Communications, Inc., 1322 Industrial Avenue, Escondido, Ca. 92025. (714) 747-0374.

MANUALS NEEDED. RCA ET-4336-H, Waterman S-14 Pocketscope and HP 524B Counter. State price first letter. W1KCR, Claremont, New Hampshire.

TRAVEL-PAK QSL KIT — Send call and 25¢; receive your call sample kit in return. Samco, Box 203, Wynantskill, N. Y. 12198.

ARTS ON 7103 KHZ DAILY 1300Z to 1900Z for emergency or routine traffic. Amateur Radio Telegraph Society, 2730 South Tabor Avenue, Silver City, New Mexico 88061.

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TELEVISION SYNC GENERATOR. Broadcast standards. Best offer or swap for VHF or UHF transceiver, CB SSB, etc. Manual, W9YLU. 6731 N. Hermitage, Chicago, IL 60626.

MANUALS for most ham gear, 1939/70. List $1.00. Send SASE (or 25¢) for one specific model quote. Hobby Industry, WØJJK, Box H864, Council Bluffs, Iowa 51501.

THE PERFECT GIFT FOR A HAM — His/her callsign cut into solid redwood. Type one: just the call, approximately 18" x 5½", $7.50. Type two: the call plus any handle, seven letters or less, approximately 18" x 9½", $12.50. Callsigns Ltd., P. O. Box 101, Hurley, N. Y. 12443. N. Y. State residents add sales tax.

RADIO MUSEUM now open. Free Admission. 15,000 pieces of equipment from 1850 telegraph instruments to amateur and commercial transmitters of the 1920's. Amateur station W2AN. Write for information. Antique Wireless Assn., Main St., Holcomb, N. Y. 14469.

OSCILLOSCOPE AN/USM-140D, dual trace, 24 MHz, $325. HP-140A/1415A time domain reflectometer, $600. HP-130BR oscilloscope 300 kHz, $200. B & K 445 CRT tester, $50. AN/URM-32A frequency meter, $100. Boonton 207H univerter, $150. HP-302A wave analyzer, $550. Hickok 1600B VTVM, $75. AN/USM-34 VTVM, $35. Both with RF probes. All instruments in working condition with manuals. James Walter, 2697 Nickel, San Pablo, Ca. 94806.

SIDESWIPER only $13. Airmailed USA. Kungsimport, Box 257, Kungsbacka, Sweden.

MOTOROLA HT220, HT200, Pageboy, and other popular 2M FM transceiver (Standard, Regency, etc.) service and modifications performed at reasonable rates. WA4FRV (804) 320-4439 evenings.

More Details? CHECK-OFF Page 134

6 Digit LED Clock Kit - 12/24 hr.

$9.95 ea. QTY. 12 OR MORE — $10.95 ea. QTY. 6-11 — $11.95 ea. QTY. 1-5

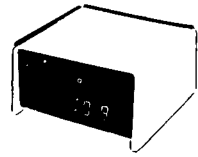
KIT INCLUDES
- INSTRUCTIONS
- QUALITY COMPONENTS
- 50 or 60 Hz OPERATION
- 12 or 24 HR OPERATION

6-LED Readouts(FND-359 Red, com. cathode)
1-MM5314 Clock Chip (24 pin)
13-Transistors
3-Switches
6-Capacitors
5-Diodes
9-Resistors
24-Molex pins for IC socket

**LARGE .4" DIGITS!
ORDER KIT #850-4
AN INCREDIBLE VALUE!**

"Kit #850-4 will furnish a complete set of clock components as listed. The only additional items required are a 7-12 VAC transformer, a circuit board and a cabinet, if desired."

Printed Circuit Board for kit # 850-4 (etched & drilled fiberglass) $2.95
Mini-Brite Red LED's (for colon in clock display) pkg. of 5 .... 1.00
Molded Plug Transformer 115/10 VAC (with cord) 2.50
NOTE: Entire Clock may be assembled on one PC Board or Board may be cut to remote display.
Kit # 850-4 will fit Plexiglas Cabinet II.



PLEXIGLAS CABINETS

Great for Clocks or any LED Digital project. Clear-Red Chassis serves as Bezel to increase contrast of digital displays.

CABINET I
3"H,6¼"W,5½"D
CABINET II
2½"H,5"W,4"D

Black, White or Clear Cover

ANY SIZE/COLOR **$6.50** ea. **2/$12.**

RED OR GREY PLEXIGLAS FOR DIGITAL BEZELS
3"x6"x1/8" **95¢** ea. **4/$3**

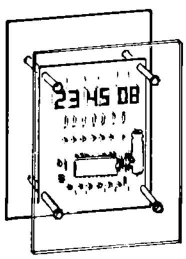
60 HZ.

XTAL TIME BASE
Will enable Digital Clock Kits or Clock-Calendar Kits to operate from 12V DC.
1"x2" PC Board
Power Req: 5-15VDC (2.5 MA. TYP.)
Easy 3 wire hookup
Accuracy: ± 2PPM (Adjustable)
Complete Kit **$4.95** ea.
#TB-1

**SEE THE WORKS Clock Kit
Clear Plexiglas Stand**

- 6Big .4" digits
- 12 or 24 hr. time
- 3 set switches (back)
- Plug transformer
- all parts included

Plexiglas is Pre-cut & drilled
Size: 6"H,4⅓"W,3"D



A SUPER LOOKING CLOCK!

Kit #850-4CP **$23.50** ea. **2/$45.**

6 Digit-LED Clock-Calendar-Alarm Kit

• 12/24 HR TIME • CHOICE OF DIGITS • 28-30-31 DAY CALENDAR • AC FAILURE/BATTERY BACK-UP • 24 HR ALARM - 10 MIN. SNOOZE • ALTERNATES TIME [8 SEC] AND DATE [2 SEC] OR DISPLAYS TIME ONLY AND DATE ON DEMAND • THIS KIT USES THE FANTASTIC CT-7001 CHIP. FOR THE PERSON THAT WANTS A SUPER CLOCK KIT-TOO MANY FEATURES TO LIST!

#7001 B [6 - .4" Digits/Man-64]	$39.95
#7001C [4 - .6" Digits/2-.3" Seconds]	$42.95
#7001 D [4-.8" Digits/2-.3" Seconds]	$45.95

ALL KITS ARE COMPLETE INCLUDING IC SOCKET, TRANSFORMER, LINE CORD, SWITCHES, etc.
Cabinet not included (Ideal fit in Cabinet I above)

½" DIGIT CLOCK KIT

A COMPLETE KIT LESS CABINET. FEATURES: MM5314 IC, 12/24 HR, 50/60 H2, 6-FND-503 LED'S, PLUG-TRANSFORMER, LINE CORD, etc.
[Ideal Fit in Cabinet II]
Kit #5314-5. .... **$19.95** ea. **2/$38.**

JUMBO DIGIT CONVERSION KIT

Convert small digit LED clock to large .5" displays.Kit includes 6-.5" LED's,Multiplex PC Board & easy hook-up info.

**Kit #JD-1CC For common Cathode
Kit #JD-1CA For common Anode** **$9.95** ea. **2/$19.**

Fairchild Super Digit FND-359

.4" Char. Ht. 7 segment LED RED Com. Cath. Direct pin replacement for popular FND-70.

**95¢ ea, 10/$8.50
100/$79.00**

SET OF 6 FND-359 WITH MULTIPLEX PC BOARD $6.95

CLOCK IC'S

CT-7001 ...... 7.95	CT-7002 ..... 13.95
MM5314N.... 3.95	MM5369N.... 2.50
MM5316N... 4.95	MM5375AB.. 3.95

IC SOCKETS — Solder Tail

Pin	Profile	1-24	25	100
14	LP	$.25	$.22	$.20
16	LP	.28	.25	.23
18	LP	.31	.28	.26
24	SP	.50	.45	.40
28	SP	.60	.55	.50
40	LP	.75	.70	.65

NYLON WIRE TIES
8" for bundle dia. 1/4"-1-3/4" 100/$1.95
4" for bundle dia. 1/16"-3/4" 100/$1.75

25 AMP BRIDGE

**$1.95 ea.
3/$5.00**
100 PIV

Form Inexpensive Sockets **MOLEX PINS**
**100 for $1.25
Reel of 1000 - $8.50**

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IN4002	12/$1.00
IN4003	12/$1.00
IN4005	11/$1.00
IN4007	10/$1.00
2.5A/1000PIV	5/$1.00
IN5400	5/$1.00
IN914	20/$1.00
IN4148	20/$1.00

SWITCHES

ROCKER SPDT	6/$1.
MINI-SLIDE SPDT	5/$1.
REG. SLIDE DPDT	6/$1.
PUSH BUTTON N.O.	3/$1.

TRANSISTORS 5/$1.00
All Prime Marked Units

2N2222A	NPN	TO-18
2N3415	NPN	TO-92
2N3704	NPN	TO-92
2N4249	PNP	TO-92
2N4400	NPN	TO-92
2N4437	NPN	TO-92
2N5089	NPN	TO-92
2N6027	PUT	TO-92

**TRANSISTOR SOCKET GOLD PINS
5/$1.00**

TELEPHONE FORMAT KEYBOARD
BY Chomerics

2-1/9"x3"
5/32" thick
**$4.95
6/$28.**

7-SEG LED

COMMON CATHODE

HP5082-7702	RED	.3"	1.25
FND-71	±1	.25"	.75
FND-359	RED	.4"	.95
FND-503	RED	.5"	1.35
FND-803	RED	.8"	3.50
DL-33MMB	RED	3x.1"	.75
DL-750	RED	.6"	2.95
XAN-654	GREEN	.6"	2.95 (no D.P.)
XAN-664	RED	.6"	2.95 (no D.P.)

JUMBO RED LED's Pkg. of 50 $3.95

10K 10 TURN SPECTROL POT SPECTROL 95c
3/8"x3/8"x1/4" High #50-4-11-103 4/$3.00

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flea market

FIGHT TVI with the RSO Low Pass Filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada. MIS 3B4.

RTTY — NS-1A PLL TU (HR 8/76). Wired/tested $29.95 ppd. Board $4.75 ppd. SASE for info. Nat Stinnette Electronics, Tavares, FL 32778.

TELETYPEWRITER PARTS, gears, manuals, supplies, tape, toroids. SASE list. Typetronics, Box 8873, Ft. Lauderdale, Fl. 33310. Buy parts, late machines.

TECH MANUALS for Govt. surplus gear — $6.50 each: SP-600JX, URM-25D, OS-8A/U, PRC-8, 9, 10. Thousands more available. Send 50¢ (coin) for 22-page list. W3IHD, 7218 Roanne Drive, Washington, D. C. 20021.

QSL'S — BROWNIE W3CJI — 3035B Lehigh. Allentown, Pa. 18103. Samples with cut catalog 50¢.

GOOD CONDITION — Conrac video monitors, model CUA17 with rack mount frame and non glare face plate. Ideal for use with slow to fast scan converters, $100 each. Robot model 70B SSTV monitor in original packing, $300. W5DFC, (713) 862-7875.

EXCLUSIVELY HAM TELETYPE 21st year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 35¢. $3.50 per year. Box 837, Royal Oak, Michigan 48068.

BUY — SELL — TRADE. Write for free mailer. Give name, address and call letters. Complete stock of major brands, new and reconditioned equipment. Call for best deals. We buy Collins, Drake, Swan, etc. SSB & FM. Associated Radio, 8012 Conser, Overland Park, Ks. 66204. 913-381-5901.

OSCAR 7, SSB-CW TRANSMIT CONVERTERS. For 28 or 50 MHz input at 20 mw. 432 MHz output at 1 watt. Solid state, for 12 volt supply. 35 watt solid state amplifier available for this converter. Units designed and built by WØENC. Write for information. UHF-VHF Communications, 53 St. Andrew, Rapid City, S. D. 57701.

CIRCUIT BOARDS. Artwork, negatives, etching. SASE for details. Karl Raup, WB4OXG, Box 8013, Orlando, Fla. 32806.

FREE! Over 100 electronics items. "Electronics Sourcebook" tells how, where to obtain free samples, manuals, publications. Write Technical Publications, 1405 Richland, Metairie, La. 70001.

VARIABLE AND TRIMMER CAPACITORS — Millen, Johnson, Hammarlund, Erie, Arco. In stock for immediate shipment, write for free price list "A". D & V Radio Parts, 12805 W. Sarle Rd., R#2, Freeland, Mich. 48623.

YOU CAN INSTALL YOUR OWN SECURITY SYSTEM at a price you never dreamed possible. Our 1976 Burglar-Fire Alarm Catalog describes our complete line of commercial security equipment. It has everything you need to produce a durable, custom installation — just as good as the professionals. Free Catalog. Emel Electronics, P. O. Box 146-D, Sheffield, Mass. 01257.

HEATH HW-101 with CW filter, HP-23B and SB-600. With new tubes and aligned. Used less than one hour on tubes $380. Tennelec MS-2 scanner, brand new $240. Clegg 66er rarely used $100. Kenwood TS-700A new $635. Also Yaesu FR-101 digital receiver and FL-101 transmitter. Write or call. All equipment meets factory specifications and is in mint condition in performance and looks. WBØGOA, Mike Folta, Laddonia, MO 63352. 314-373-2325.

SB-34, mint condition, with mic and xtal calibrator, $240. John Abdnour, WB9TQB, 806 Arcadia, Bloomington, IL 61701.

YOUR AD belongs here too. Why not send it in today.

Coming Events

MEMPHIS IS BEAUTIFUL IN OCTOBER! The Memphis Hamfest, bigger and better than the 3,500 who attended last year, will be held at State Technical Institute, Interstate 40 at Macon Road, on Saturday and Sunday, October 2 and 3. Demonstrations, displays, MARS meetings, flea market, ladies' flea market, too! Hospitality room, informal dinners, XYL entertainment, many outstanding prizes. Dealers and distributors welcome, too! Contact Harry Simpson, W4SCF, Box 27015, Memphis, TN 38127, phone 901 358-5707.

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TOUCHTONE ENCODER KIT
SINGLE IC GENERATES THE STANDARD 16 TONE PAIRS FOR TOUCHTONE DIALING. KIT FEATURES:
Small Size (1.5 x 2.25")
NO Tuning Required
PTT Circuit has built-in 1 second delay
Low Level output is adjustable
Audio output circuit for tone monitoring or acoustic coupling
Kit includes all parts
Plug compatiblewith DIGITRAN® KEYBOARDS
Complete with PC Board
TE-01 $12.00
Same as above but less Audio output and PTT circuit. Smaller size permits use with handitalkies. $10.00 TE–02

PS-01A POWER SUPPLY KIT
A COMPACT, WELL REGULATED TRIPLE OUTPUT POWER SUPPLY. GIVES +5VDC @ 1.5A and +15 @ 150MA AND –15@150MA. THE PS–01A USES THE SUPERIOR 78L SERIES OF 3 TERMINAL REGULATORS.

SPECS:

Voltage:	+5@1.5A
	±15@150MA
Ripple:	Less than 10mv
Regulation:	Load 50mv
	Line 100mv

$14.95
PS–01B AS ABOVE BUT WITH ±12 OUTPUT INSTEAD OF ±15.

SAJ110
7 STAGE FREQ. DIVIDER FOR MUSIC
.99

9 VOLT BATTERY CLIPS
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50 Hz. TIMEBASE KIT
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PNP POWER
TAB TRANSISTOR SIMILAR TO TIP30 IN SPECS. HOUSE NO. GUARANTEED
3/1.00

HARDWARE
A VERY USEABLE ASSORTMENT OF 2–56, 4–40, 6–32 and 8–32 SCREWS AND NUTS
A FULL 225 gms OF NEW HARDWARE
$1.95
3/5.00

ELECTROLYTIC CAPS
WE BOUGHT A WHOLE TRUCKLOAD OF VARIOUS VALUES FROM 20 mfd to 100 mfd ALL FROM 15 to 50 VOLT RATINGS.
TRUST US!
8 PIECES
$1.00

2N 3772
150W 60V NPN POWER TRANSISTORS
Most popular unit for high current power supplies or audio work.
NEW–GUARANTEED
$1.75
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100V 60MA
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HARD TO FIND!
LOW VALUE HIGH WATTAGE WIREWOUND RESISTORS
3/1.00 .1ohm@10W .032ohm@10W 2ohm@10W

AIRCRAFT/MOBILE CLOCK TIMER KIT
A FULL 6 DIGIT LED READOUT CLOCK THAT GIVES 24 HOUR (ZULU) REALTIME AND ELAPSED TIME IN SECONDS, MINUTES AND HOURS UP TO 24 HOURS. READOUTS CAN BE TURNED OFF OR DIMMED. PC BOARDS ARE SMALL TO FIT INSIDE A STANDARD INSTRUMENT CASE (2¼" Dia.)
FEATURES:
Real Time (24 hr.)
Elapsed Time
2 MOS Chips
3" LED Readouts
Special Noise Filter
Quality plated boards
12V NEG GND
MK–03 **$26.95**
KIT IS SUPPLIED WITHOUT SWITCHES OR CASE

2N3904 NPN General Purpose Switching **12/1.00**

2N2222 UNMARKED TO–18 NPN **12/1.00**

2N4861 N CHANNEL FET GEN. PURPOSE FET **3/1.00**

6 DIGIT ALARM CLOCK CHIP
MM5375AB
NEW LOW PRICE!
$2.95
DRILLED AND PLATED PC BOARDS FOR 5375
Com Cathode – 1.50
Com Anode – 1.50

SCREAM!
AS LOUD AS YOU WANT, BUT OUR LITTLE WARBLE KIT IS LOUDER! COMPLETE DUAL TONE WARBLE ALARM WITH 10 WATT OUTPUT. INCLUDES ALL PARTS AND BOARD.
$2.50
GREAT FOR BURGLER ALARMS
6–15VDC
(Speaker not included)

AUDIBLE CONTINUITY TESTER KITS
A FANTASTIC LITTLE DEVICE THAT HAS A HUNDRED AND ONE USES AROUND THE SHOP AND HOME.
USES:
Code Practice **$1.95**
Continuity Test
Transistor Test
LED & DIODE Test
Tone varies with resistance
REQUIRES 3 to 6 VOLT BATTERIES (not included)

50 OHM COAX
SUPER QUALITY COAX SLIGHTLY LARGER IN DIAMETER THAN RG–74 BUT BETTER SPECS
100 ft. ROLL
$3.99

CONTEST ELECTRONIC DESIGN
Monthly Prize Given for best circuit.
1. All entries become property of BULLET.
2. Entry should include schematic and circuit description.
WATCH THIS SPOT FOR WINNER'S NAME!
$100. GIFT COUPON FROM BULLET

SORRY – NO COD's
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ADD 5% FOR POSTAGE
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ORDERS UNDER $10. ADD 60c HANDLING
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More Details? CHECK-OFF Page 134

flea market

FLEA MARKET sponsored by LIMARC, October 10, rain date Oct. 24. N. Y. Institute of Technology, Rte. 25A & Whitney Lane, Old Westbury, L. I., N. Y. Largest flea market in the N.Y.C. area. Talk-in on 25/85 & 52. $1.25 per buyer, $2.50 per seller space. Hank Wener, WB2ALW, 516 484-4322.

NOW YOU ALL COME, YOU HEAR. Where? Why ARRL Hudson Division Convention, November 13 and 14, Playboy Resort and Country Club at Great Gorge, McAfee, New Jersey. Many exhibits, giant indoor flea market, FCC and ARRL forums, FCC exams, special YL programs, technical sessions and a Saturday night banquet with Jean Shepherd, K2ORS, world traveler, columnist and famed radio and TV personality as the speaker. For information write to Al Piddington, WA2FAK, 4 Acorn Drive, East Northport, N. Y. 11733.

6TH ANNUAL SOUTHERN MINNESOTA SWAPFEST — Minnesota's largest ham gathering — Saturday, October 9th, 9:00-4:00 — Contact VARS, Box 3, Waseca, Minn. 56093 — Talk-in on 94 or call 507-835-2679.

SAVE THIS DATE: October 9, 1976 for the 22nd Annual VHF Conference. Western Michigan University, 1 & ET Bldg., Kalamazoo, Michigan. Registration 8 a.m. — full day of activity. Write: Dr. Glade Wilcox, Chm., W9UHF/8, Dept. of Electrical Engineering, Western Michigan University, Kalamazoo, MI 49008. Get on the mailing list for yourself and your club.

THE NEXT CEDAR VALLEY AMATEUR RADIO CLUB HAMFEST is Sunday, October 3, in Cedar Rapids. Manufacturers and dealers welcome. ARRL representation. Held at Hawkeye Downs Exhibition Building, ample parking, camping. Tickets $1.50 advance; $2.00 at gate. First table - $3.00; others - $5.00. Saturday afternoon setup available. Talk in frequencies 146.16/.76, 146.52, 3.970 MHz. Prizes are Collins 75S-3C Receiver, Hy-Gain TH-3MK3 Beam, Wilson WE-224 Mobile XCVR, Wilson 1402SM H/T, Heathkit HW-8 QRP XCVR, plus more. Write: CVARC Hamfest, P. O. Box 994, Cedar Rapids, Iowa 52406.

INDIANA — Marshall County swap-n-shop. Sunday, October 31, 1976, at the Plymouth Indiana National Guard Armory located at 1220 W. Madison St. from 7:00 a.m. to 4:00 p.m. Talk-in on 146.07-67 and 146-94 simplex. Free tables, no charge for set-up! Tickets $2.00 at the door. Food, drink and door prizes. For further information contact WA9INM, Route 3, Box 526, Plymouth, Indiana 46563.

PASCAGOULA EXPLORER AIR SQUADRON will holds its annual Air Show Oct. 15, 16 and 17. In conjunction, the Jackson County Amateur Radio Club will operate mobile units and a multi-band radio station. The special call of KM5BSA will be applied for. Operators are needed. KM5BSA will also operate from here in the Nineteenth (19) International Boy Scouts Jamboree on the air. We hope any Explorers with an amateur license will plan to be present and take shifts as operators. Any scout or scouter so licensed is asked to contact W5UCY as soon as possible.

CEDAR VALLEY HAMFEST (Cedar Rapids, IA, home of Collins Radio) Sunday, October 3. Hawkeye Downs Exhibition Bldg. Prizes. Talk-in 146.16-.76, 146.52, 3.970 MHz. Advance tickets CVARC, P. O. Box 994, Cedar Rapids, IA 52401.

Stolen Equipment

REGENCY HR-2B; Serial #49-02817. Stolen from automobile. Microphone cable is hard-wired to set (instead of conventional cable-jack). Kristen N. Johnson, WA1TJP, 86 Alton Rd., Quincy, Mass. 02169.

EBC-144JR, 2 meter transceiver. Serial number 7514359A, and "Tuna Two" home-brew antenna with magnetic mount. Unit is inscribed on bottom with: Name: J. C. Maikisch; Call: WA2OFT; SS #: 089-32-6899; N. J. Dr. Lic. #: MO1814076312424; Tel. #: (201) 538-1667. Any info, contact above or Bernards Township Police (201) 766-1122.

More Details? CHECK-OFF Page 134

More Details? CHECK-OFF Page 134

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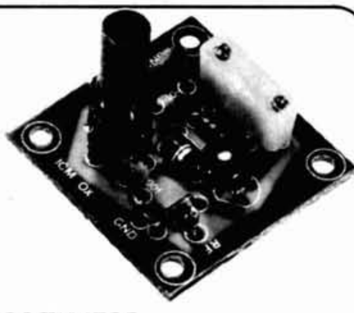
for the experimenter!

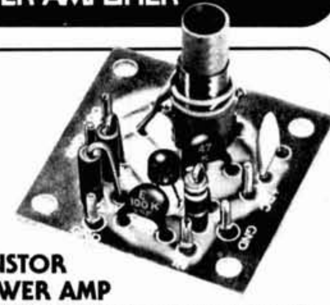
INTERNATIONAL CRYSTALS & KITS

OSCILLATORS • RF MIXER • RF AMPLIFIER • POWER AMPLIFIER

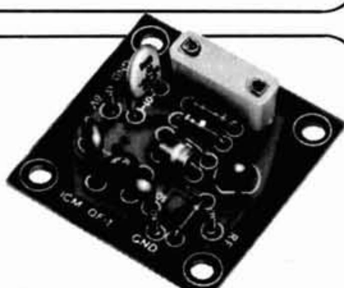


OX OSCILLATOR
Crystal controlled transistor type. 3 to 20 MHz, OX-Lo, Cat. No. 035100. 20 to 60 MHz, OX-Hi, Cat. No. 035101
Specify when ordering.
$3.95 ea.

MXX-1 TRANSISTOR RF MIXER
A single tuned circuit intended for signal conversion in the 30 to 170 MHz range. Harmonics of the OX or OF-1 oscillator are used for injection in the 60 to 179 MHz range. 3 to 20 MHz, Lo Kit, Cat. No. 035105. 20 to 170 MHz, Hi Kit, Cat. No. 035106
Specify when ordering.
$4.50 ea.



PAX-1 TRANSISTOR RF POWER AMP
A single tuned output amplifier designed to follow the OX or OF-1 oscillator. Outputs up to 200 mw, depending on frequency and voltage. Amplifier can be amplitude modulated. 3 to 30 MHz, Cat. No. 035104
Specify when ordering.
$4.75 ea.



OF-1 OSCILLATOR
Resistor/capacitor circuit provides osc over a range of freq with the desired crystal. 2 to 22 MHz, OF-1 LO, Cat. No. 035108. 18 to 60 MHz, OF-1 HI, Cat. No. 035109
Specify when ordering.
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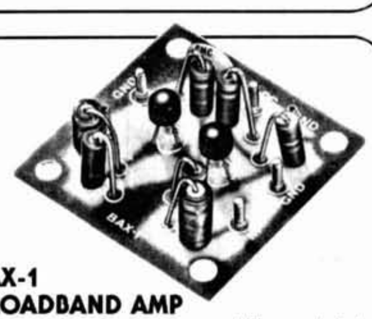
SAX-1 TRANSISTOR RF AMP
A small signal amplifier to drive the MXX-1 Mixer. Single tuned input and link output. 3 to 20 MHz, Lo Kit, Cat. No. 035102. 20 to 170 MHz, Hi Kit, Cat. No. 035103.
Specify when ordering.
$4.50 ea.



BAX-1 BROADBAND AMP
General purpose amplifier which may be used as a tuned or untuned unit in RF and audio applications. 20 Hz to 150 MHz with 6 to 30 db gain. Cat No. 035107
Specify when ordering
$4.75 ea.

.02% Calibration Tolerance
EXPERIMENTER CRYSTALS
(HC 6/U Holder)

Cat. No.	Specifications	
031080	3 to 20 MHz — for use in OX OSC Lo *Specify when ordering*	$4.95 ea.
031081	20 to 60 MHz — For use in OX OSC Hi *Specify when ordering*	$4.95 ea.
031300	3 to 20 MHz — For use in OF-1L OSC *Specify when ordering*	$4.25 ea.
031310	20 to 60 MHz — For use in OF-1H OSC *Specify when ordering.*	$4.25 ea.

Shipping and postage (inside U.S., Canada and Mexico only) will be prepaid by International. Prices quoted for U.S., Canada and Mexico orders only. Orders for shipment to other countries will be quoted on request. Address orders to: M/S Dept., P.O. Box 32497, Oklahoma City, Oklahoma 73132.

ICM

International Crystal Mfg. Co., Inc.
10 North Lee
Oklahoma City, Oklahoma 73102

More Details? CHECK-OFF Page 134

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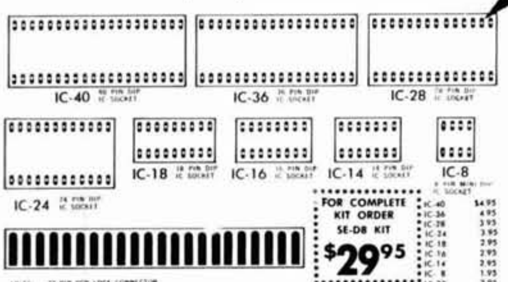

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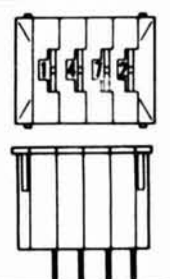

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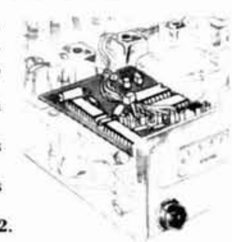

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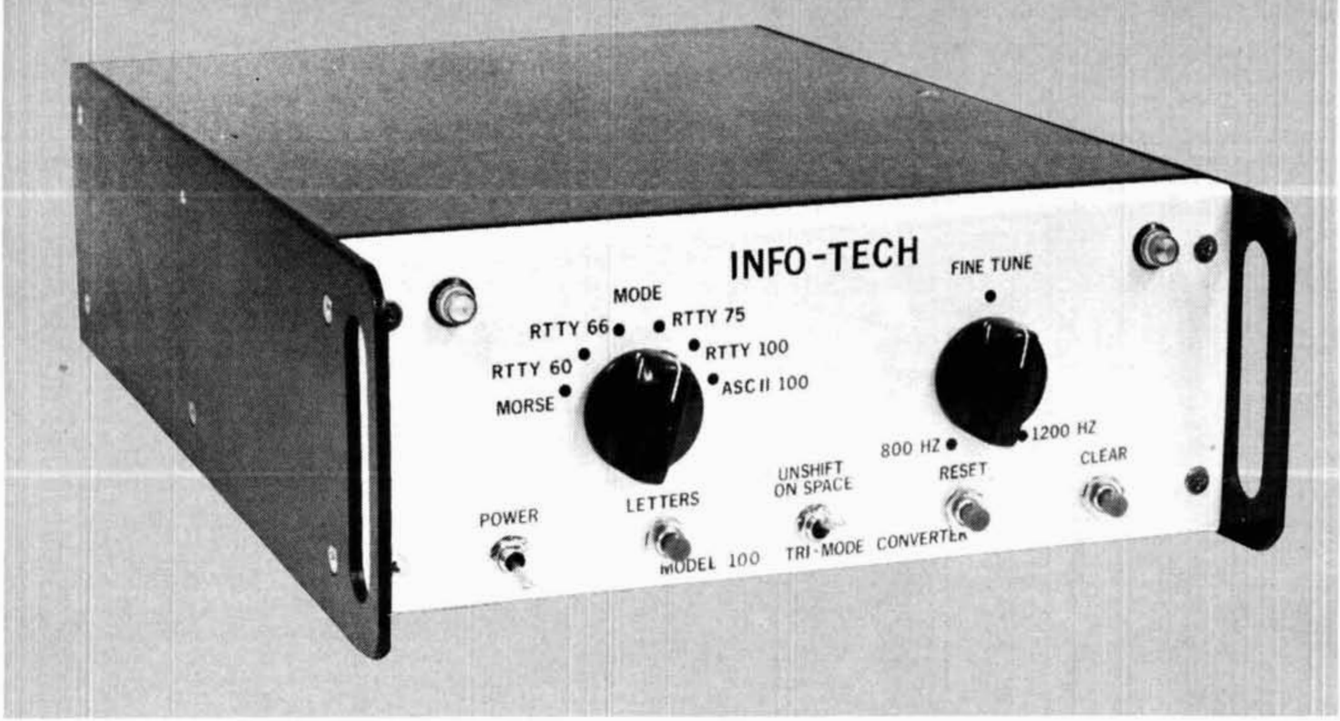

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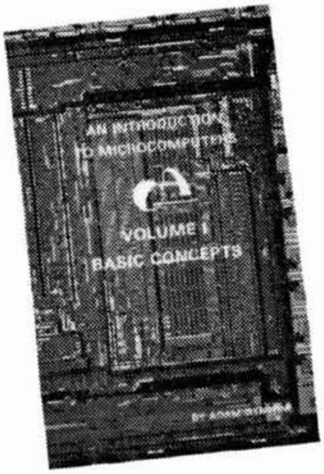

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October 1976

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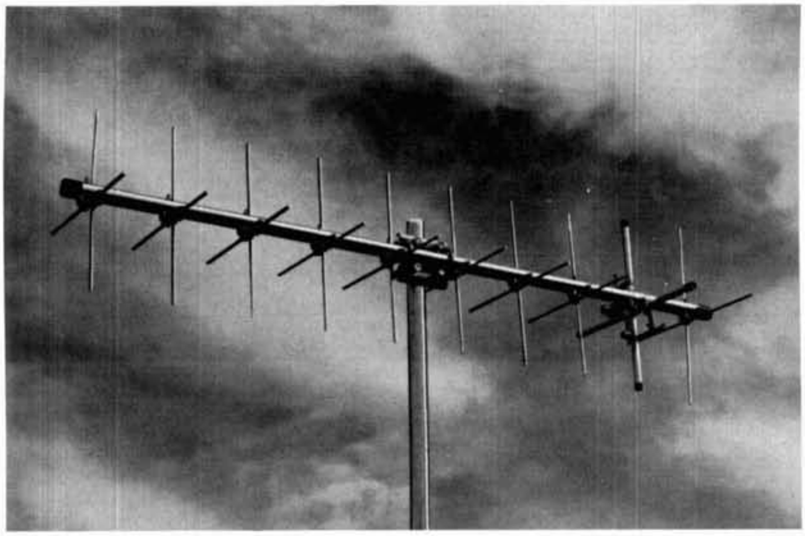
2 METERS...



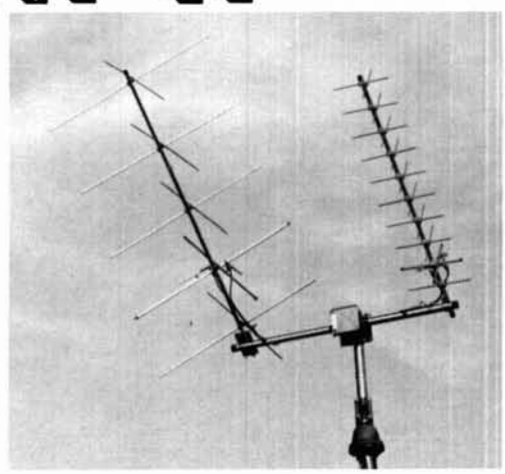
. . . Alive with activity at *both* ends of the band! Be a part of the total 2 Meter picture with the Cush Craft Twist Antenna. Actually two, easily assembled, 10 element yagis in one — the vertical elements are cut for the high end, the horizontal elements for the low end, and separate feed lines are used. The A147-20T is tailored to meet the demands of the operator who enjoys the best of both worlds — FM and SSB/CW.

A147-20T $47.50

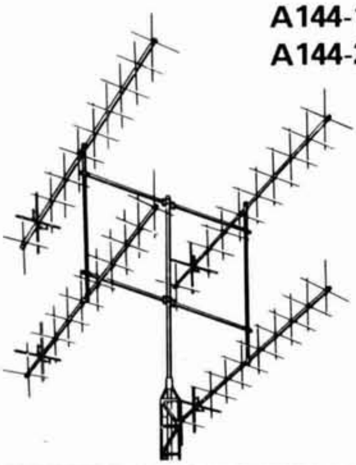
CQ OSCAR ....

. . . Extend your horizon — Explore the exciting new world of amateur satellite communications using low power with our multi-polarized Twist Antennas. All models include phasing harness for selectable linear or right/left circular polarization. Two of these Twists may be mounted on the A14T-MB mounting boom which is complete with a pre-drilled plate for a readily available mast-through rotator.

Face this challenging frontier—Become a Specialist!



A144-10T $32.50 A432-20T $45.00
A144-20T $47.50 A14T-MB $13.95



PERFORMANCE ARRAYS . . .

Enjoy fade-free contacts on VHF/UHF with Twist Antennas and Arrays. Excellent for scatter and other long-haul techniques. Double your effective radiated power by stacking two Twists, or quadruple ERP by stacking four Twists. Arrays are easily assembled for your special communications requirement. Write for stacking and phasing harness details concerning amateur and commercial frequencies.

Dependable communications — *Now Yours!*

SPECIFICATIONS

Model	A147-20T	A144-10T	A144-20T	A432-20T
Center Freq. (MHz)	144.5/146.5	145.9	145.9	432
No. Elements	10/10	10	20	20
Weight (lbs.)	6	3.5	6	3.5
Wind Surf. Area (ft.²)	1.42	.74	1.42	.37
Mounting	Center	Rear	Center	Rear
Dimensions (Inches)	40x40x140	40x40x70	40x40x140	14x14x57
Front-to-Back Ratio (dB)	22	22	22	22
Forward Gain (dBd) circular	—	10.8	13.6	13.6
linear	12.4	9.6	12.4	12.4

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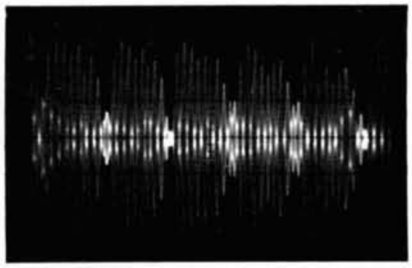
More Details? CHECK-OFF Page 134

400% MORE RF POWER

WITH THE NEW MFJ SUPER LOGARITHMIC SPEECH PROCESSOR

SSB Signal **Before** Processing. The weak valleys and high peaks give a low average power output, resulting in a real lack of punch.



SSB Signal **After** Processing. The once weak valleys are now strong peaks giving up to a 400% increase in average power output. The result is a signal full of punch with power to slice through QRM.

Up To 400% More RF Power is yours with this plug-in unit. Simply plug the MFJ Super Logarithmic Speech Processor between your microphone and transmitter and your voice is suddenly transformed from a whisper to a **Dynamic Output.**

Your signal has power to slice through QRM and you go from barely readable to "solid copy OM".

Look that happens to the RF power output on our NCX-3. It was tuned for normal SSB operation and then left untouched for these "before" and "after" oscillograms.

Three active filters concentrate power on those frequencies that yield maximum intelligence. Adds strength in weak valleys of normal speech patterns. This is accomplished through use of an IC logarithmic amplifier with a dynamic range of 30 dB for clean audio with **minimum distortion.**

The input to the LSP-520BX and LSP-520BX II is completely filtered for RF protection. A 9 volt transistor battery provides months of operation. These quality units are designed and manufactured entirely in the U.S.A.

TWO MODELS TO CHOOSE FROM — PLUGS BETWEEN YOUR MICROPHONE AND TRANSMITTER

$49.95

LSP-520BX. 2-3/16 x 3¼ x 4 inches. Uses 9 volt battery. RF protected. 3 conductor, ¼" phone jacks for input and output.

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LSP-520BX II. Same as LSP-520BX but in a beautiful 2⅛ x 3⅝ x 5-9/16 inch Ten-Tec enclosure with uncommitted 4 pin Mic jack, output cable, rotary function switch.

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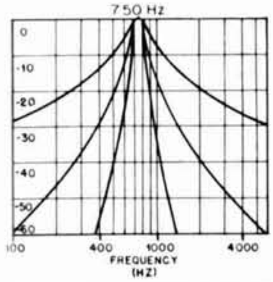
By far the leader. Over 5000 in use. Razor sharp selectivity. 80 Hz bandwidth, extremely steep skirts (See Response Curves).

● No ringing ● Bandwidth: 80 Hz, 110 Hz, 180 Hz (selectable) ● Skirt rejection: at least 60 dB down one octave from center frequency for 80 Hz bandwidth ● Center frequency 750 Hz ● Drastically reduces all background noise (up to 15 dB improvement in S/N ratio) ● 8 pole active filter uses ICs ● 9 volt transistor battery provides months of operation ● 400 Hz or 1000 Hz center frequency optional. Add $3.00 ● Size: 2-3/16 x 3¼ x 4 inches.



CWF-2PC wired PC board $18.95

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CMOS-8043 ELECTRONIC KEYER

● Uses Curtis 8043 Keyer IC ● Built-in key ● Dot memory ● Iambic operation with external squeeze key ● 8 to 50 WPM ● Sidetone and speaker ● Volume, tone, weight controls ● Solid State Keying ±300 volts max. ● Uses 4 penlight cells ● 2-3/16 x 3¼ x 4 inches.

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● Clearly copy weak unreadable signals (increase signal 3 to 5 "S" units) ● More than 20 dB low noise gain ● Separate input and output tuning controls give maximum gain and RF selectivity to significantly reject out-of-band signals and reduce image responses ● Dual gate MOS FET for low noise and strong signal handling abilities ● Completely stable ● Optimized for 10 through 30 MHz ● Uses 9 volt battery ● Beautiful Ten-Tec enclosure ● 2⅛ x 3⅝ x 5-9/16 in.

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MFJ-16010 ANTENNA TUNER

Now you can operate all bands — 160 through 10 meters — with a single random wire and run your full transceiver power output — up to 200 watts RF power OUTPUT.

● All bands 160 thru 10 meters ● Matches almost any random length of wire to any Transceiver or Transmitter ● Rated at 200 watts RF **Output** to handle virtually any transceiver running barefoot ● Ultra compact 2-3/16 x 3¼ x 4 inches ● Matches low and high impedances by interchanging input and output ● SO-239 coaxial connectors ● Unique wide range, high performance, 12 position tapped inductor. Uses two stacked toroid cores ● High quality specially designed variable capacitor ● Heavy duty construction ● All aluminum cabinet.

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