QEX

April 1982

Number 3

The ARRL Experimenters' Exchange

Well, this is the third issue of QEX. That may not seem to be earth-shaking news unless you have had experience getting a new publication off the ground. Many newsletters fail after a terrific premiere edition because they run out of material. That is not the case with QEX. We received enough for two issues before the first one went to press. More manuscripts have come in since then. At this point, after putting this issue together, we have sufficient material for the next one. So, it looks like there is plenty of support for a newsletter for Amateur Radio experimenters.

Correspondence from a few readers indicates that QEX is too much this or too little that, according to their expectations or desires. One such comparison is how much theoretical vs. practical coverage. Opinion is one thing. Getting the manuscripts in the 'right' mix is another. As a reader-generated publication, QEX is dependent on the doer/thinker taking time out from a very busy schedule to pick up the quill. So, if there is a subject - of interest to experimenters - which has not been properly covered, possibly you could get yourself up to speed to write such an article if you start researching the subject now. According to one definition, a 'genius' is the first person to write down something that everyone else knows.

QEX can also be used to ask experimenters how to do something - in order not to reinvent the wheel. If the item has not yet been invented, possibly merely asking would stimulate someone into designing it and publishing the results in QEX. That's the essence of several converations that I have had with ARRL Atlantic Division Vice Director, Hugh Turnbull, W3ABC. Several items on his wish list are:

An off-the-air monitor for the various bands and modes (hf-ssb, vhf-fm, etc.) in common use today to check whether your signal is clean and legal.

An inexpensive instrument to measure deviation and output power of a vhf-fm or uhf-fm transmitter.

A method of calibrating a frequency counter.

If you have solutions to any of the above, please write an article about it.

Switching subjects, some weeks ago, I had a conversation with Dr. Michael Marcus of the FCC's Office of Science and Technology. He is very interested in seeing Amateur Radio break new ground and help solve technical problems through experimentation. Dr. Marcus was instrumental in getting hams to experiment with spread spectrum. He now suggests that hams look into what seems to be a barrier to the use of high signaling speeds for mobile users of the vhf/uhf spectrum. If the ends of a circuit are fixed, vhf and above can be used to send data at terrific rates - megabauds! But if the receiver or transmitter is mobile, the practical speed appears to be limited to around 1000 baud due to fading and intersymbol interference. In fact, commercial paging services which not only 'beep' you these days but can send you a ten-digit phone number or other code, keep their signaling speed around 200 baud or so. Here is an area where Amateur Radio experimenters can help move technology ahead. A practical ham application might be high-speed paging of individuals carrying handheld transceivers. Another could be how to send packets at 1200 baud and above to/from mobiles. What's the solution? Error-correcting codes? A new modulation scheme? Right now, we're limited by FCC rules to 1200 baud ASCII on vhf. But that could change very soon with favorable FCC action on Docket 81-699 (ARRL petition RM-3788) which would permit new codes and higher signaling speeds on vhf.

If you have some ideas along the above lines, we'd welcome an article on your solution in QEX. On the other hand, possibly you have something to add to the wish list. - W4RI

A Multimode, Fm-able VXO

By Worthie L. Doyle,* N7WD

Summary

Here is a crystal oscillator with the following features:

(1) It requires no tank circuits, yet can be adjusted so the crystal oscillates at its fundamental or overtone frequencies.

(2) It can produce output at frequencies fj/N (ed. note., read f sub j throughout), where N is a small integer and fj (j=1, 3, 5...) are the crystal fundamental and overtone frequencies.

(3) It can be fm-ed directly, without the use of voltage-controlled reactances.

Circuit Operation

The circuit appears in Fig. 1 and makes use of a one-shot multivibrator. Discussion centers around the TTL 74121, but other one shots should be similarly usable.

For the moment, ignore both the crystal and the audio input and recall that the one shot will generate a pulse after either of two events. First a pulse will be produced if the B input is above a certain threshold and one or both of the A inputs brought from high to low. Second, if one of the A inputs is held low, a pulse will be produced if the B input rises from below to above its threshold. Trigger action at B has hysteresis of about 0.2 volt, so the swing at B must go below the lower hysteresis level and above the upper to insure triggering. It is assumed that bias at B has been adjusted to about 1.6 volts, so that B is a little above the trigger hysteresis interval when the circuit is in the resting state. The value is not critical, but if your oscillator doesn't go, this setting should be checked.

Pulse length is determined by the R-C time constant formed by C and an internal resistance of about 2000 ohms between pins 9 and 11 of the one shot. External resistors can also be used; interested readers can consult the IC data books. For rf use, short pulses are desired, implying small resistance values. The internal resistance is near the minimum suggested value and is convenient to use.

Here is the reasoning that led to the oscillator circuit of Fig. 1. Suppose that the key from A1 to ground is open. The one shot is at rest, and Q is low. When the key is closed, the first set of conditions for a pulse is satisfied, and Q goes high for a time determined by R-C.

Now consider the effect of this pulse on the crystal. The crystal terminal at B is initially at about 1.6 volts. Q has just gone from about 0.22 to 3.3 volts, so that the crystal has received about a 3-volt tap and is presumably vibrating like a struck gong. This will occur regardless of the duration of the pulse; however, it is clear that there is some optimal pulse length that will induce the greatest amplitude of vibration. Simple physical reasoning suggests that the greatest energy will be imparted to the crystal if the pulse length is half the crystal's natural period of vibration. If the pulse is longer, it eventually begins to push in the opposite direction to that desired, since the crystal will have swung away from the tap, so to speak, and will now be coming back toward it after half a period. Shorter pulses, down to perhaps an eighth of a period, will also have advantages to be discussed later.

Let's take an example. Suppose the crystal frequency is 7.1 MHz. The period is then about 140 ns. Half the period is 70 ns. From the 74121 curves we read that the standard value of 33 pF will produce a pulse length between 65 and 70 ns; hence this will be a suitable value for C (a low-power microscope is useful for reading the curves in manufacturers' data books).

Detouring momentarily, we can estimate the highest useful frequency for the 74121. Doubling the data book's typical propagation delay time of 35 ns from B input to Q output, we get a minimum period of about 70 ns, corresponding to a maximum frequency of about 14.3 MHz. Of course there will be variations among individual devices, but apparently most 74121s should go at 14+ MHz.

So far we have explained how the crystal starts when the key is closed. Now we need to explain what keeps it going. With A1 held low and the first pulse ended, the voltage at B will have swung up, then back down, and now after one full period will be swinging back up again from energy stored in the crystal. This brings B up through the trigger threshold, and a new one-shot pulse is initiated according to the second set of conditions for producing a pulse. This process will repeat on every cycle of crystal oscillation as long as A1 remains low.

*1120 Bethel Ave., Port Orchard, WA 98366.

Although the one shot is not an amplifier, it may help to notice that the crystal feedback branch is like a series-resonant circuit connecting the output to the input of a non-inverting amplifier (B and Q move in the same direction). Thus the feedback configuration is what would be suitable if the one shot were an amplifier.

Related Frequencies

While reading about the fundamental-frequency crystal oscillator, the reader may naturally have wondered what happens if the timing C is made too big or is otherwise varied. To check this, the circuit of Fig. 1 was used to experiment with three crystals with fundamentals of 1942, 7847 and 18915 kHz. Ths internal 2 k-ohm timing resistor was used, with external C that could be varied from a minimum of about 10 pF to a maximum of 720 pF (three different variable Cs).

With sufficient C, it was easy to get stable, keyable oscillator at f/2 and f/3. To get 971 kHz with the 1942 crystal, for example, took most of the 720 pF. With this crystal, stable output was also obtained at the third overtone, about 5705, and half of it, about 2853, as well as the fifth, 9485, and half of that, 4742.

A similar array of frequencies was available from the 18915-kHz crystal. Multiples which are the ratio of small integers occured: 3/10, 3/5, 1/2, 1/3, 1/4, 3/4, 3/7, and 5/12 were all measured, with values of C between 20 and 40 pF. My 74121 apparently will not go at 18915 kHz but did produce 14220 (3/4f) with this crystal. If nothing else, the multi-mode, one-shot crystal oscillator should greatly extend the range of one's crystal junk box.

Explanations

The simple explanation offered for the fundamental-frequency oscillation will obviously not do for strange ratios like 3/4. Here is a try, based on the response of a series-tuned circuit to step excitation, (ref) and specifically for the 74121.

Near any natural frequency the crystal can be regarded as a series L, R, C circuit, with R including the circuit loading (this oscillator uses the crystal in series resonance). Although the one-shot pulses have finite rise and fall times, they may be considered a positive step followed very shortly by an equal negative step. Now the response of series R, L, C to a step is a damped oscillation at the circuit's natural resonant frequency. Since the crystal has very high Q, even after circuit loading, the response is essentially undamped for the few cycles that matter here.

Each edge of the one-shot pulse produces such a response from the crystal. The oscillation from the falling edge is identical to that from the rising edge, but opposite in polarity and delayed by the pulse length. These results can be summarized by saying that after the pulse has ended, the crystal is swinging with the superposition of two equal oscillations having a phase difference determined by the pulse duration.

It is immediately clear now why some pulse lengths produce no oscillation. There wil be a series of such lengths that lead to near cancellation of the two superposed oscillations. If these do not also have enough energy at one of the crystal's overtones or spurious frequencies, no sustained oscillation will ensue. Since it takes 0.2 volt to swing B through the hysteresis region, there is a threshold of crystal response below which the one shot is not stimulated to produce a second pulse.

The superposition of two sinusoids of the same frequency but different phases is another sinusoid with the same frequency, but amplitude and phase determined by the relative phases of the two originals. Fig. 2 illustrates crudely what happens after the first pulse. Notice that the second one-shot pulse occurs at $T1 > T0$ in this example, so that the period of oscillation, whatever it eventually settles down to, would be expected to be greater than the natural period, T0.

Fig. 3 illustrates the situation when producing fj/2. Action here is similar to that of low-frequency pulse dividers using timers like the 555.

Overtone Excitation

As noted above, the one-shot pulse will have a local maximum of energy at a particular frequency if the pulse length is half the corresponding period. This rule can be extended, for overtones particularly, by noting that energy also goes through a maximum when the pulse length differs from half the period by a whole number of periods.

This extended rule can be used to estimate C required for overtone excitation. Consider, for example, the third overtone of an 18915 crystal. The 74121 cannot begin to produce a pulse short enough to have maximum energy at its third overtone. Indeed, the circuit of Fig. 1 would not even produce its fundamental. However, a local energy maximum at the third overtone will occur when the pulse length is about 7/6 of the fundamental period, T. Since a half cycle at the third overtone lasts about 1/6 T, this pulse length should lead to output with a period near 8/6 T or frequency near 3/4 f1, i.e. at f3/4. This is in fact one of the output frequencies obtained with the 18915 crystal, approximately 14220. Indicentally, 56.88-MHz energy should be extractable at point B in the circuit, despite 74121 limitations.

Since 5/6 T also differs by about half a period from the period of the third overtone, it should also excite the third overtone preferentially. The output period expected is 4/6 T, corresponding to f3/2 or half the third overtone. This could not be achieved with the 18915 crystal but was obtained with lower-frequency crystals.

In practice the rule is only used to make a guess at C. Actual C will be a trimmer that is set for most-reliable of the desired mode. Except for fundamental oscillators, such a trimmer will practically always be needed.

The above explanation is undoubtedly oversimplified. Performance will also be affected by the relative enthusiasm of the crystal, in its mounting, for its various modes of oscillation.

VXO Use

If you watch frequency while varying C, say on a counter with 0.1 second or shorter counting interval, it will be obvious when you are in a range of C that corresponds to a stable mode of oscillation controlled by the crystal. Frequency will vary slightly and smoothly as C is varied, with mode jumps or cessation of oscillation at each end of the appropriate range of C. Behavior is like that of a regular VXO (variable crystal oscillator). With the 7847 crystal mentioned earlier, a 14-to-50 pF variable gave frequency variation from 7847 to 7832 kHz.

In a practical VXO it will be necessary to restrict the range of C to avoid embarrassing cessation of oscillations, otherwise it will be necessary to unground A momentarily to restart oscillation. This inconvenience does not arise if the oscillator is keyed at A; however, range of C should still be restricted to that which gives only the desired mode.

Obviously, frequency can also be varied by reactance in series with the crystal, as for any series-mode crystal oscillator. Such a variable L or C might be useful to trim the VXO range for calibration setting.

Direct Fm

Since a new pulse is produced at the instant that the voltage at B rises through the trigger level, varying the voltage at B will lead to a varying frequency. To check the amount and linearity of frequency variation, trials were made with three different crystals.

The first crystal was marked 9030. In a range from about 1.2 to 2.2 volts, frequency varied linearly over a range of 1240 Hz. The sixteenth harmonic was tuned in on a 2-meter fm receiver. It sounded clean and quiet. Next, an audio oscillator set to 800 Hz was connected to the audio input in Fig. 1. To produce a loudness that sounded about equal to one of the occasional clean signals on the band took about 0.15 volt peak af. From the dc frequency measurements, this should correspond to a peak deviation of about 3 kHz at 2 meters, which seems reasonable. With this crystal, oscillation stopped when the voltage at B was varied below 0.8 or above 2.6 volts.

The second crystal tried was marked 6091. It oscillated with B anywhere between 0 and 3 volts, but frequency varied only by about 160 Hz per volt. To get 3 kHz deviation at 2 meters would have required about 0.75 volt peak af.

The third crystal, evidently third overtone, was marked 36.000. It oscillated at its fundamental with voltage at B between 0.75 and 2.25 volts. Frequency variation was an amazing 8.8 kHz per volt. This confirms what others have noted - that overtone crystals seem to make the best VXO crystals. One of the interesting possibilities with such a crystal is a transceiver with the fm signal generated directly at 10.7 MHz (using a third overtone crystal at about 32.1), and then converted to the operating frequency rather than multiplied up. For a peak deviation of, say, 3 kHz at 10.7 MHz, this crystal would require only about 0.38 volt peak af.

From these limited trials, it appears that this VXO, at least at a fundamental frequency can simultaneously provide frequency modulation, VXO action by varying C, and, if desired, calibration adjustment by setting a reactance in series with the crystal, all with a remarkably simple circuit.

Odds and Ends

If an A is wired to ground, the one-shot oscillator normally starts when power is applied. This could be due to the R-C delay at the B input from the rf-bypass capacitor at the audio end of the rf choke. This would cause the voltage at B to move slowly up through the trigger threshold just after power is applied, thus initiating the first pulse.

Some experiments were also run with one section of a 74123. With this one shot, best results were obtained with B biased at about 0.75 volt, a puzzling circumstance. All the one shots used for thise trials were from a plastic bag of "U-test-em-n-weep" devices, although results seemed consistent among functioning devices of the same type. The 74123 is retriggerable, while the 74121 ignores its inputs until its present pulse has ended, so some modes might be possible with one device but not the other.

A useful experimental adjunct is a 50-mA meter in a supply lead to signal whether oscillation is in progress.

Since pulse length depends on R-C, it can be adjusted with R, holding C fixed. I haven't tried this approach because the usual dirty, noisy pot seems likely to lead to random pulse jitter, hence more oscillator noise.

It should also be pointed out that the one-shot VXO is probably unsuitable for applications requiring very-low-noise sidebands. The crystal stabilizes the pulse repetition frequency, but the R-C timing still controls the pulse lengths, which will entail some fluctuation, though the 2-meter signals sounded clean,

I hope that some readers will find this curious oscillator useful and interesting. Clearly there is room for more experimentation. Although these ideas are original with me, it would be unusual if something similar had not been published somewhere. I will appreciate any literature references from readers with access to engineering publications.

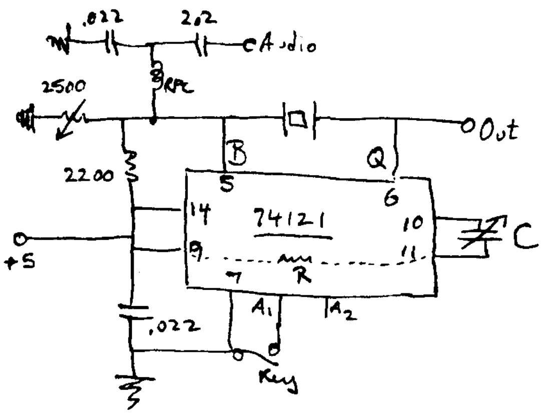


Fig. 1 Multi-mode, FMable VXO

Reference

Guillemin, "Introductory Circuit Theory," Wiley, 1953, pp 243-251.

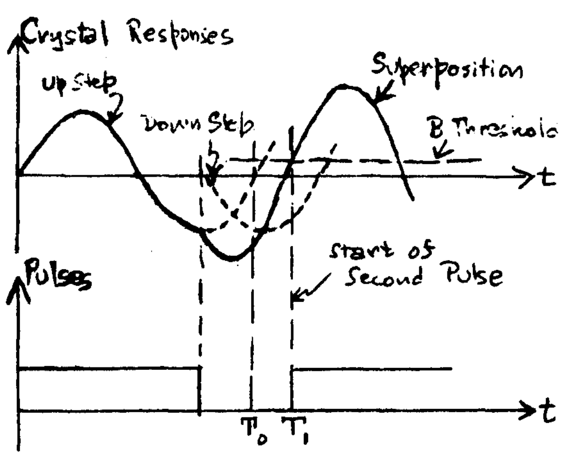


Fig. 2 Crystal Response to Pulses

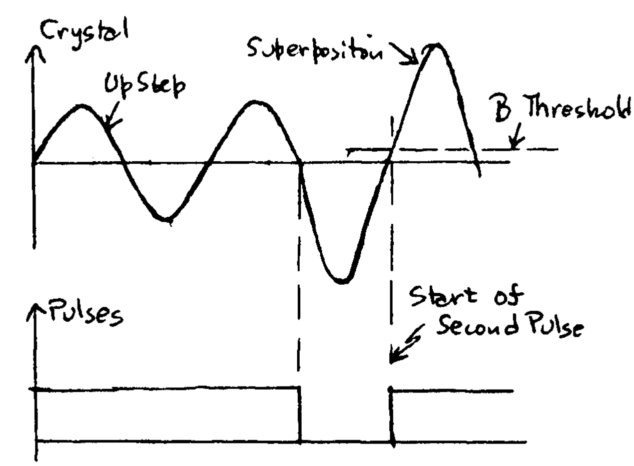


Fig. 3 Division by 2

Wish List

The application of a Dual-Tone Multi-Frequency (DTMF) low-band filter as a cw audio filter does not seem to have been exploited in ham literature.

As you are probably aware the dual tones are one from the low group (697 to 941 Hz) and one from the high group (1209 to 1477 Hz) sent together. In a three-chip decoder the first two chips are a low-bandpass and a high-bandpass filter, with both outputs sent to the decoder itself. The low-bandpass filter, with both outputs sent to the decoder itself. The low bandpass is really a little wider, 680 to 960 or thereabouts, to allow for frequency tolerance, pass band ripple is very small, and cut off is very steep on both sides. Those bandwidth, ripple and skirt characteristics make it almost ideal as a cw audio filter. It would also do away with the bulk of an L-C filter and the critical component tolerances of an op-amp active filter. Would a QEX reader like to design one? - W.A. "Spud" Monahan, K6KH, 817 Pacific Ave., Manhattan Beach, CA 90266, 213-642-1143 X2312 (work), 213-374-8289 (home).

A PTO Substitute for the R-390A

By W.S. Hoehl, Jr.,* WB4MUZ

A large number of Collins R-390A receivers have recently been released through the MARS program. Unfortunately, these receivers were cannibalized of there permeability-tuned local oscillators before they were released. This converted them into first-rate boat anchors. The correct PTO unit for them occasionally appears on the surplus market, but it is both hard to find and expensive. This article describes a PTO substitute which is almost as precise as the original PTO and which will restore these receivers to useful service.

One major advantage of permeability-tuned oscillators is their linear tuning characteristic. This is achieved by giving the oscillator coil a variable-pitch winding. A cam, which must be individually adjusted for each unit, compensates for winding irregularities by adding to, or subtracting from, tuning leadscrew rotation as a function of slug position in the coil.

The PTO substitute is tuned by a Varicap diode. Varicap-tuned oscillators have an approximately logarithmic control voltage vs. frequency characterestic. The tuning control must generate a voltage which is the inverse of this function - an exponential - to provide the desired linear tuning characteristic.

Several possible methods for generating this exponential function are shown in Fig. 1. The first makes a straight-line approximation of the desired curve with a diode function generator. Its chief drawback is the large number of line segments needed to reproduce the function accurately. A generator with ten segments would deviate from linearity by about 5 kHz at several points. The second approach uses analog computer techniques to calculate a power-series approximation of the function. This method is frequently used in instrumentation to provide a linear readout from a non-linear measurement, like a thermocouple. It is capable of a high degree of accuracy; a third-order approximation would, theoretically, deviate from linearity by no more than 3 kHz. A fourth-order approximation would deviate less than 500 Hz. However, a large number of components are required, and the network is difficult to adjust properly.

The third approach combines a truncated power series of the form

$$A_0 + A_1/X + A_2/X^2 + \ldots$$

with the diode function generator. Only the first two terms are used. This series fits the exponential function over about 60% of the tuning range. A two-breakpoint diode function generator is used to correct the end points. This method requires fewer components than the others mentioned and is easy to adjust. Fig. 2 shows its theoretical deviation from linearity along with the measured deviation of the PTO substitute using this technique.

Circuit Description

Fig. 3 shows the circuit of the PTO substitute's oscillator section. It tunes from 2.455 to 3.455 MHz as the control voltage is varied from about 2.2 to 22.5 volts. Transistor Q1 is the oscillator; transistors Q2, Q3 and filter L2, L3 buffer the oscillator and couple it to the receiver's third mixer. A Hartley oscillator circuit was selected to minimize shunt capacitance across the oscillator tank circuit. The 2:1 capacitance change required by the tuning range is about all that can be realized with readily available tuning diodes after temperature compensation and circuit stray capacitances are allowed for.

The linearization network is shown in Fig. 4. Operational amplifier U1 generates the A1/X term of the power series as the 10-turn tuning potentiometer is varied. Amplifier U2, diodes D1 - D4 and trimpots R4 - R7 form the two-breakpoint diode function generator. Trimpots R1, R2 and R3 adjust the shape of the control-voltage curve. R4 and R5 correct the low-frequency end of the tuning range; R6 and R7 adjust the high-frequency end correction.

Power for the PTO substitute is obtained by rectifying and regulating a 25-Vac line in the receiver. The power supply circuit is shown in Fig. 5. It delivers approximately 24 Vdc.

Construction

The oscillator/buffer is built on a 1 7/8 x 2 3/4 inch circuit board and mounted in a 1 1/2 x 2 x 4 inch minibox. The oscillator tank circuit is assembled on lugs in the coil form and mounted directly in the minibox. The etching pattern and parts layout for the oscillator board is

*219 Jackson Circle, Tullahoma, TN 37388.

shown in Fig. 6.

The power supply and linearization circuits are assembled on a 3 1/8 x 8 inch circuit board. The etching pattern and parts layout for this board are shown in Fig. 7. The oscillator/buffer unit and tuning potentiometer are mounted on the front of this board. Brackets attached to it mount the entire PTO substitute in the receiver. The general mechanical layout of the PTO substitute is shown in the photographs.

The tuning potentiometer shaft must be carefully aligned with the receiver's KILOCYCLE CHANGE tuning shaft. Its center is 2 1/32 x 2 2/4 inch from the bottom and right wall, espectively, of the receiver's PTO compartment. Be sure the mounting brackets are set square when assembling the unit. Elongate the screw holes in the main mounting bracket so the unit may be shifted from side to side for the best shaft alignment. A Johnson flexible coupling is used to connect the shafts.

Parts procurement is a major problem facing anyone who builds equipment these days. Most of the parts in the PTO substitute are readily available items. The only item that may be difficult to get is the subminiature co-ax connector for local oscillator input to the receiver's rf subchassis. This connector is an MB series, Amphenol part number 48850. The mate for the PTO power plug (P109) is Amphenol part number 126-012. Both of these connectors are listed by Newark and should be available from other industrial distributors.

Inductance values, as well as winding information, are given for coils so other forms may be substituted if necessary. L1 is wound on a 3/4-inch diameter form, National part number XR-72. L2 and L3 are wound on 5-mm diameter forms from a Radio Shack coil assortment package. Resistance values other than 10k ohm may be used for the tuning potentiometer if the values of R1 and R3 are adjusted to keep the same resistance ratios. Any value between 5k and 50k should be satisfactory. The tuning pot should be a Bourns model 3400 or equivalent. Smaller diameter pots and lower resistance values are not recommended because their reduced resolution may cause the receiver to tune with noticeable frequency jumps.

One final caution: do not use junkbox or bargain-counter parts in the linearization network, since it determines the ultimate frequency stability of the unit. a 1 mV drift in control voltage will shift the oscillator's frequency by about 150 Hz - enough to throw a cw signal completely out of the receiver's passband at maximum selectivity. Use 2%-tolerance or better metal-film resistors and good quality integrated circuits. Wirewound trimpots are preferable to Cermet-element trimpots, since they generally have a smaller temperature coefficient.

Receiver Modifications

Two receiver modifications are required to install the PTO substitute in the R-390A. First, the beat-frequency oscillator tube heater must be wired to the receiver's 6.3-V filament bus. This is done as follows:

1. Remove the i-f subchassis from the receiver.

2. Disconnect the wire on pin 7 of the ballast tube (to pin 4 of the BFO tube, V505) and ground it.

3. Install a wire from V505, pin 5 to V506, pin 9.

The PTO substitute derives its supply voltages from the 25-Vac source originally used for the PTO's oven heater. The wires from pins D and E in P109 (the PTO power plug) which supply this voltage must be moved from the normally open to the common terminal on the OVENS switch located on the receiver's back panel.

If you are unable to obtain the MB series co-ax connector to mate the rf subchassis, one may be salvaged from the receiver's i-f output cable. The MB-to-BNC adapter on the rear panel may be replaced with a chassis-mount BNC connector if it is desired to keep the i-f output feature.

Adjustment and Calibration

Alignment of the PTO substitute involves setting the oscillator to tune the proper frequency range, tuning the output filter, and calibrating the linearization network. The following procedure is suggested.

First, adjust the oscillator and filter. Temporarily disconnect the control input from the linearization board and connect the tuning pot directly across the 24-Vdc supply. Connect the pot wiper to the control input. Then connect the PTO substitute to P109 and turn the receiver on STANDBY. Couple a counter to the oscillator's output and set the control voltage to 22.5 volts. Adjust L1 for an output of 3.455 MHz. Then set the output to 2.455 MHz with the tuning pot and measure the control voltage. It should be about 2.2 volts. Adjust L2 and L3 so the oscillator's output level is reasonably flat over the 2.455 to 3.455 MHz range.

Reconnect the tuning pot and control input to the linearization board. Temporarily lift one end of diodes D1 and D3. Set R1 to 3.5k ohms, R2 to 650 ohms and R3 to 3.4k ohms.

The linearization network is calibrated as follows:

1. Set the tuning pot 2 turns clockwise.

2. Adjust R1 for an output frequency of 3.255 MHz.

3. Set the tuning pot to 7 turns clockwise.

4. Adjust R2 for an output frequency of 2.755 MHz.

5. Repeat steps 1 - 4 until both points are set.

6. Set the tuning pot to 4.5 turns clockwise and note the output frequency.

7. Take the difference between this reading and 3.005 MHz. Adjust R3 to shift the frequency about five times further away in the same direction. For example:

If F4.5 = 2.970 MHz,
delta = 2970-3005 = -35 kHz
5 delta = -165 kHz
F'4.5 = 3005 + (-165) = 2840

Adjust R3 for an output frequency of 2.840 MHz.

8. Repeat steps 1 - 7 until all three points are set.

9. Set the tuning pot to 1.5 and 7.5 turns clockwise. Note the output frequencies at these points. They should be about 3.307 and 2.702 MHz respectively. Then reconnect diodes D1 and D3.

10. Set the tuning pot fully counter-clockwise and adjust R6 for an output of 3.455 MHz.

11. Set the tuning pot 1.5 turns clockwise and adjust R7 until the output frequency either just starts or just stops changing. It should be about that measured in step 9.

12. Repeat steps 10 and 11 until both points are set.

13. Set the tuning pot fully clockwise and adjust R5 for an output frequency of 2.455 MHz.

14. Set the tuning pot 7.5 turns clockwise and adjust R4 until the output frequency either starts or stops changing, as in step 11.

15. Repeat steps 13 and 14 until both points are set.

Installation

The PTO substitute may now be mounted in the receiver. Set the KILOCYCLE CHANGE tuning control so the frequency dial reads XX.000. Set the tuning pot shaft fully counterclockwise. Couple the pot shaft to the KILOCYCLE CHANGE tuning shaft with a flexible coupling and position the unit for best shaft alignment. Then tighten the mounting screws. Turn on the receiver and tune it to WWV. If necessary, loosen the flexible coupling and turn the tuning pot shaft to set the end point exactly. Then check the high-frequency end point by selecting the next lower tuning range and finding WWV at its top. Be careful at the ends of the tuning range: 10-turn pots don't have any overtravel. Touch up the end-point adjustments, R5 and R6, if necessary.

A dial calibration curve for the PTO substitute may be plotted by tuning across the a-m broadcast band. Carefully zero beating the carrier in each channel provides 10-kHz check points across the entire tuning range. The dial reading should not show more than about 3.5-kHz error at any point.

Conclusion

The PTO substitute described in this article has been in use for several months. Although it is designed for a particular receiver, the same idea could be adapted to other pieces of equipment having similar oscillator requirements. Appendix A outlines the design procedure for the linearization network.

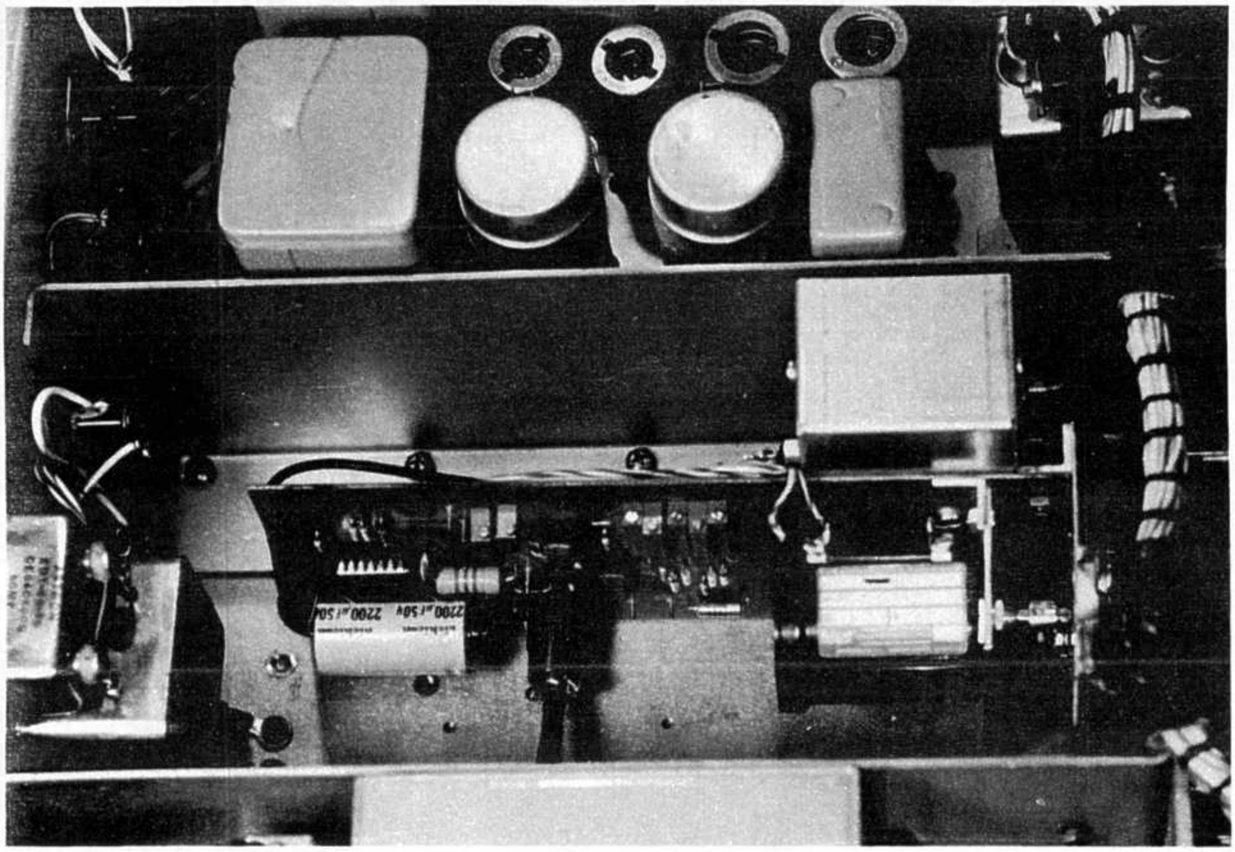
Reference

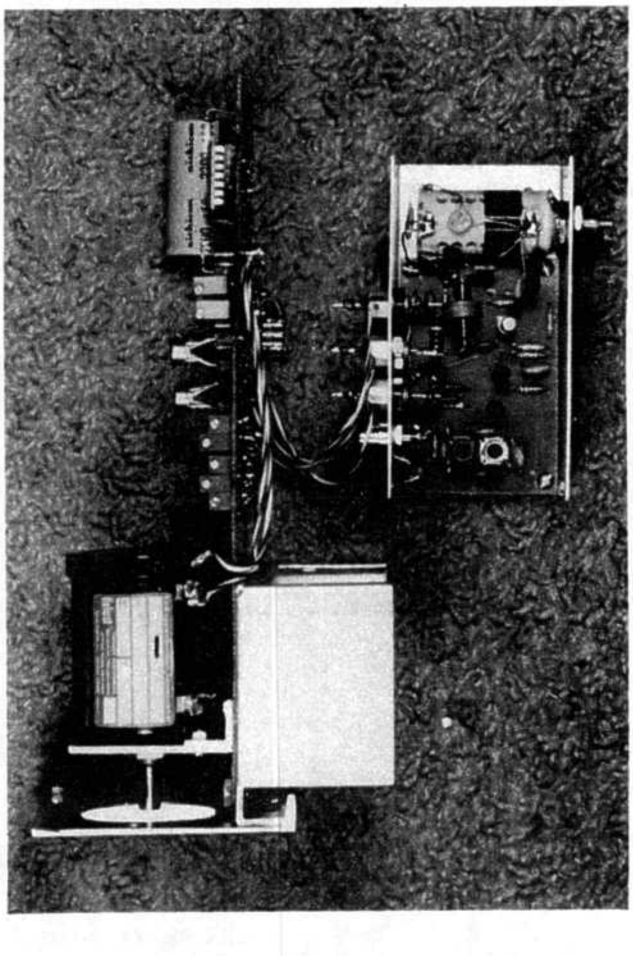
Graeme, Tobey, Huelsman, "Operational Amplifiers - Design and Application," Burr-Brown Research Corp., 1971.

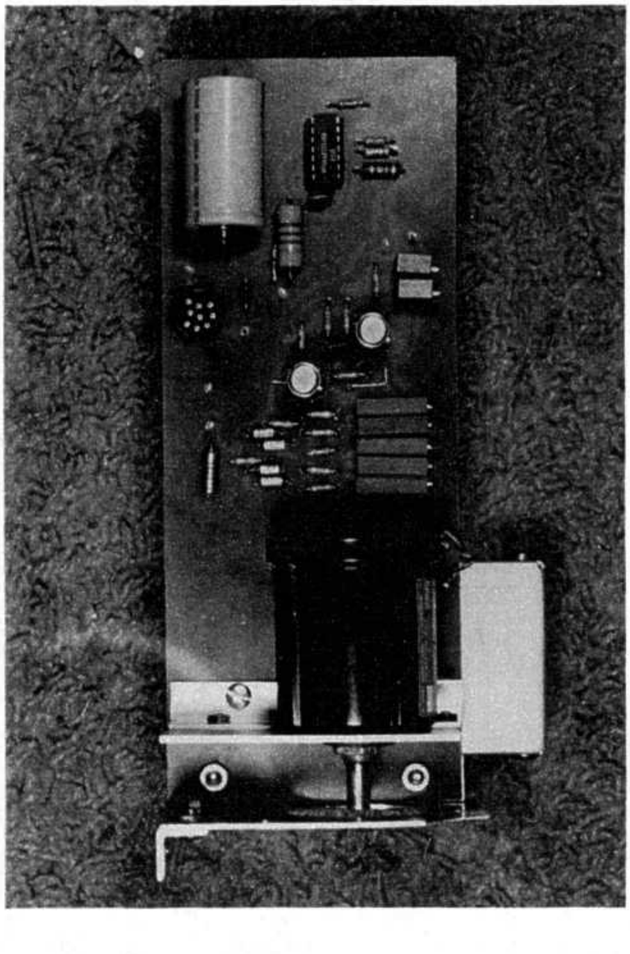
Literature

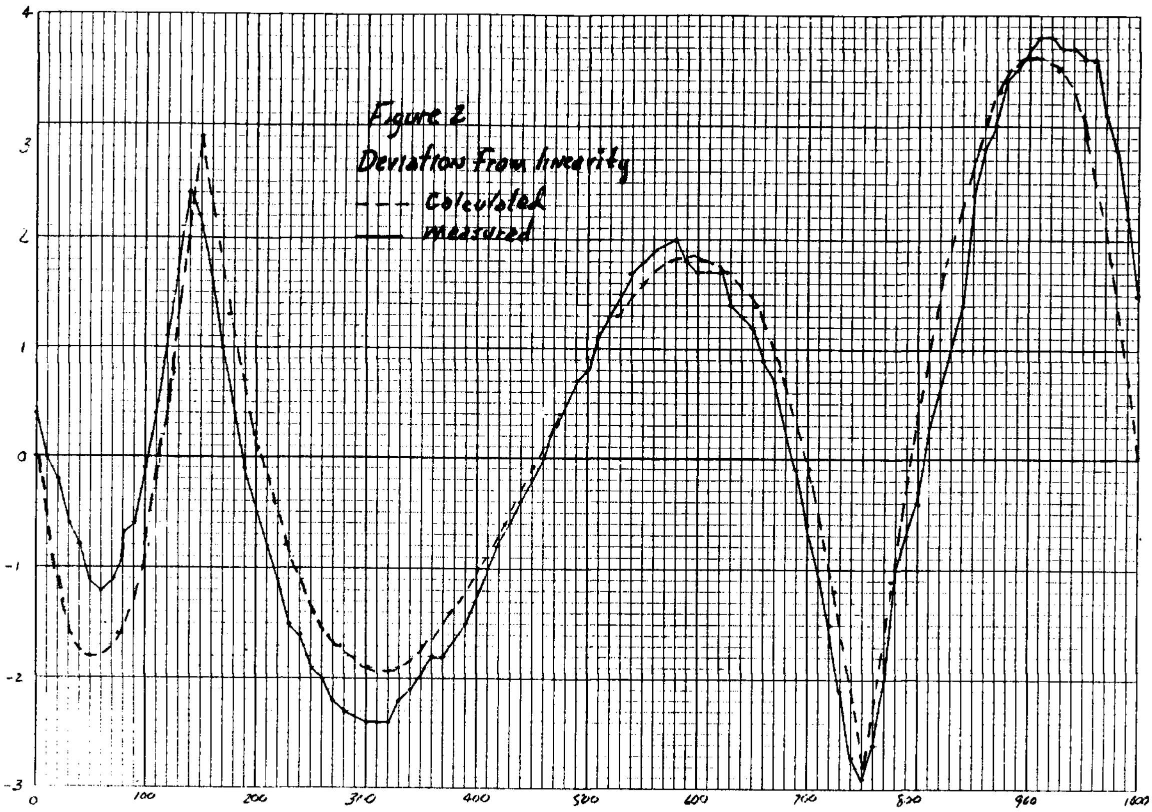
"Data Communications - A User's Handbook" is a new publication of graphic illustrations and to-the-point explanations of a number of data communications subjects. It is published by Racal-Vadic and is currently available by calling one of their dealers serving your area. Sections cover telephone systems, telecom services, why modems, network interface, modes and protocols, modems, automatic calling units/accessories, and diagnostics.

If you have difficulty finding your local dealer, contact: Racal-Vadic, 222 Caspian Dr., Sunnyvale, CA 94086, 408-744-0810.



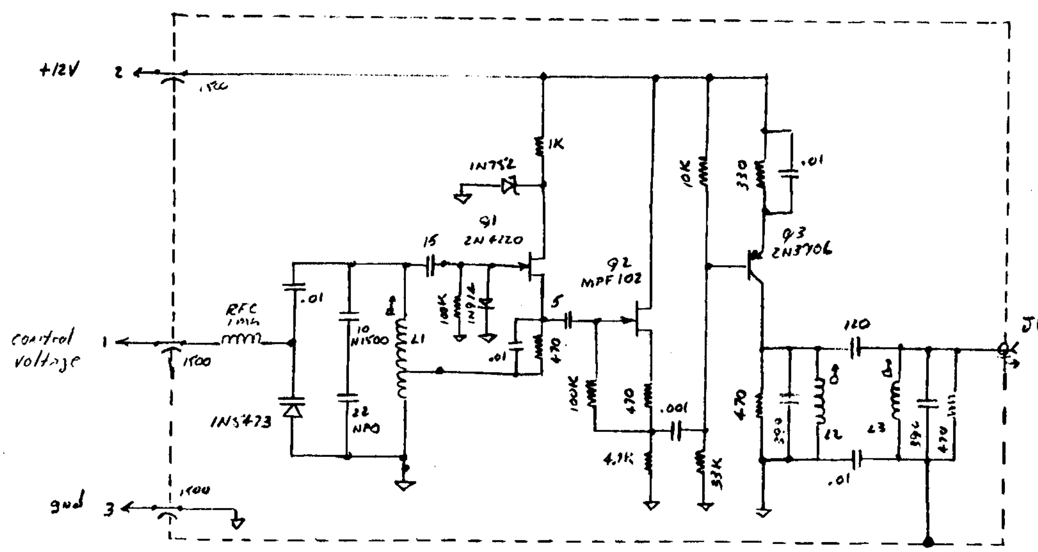


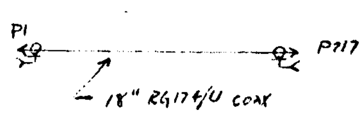




↓ = circuit board gnd

L1: 50 μh; 45 turns #34, tapped 15t from gnd; National XR72 form, 3/4" dia., close wound
L2, L3: 7 μh; 25 turns #34, 5 mm dia. form, close wound
J1: phono jack





Output cable:
P1: phono plug
P717: MB series; Amphenol #48850

Figure 3

↓ = ckt. board gnd.

D1, D2, D3, D4 - 1N914

R1-R7 : 15 turn trimpots. Bourns 3005P series or equivalent

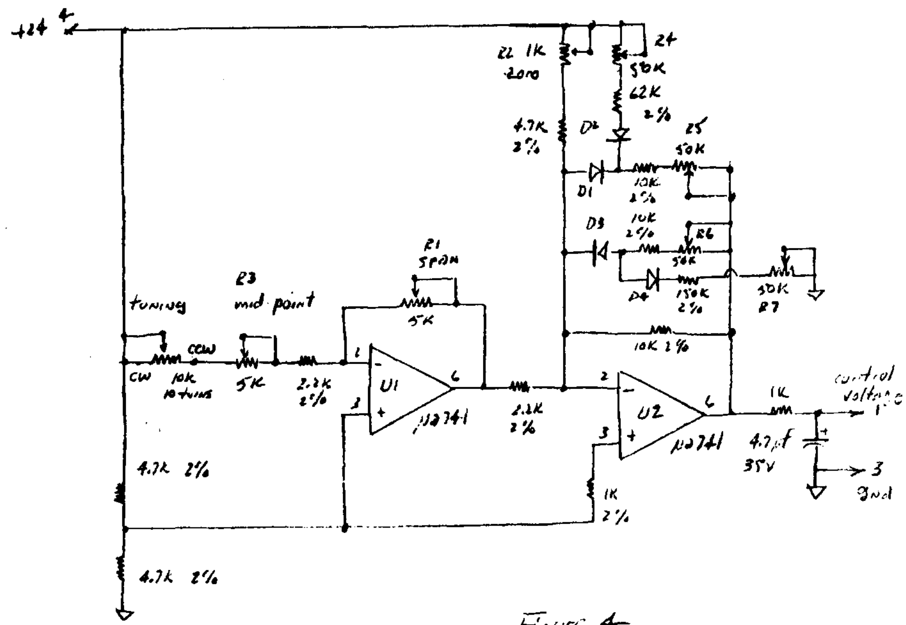
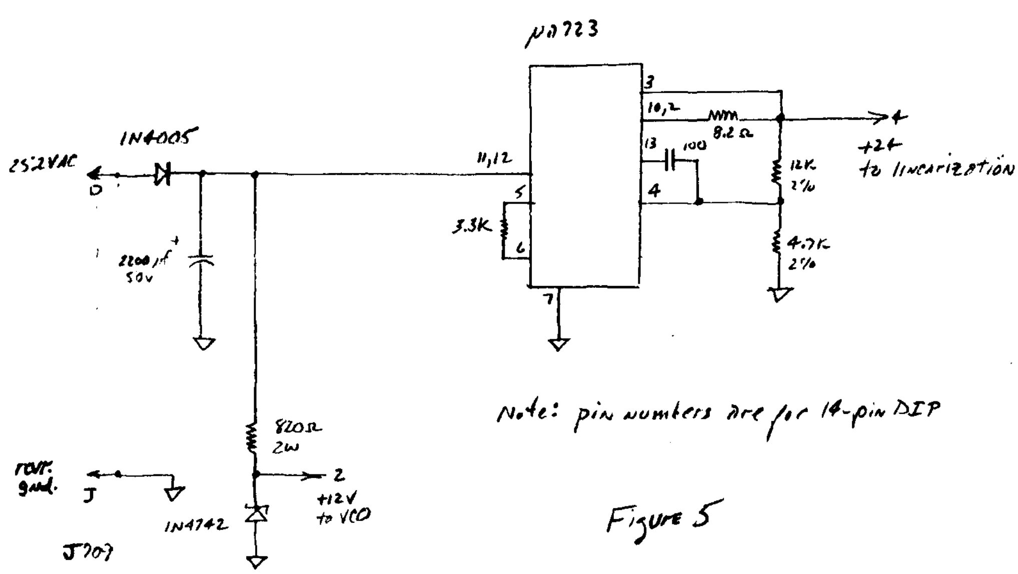


Figure 4

↓ = ckt board gnd.

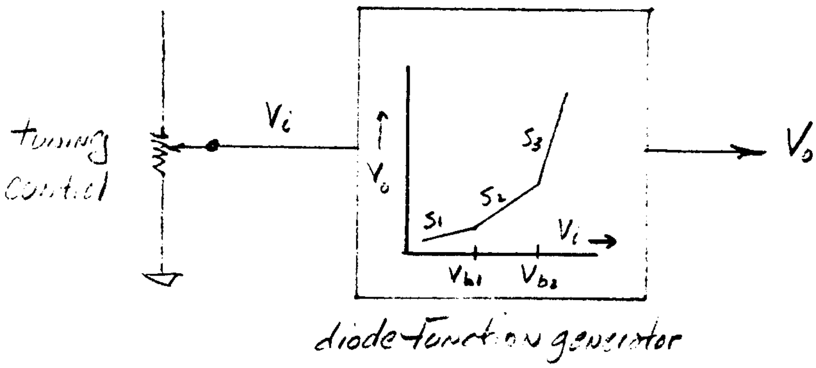


Note: pin numbers are for 14-pin DIP

Figure 5

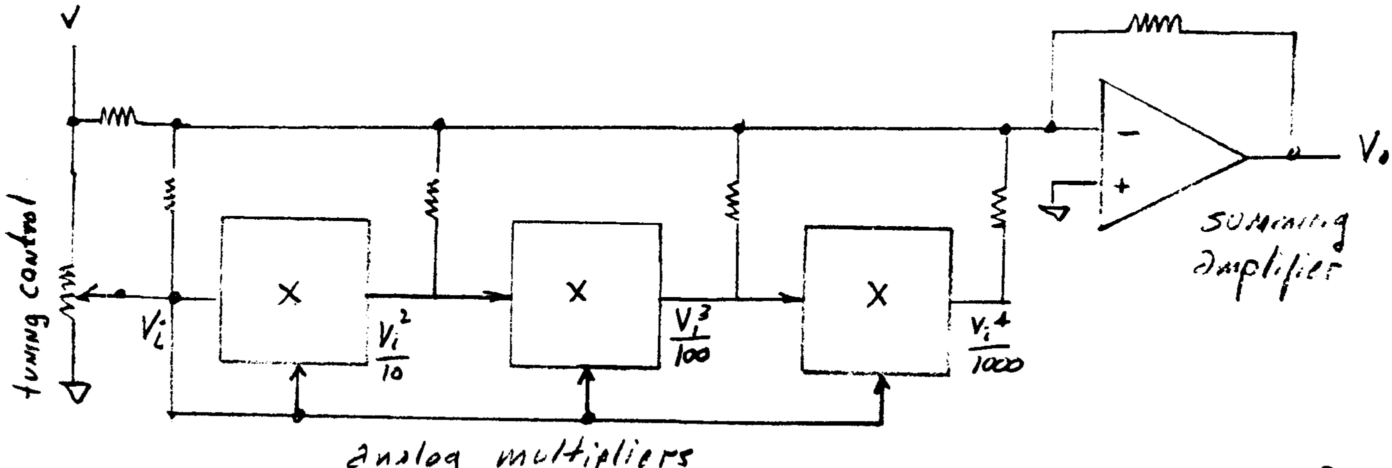
J709: Amphenol 126-012

Fig. 1 - Linearization techniques.



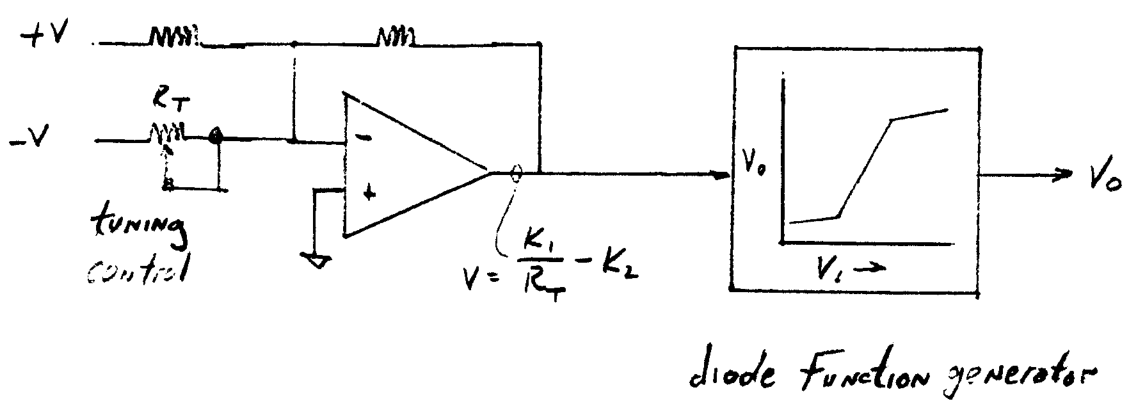
$$V_o = V_i S_1 + K_1, \quad 0 < V_i < V_{b1}$$
$$V_o = V_i S_2 + K_2, \quad V_{b1} < V_i < V_{b2}$$
$$V_o = V_i S_3 + K_3, \quad V_{b2} < V_i$$

1. Straight-line approximation



2. power-series approximation

$$V_0 = K_1 + K_2 V_i + K_3 V_i^2 + K_4 V_i^3 + K_5 V_i^4 + \cdots$$



3. Truncated series with end point correction

The series is in the form $K_1 + \frac{K_2}{R_T} + \frac{K_3}{R_T^2} + \frac{K_4}{R_T^3} + \cdots$

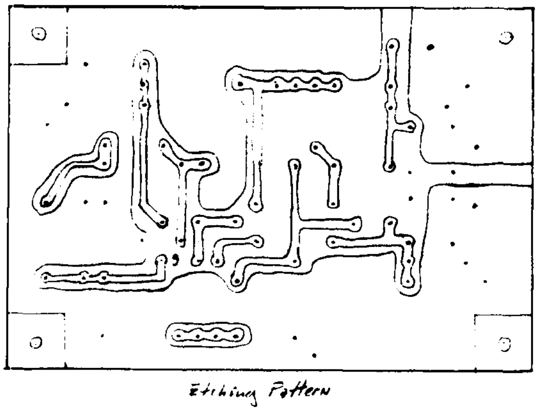


Fig. 6A - Oscillator/buffer board, foil side.

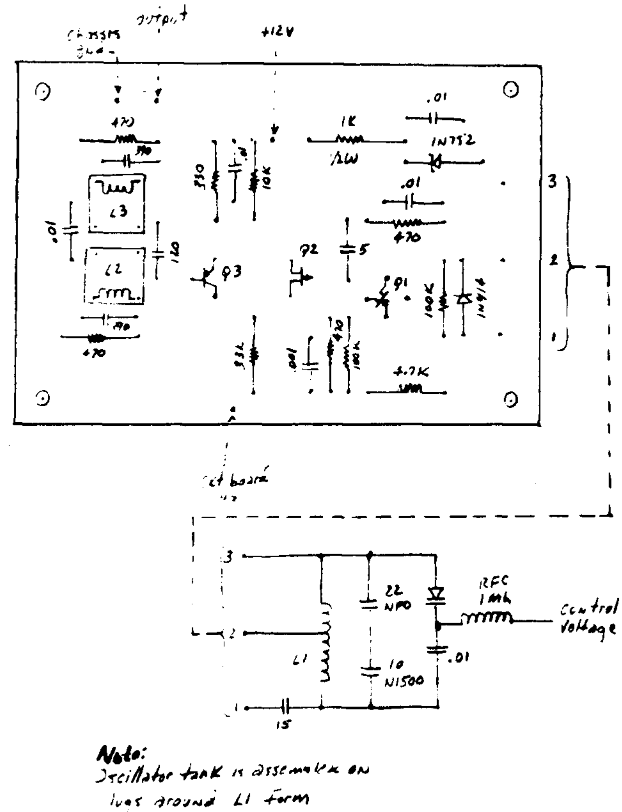


Fig. 6B - Oscillator/buffer board, component side.

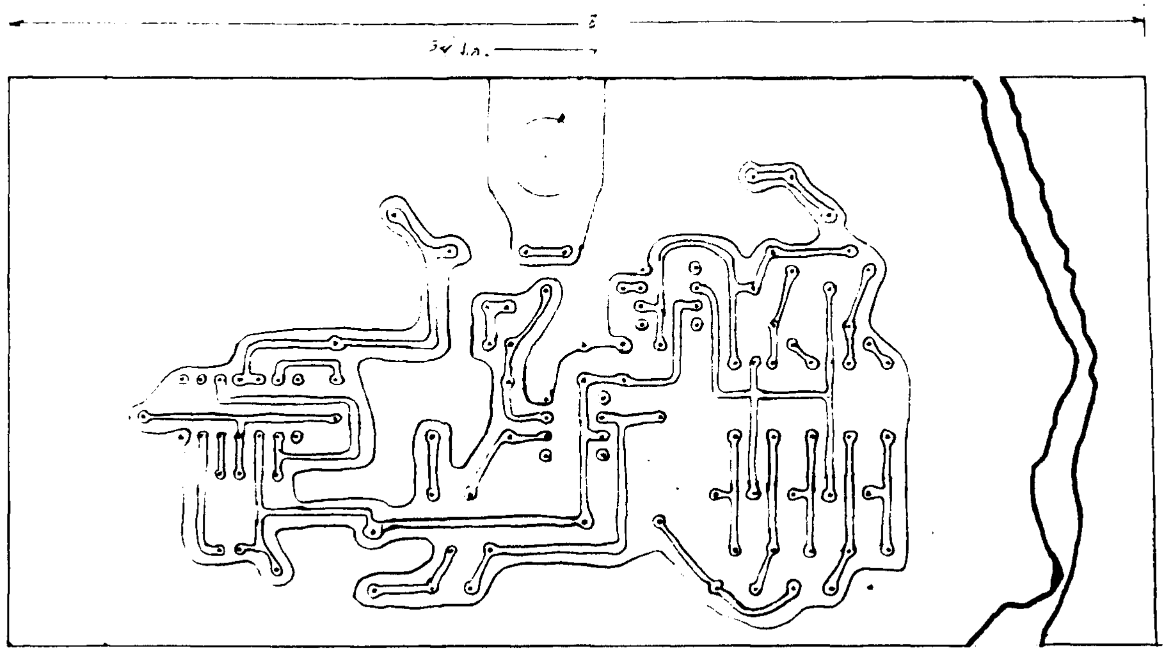


Fig. 7A - Linearization board, foil side.

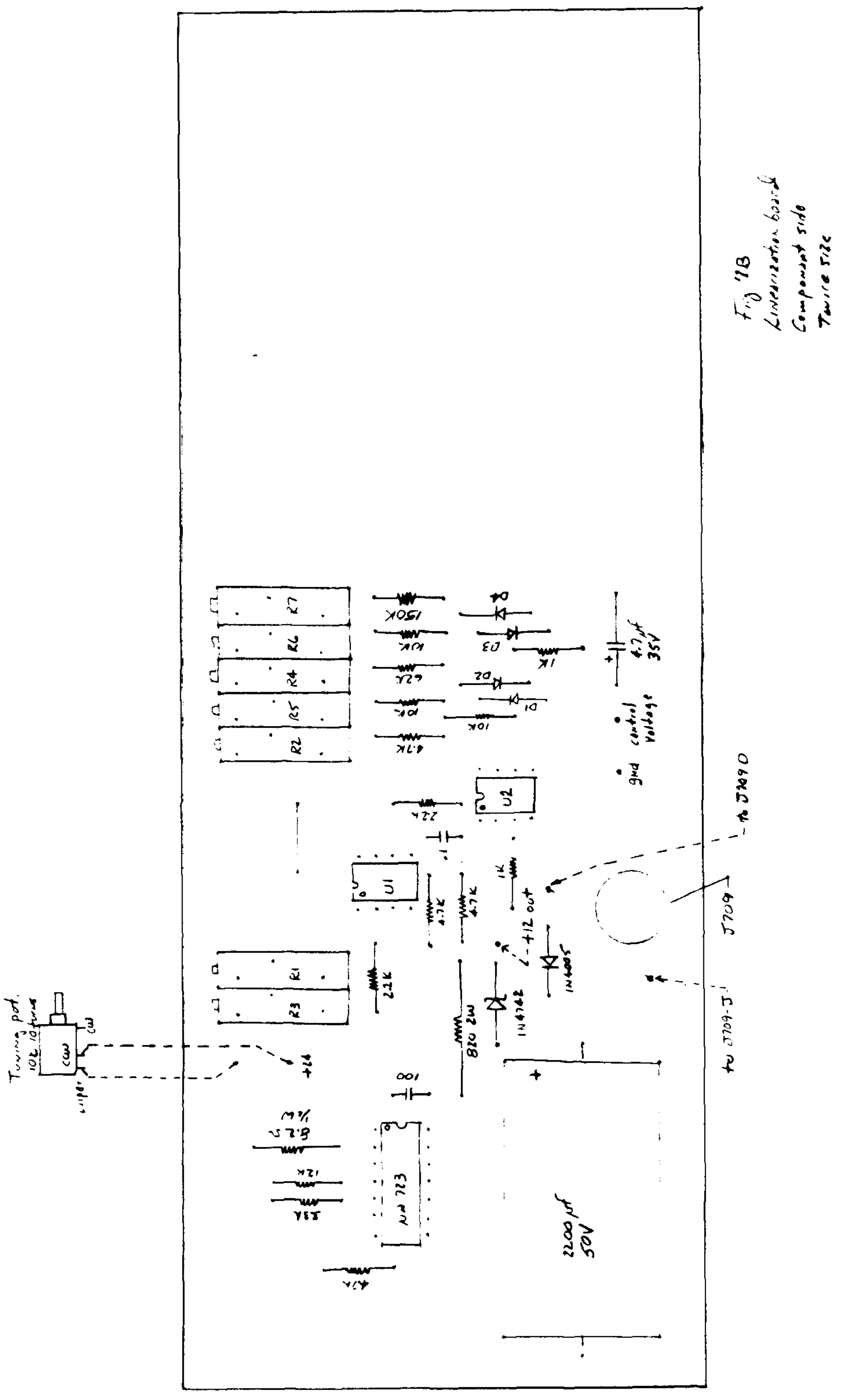


Fig. 7B - Linearization board,
component side.

APPENDIX A

Linearization Network Design Procedure

The capacitance of a Varicap diode is given by:

$$(1)\quad C_d = \frac{C_0}{\left(1+\frac{V}{0.6}\right)^{.44}}$$

Where:
V = diode reverse voltage
C_0 = capacitance for V = 0. C_0 = 136.8 pf. for the 1N5473

Solving equation 1 for V at a given diode capacitance C_d,

$$(2)\quad V = 0.6\left(\log^{-1}\left(\frac{\log\frac{C_0}{C_d}}{0.44}\right) - 1.\right)$$

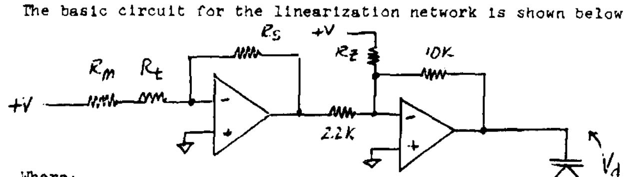
For a given frequency and coil inductance,

$$(3)\quad C_d = \frac{25330}{F^2 L} - C_s$$

Where:
F = frequency in mHz
L = inductance in microhenries
C_s = shunt capacitance in picofarads

Equations 2 and 3 permit calculation of a theoretical voltage vs. frequency curve for any value of L.

The basic circuit for the linearization network is shown below:



Where:
R_t = value of tuning pot resistance
R_m = mid-point adjustment
R_s = high-end point adjustment
R_z = low-end point adjustment

The tuning diode reverse voltage generated by this circuit is:

$$(4)\quad V_d = \frac{10(V)(R_s)}{2.2(R_t+R_m)} - \frac{10V}{R_z} + V$$

Where V = supply voltage

Equation 4 assumes a symmetrical supply of ±V for the operational amplifiers, and that the DC return for the tuning diode is referred to -V.

By selecting three points on the calculated voltage vs. frequency curve for the oscillator and substituting the corresponding diode voltages and tuning pot resistances into equation 4, values of R_m, R_s and R_z may be determined.

Example:

For the PTO substitute as constructed,

L = 49.3 µh, C_s = 15.6 pf, R_t = 0-10K

Using tuning pot settings of 2, 4½ and 7 turns,

TABLE 1

TURNS	R_t	F	C_d	V_d
2	2K	3.255 mHz	32.89 pf	14.712 V
4½	4.5K	3.005 mHz	41.30 pf	8.526 V
7	7K	2.755 mHz	52.09 pf	4.785 V

The supply voltage, V = ±12V

Substituting the values from Table 1 into equation 4,

$$(5)\quad 14.712 = 54.545\frac{R_s}{R_m+2} - \frac{120}{R_z} + 12$$

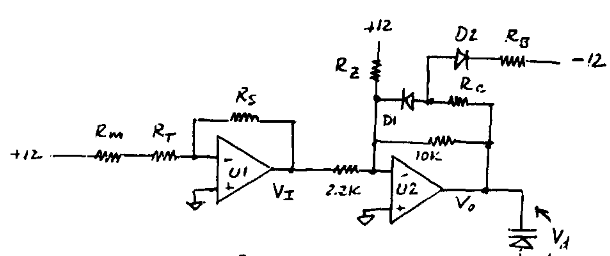
$$(6)\quad 8.526 = 54.545\frac{R_s}{R_m+4.5} - \frac{120}{R_z} + 12$$

$$(7)\quad 4.785 = 54.545\frac{R_s}{R_m+7} - \frac{120}{R_z} + 12$$

Solving this set of equations simultaneously gives:

R_s = 3.527K
R_z = 5.353K
R_m = 5.656K

The Diode Function Generator works by reducing the gain of amplifier U2 when its output voltage passes a pre-set threshold. The circuit below includes only the upper end point correction. Diode D1 switches R_c in parallel with the 10K feedback resistor when the output voltage exceeds the breakpoint voltage. D2 compensates the breakpoint voltage for temperature drift. The breakpoint voltage is determined by divider string formed of R_c and R_b. Diode Function Generators are discussed in detail in Reference 1.



$$(8)\quad V_I = -\frac{12R_s}{R_m+R_T}$$

$$(9)\quad V_o = V_d - 12$$

The gain of U2 above the breakpoint voltage is:

$$(10)\quad A_V = -\frac{10R_c}{2.2(10+R_c)}$$

The upper breakpoint is set at 1½ turns of the tuning pot. Table 2 gives the required function generator input and output voltages at the breakpoint and the end point.

TABLE 2

TURNS	R_t	F_c	F_u	V_i	V_o
0	0	3.455 mHz	3.477 mHz	-7.483 V	10.508 V
1½	1.5K	-	3.309 mHz	-5.914 V	4.467 V

Where:
F_c = required output frequency
F_u = output frequency without compensation
V_o = output voltage required to generate F_c

The required voltage gain of U2 above the breakpoint is:

$$(11)\quad A_v = \frac{\Delta V_o}{\Delta V_i} = \frac{10.508 - 4.467}{-7.483 - (-5.914)} = -3.850$$

$$(12)\quad A_v = -\frac{10R_c}{2.2(10+R_c)} = -3.850$$

R_c = 55.380K

$$(13)\quad V_B = 4.467 = \frac{-V(R_c)}{R_B} = \frac{-(-12)(55.380)}{R_B}$$

R_b = 148.772K

The same procedure is used to determine the values of R_c and R_b for the low-frequency end correction. The low-end breakpoint is set at 7½ turns of the tuning pot. The resulting resistor values are:

R_c = 64.558K
R_b = 99.396K

10-GHz Signal Sources

- Some Preliminary Comparisons

By G. H. Krauss,* WA2GFP

Above 1300 MHz, the most populated amateur band in the U.S. is believed to be 10.0 - 10.5 GHz. Most communications are by wideband fm, using Gunnplexers (R) - a varactor-tuned Gunn diode oscillator with a diode mixer-ferrite circulator receiver. Recently, Mitsubishi has made available both a superheterodyne down-converter and a transmitter, both using oscillators in which a GaAs FET is frequency-stabilized by a dielectric resonator. A dielectric-resonator oscillator (DRO) was compared to the transmitter section of a Gunnplexer, and to a rather rudimentary crystal-controlled source operating at 10,368 MHz (using a pair of cascaded SRD multipliers to generate the ninth harmonic of a driver at the "magic" microwave frequency of 1152 MHz); 10,368 MHz is a fairly well recognized "weak-signal" operating frequency.

*16 Riviera Dr., Latham, NY 12110.

CHARACTERISTIC	CRYSTAL-CONTROLLED MULTIPLIERS	MA87127-1 GUNNPLEXER	MITSUBISHI FO-1010X
Cost	?	∿$130(includes receiver)	∿$35
Power Required (not including tuning voltage, if any)	12V at 122ma. and 28V at 707ma. or 7.26W.	10V at 130 ma. or 1.30W.	6V at 62ma. or 0.372W.
Microwave Output Power	18mW.	13MW.	14.7mW.
at f= (MHz.)	10,368	10,260	10,247
Efficiency (RF/DC)	0.248%	1.00%	3.95%
Tuning Range, Mech.	±1MHz.	10.0-10.5 min	10.2-10.6 min
Elect.	------	56MHz.(+1 to+10V)	------
Stability: 1 min, T=K	0.35KHz. Δf	37KHz Δf	43KHz. Δf
(after 30 min. warmup)	∿0.0dB. ΔP_o	∿0.0dB. ΔP_o	∿0.0dB. ΔP_o
:over ±5 C	±38.4KHz. Δf	∓2630KHz. Δf	∓872KHz. Δf
	<0.1dB ΔP_o	∿0.3dB. ΔP_o	<0.1dB ΔP_o
Sideband Noise (dBc/Hz.) at Specified Δf from Carrier Freq. Δf=20KHz.	-96	-80	-80
50KHz.	-115	-93	-92
100KHz.	-118	-111	-105
200KHz.	-121	-120	-112
300KHz.	-123	-124	-120
500KHz.	-125	-126	-127

(Measurement Equipment Noise Floor at ∿-130dBc/Hz!)

Components

Conducted by Mark Forbes, KC9C*

The theme of this issue's column is speech synthesis. The first talking product introduced, and one almost everyone has at least heard of, is the Speak 'n Spell which used a chip set by Texas Instruments. More recent developments include such things as talking alarm clocks, soft-drink machines and, of course, talking repeaters. Two recently introduces chip sets of fairly reasonable price are featured here.

National Semiconductor DT1050 Digitalker

The Digitalker(tm) was introduced by National in two forms; a complete kit which sells for more than $200, and a 3-chip set which is available for about $85. The DT1050 (which is the 3-chip set) is all that the experimenter really needs; so I'll concentrate on that.

The DT1050 consists of the MM54104 speech processor chip and 2 speech ROMs. The included ROMs give the Digitalker a vocabulary of 137 words, 2 tones and 5 different silence durations. The words are addressed with an 8-bit word address input which is microprocessor compatible, or can be manually addressed with something such as a DIP switch or the like.

The vocabulary included in the ROM chips is quite sufficient for many amateur applications. The vocabulary includes the numbers zero through twenty, thirty, forty, fifty, sixty, seventy, eighty, ninety, hundred, thousand and million. Also programmed are all the letters of the alphabet (great for call signs) and words suitable for uses such as gas pumps, scales, interactive machines and several miscelaneous words such as "on," "off," "the" and even words for the phrase "over and out!"

As you can probably imagine, this circuit is fun to play with. It is a little on the expensive side, but it has so many applications that it is well worth the money - especially the first time you program in your call sign. For information on the Digitalker, contact your National distributor or:

National Semiconductor Corporation
2900 Semiconductor Drive
Santa Clara, CA 95051

*1000 Shenandoah Dr., Lafayette, IN 47905, 317-447-4272 2300-0230 UTC weekdays, until 0230 UTC on weekends.

Texas Instruments TMSK101A and TMSK202

Announced in December were two evaluation kits made by Texas Instruments using the linear predictive coding (LPC) scheme, which is the same used in the Speak 'n Spell. The TMSK101A has a vocabulary of 289 words or phrases. Unlike the digitalker, silence must be programmed in the software controller. The TMSK101A is compatible with control logic or 4-bit microprocessor systems. Included is the TMS5100 speech-synthesis processor and three vocabulary ROMs (VROM).

The TMSK202 is probably the easier of the two to use, as it is designed to interface directly with most 8-bit and 16-bit microprocessor systems. The TMSK202 contains 241 words and phrases and 15 sounds such as explosions and whistles. This synthesizer also has 3 VROMs, one of which has a female voice. The words in both kits can be accessed individually or linked together to form phrases or sentences. In addition to the VROMs, speech information can also be stored in the host microprocessor system.

The linear predictive coding method allows natural sounding speech at relatively low bit rates. For example, the TMS5100 runs at only 100 bits per second. Complete documentation and the kits are available from local TI distributors. The price for the kits is: TMSK101A - $45.00 and TMSK202 - $60.00. Other VROMs are also available for weather and time, military and avionic applications for $11.65 each. All prices are single quantity. More information can be obtained from:

Texas Instruments Incorporated
Central Literature Response Center (SC-346)
P.O. Box 202129
Dallas, TX 75220

Motorola MC12071 High-Speed Prescaler

New from Motorola is a prescaler usable from 90 to 950 MHz. The chip has two dividers, 64 for the vhf range and 256 for the uhf range. In the vhf range (90-275 MHz), the frequencies are translated to 1.4 to 4.3 MHz. The uhf range (90-950 MHz) is translated to 312 kHz to 3.71 MHz. The separate vhf band allows greater sensitivity. A data sheet and more information can be obtained from:

Motorola Semiconductors
P.O. Box 20912
Phoenix, AZ 85036

Data Communications

Conducted by
David W. Borden, K8MMO*

In this issue we continue our study of the Line Interface Program (LIP) which is the workhorse software running in the VADCG Terminal Node Controller (TNC). This program makes the packets working with the 8273 protocol controller chip. It is very complicated software which can be studied only in small modules in order not to become lost in the complex structure. It is clear that we will have to study a little of this software each month for quite a while before completing it. This time we will examine the interrupt structure of the 8Ø85 and the associated interrupt routines running in the TNC.

The 8Ø85 microprocessor has five interrupt inputs which are (here listed in priority order):

1) TRAP - Highest priority interrupt which is useful for power failure or catastrophic events. Currently causes jump to ØCØØH where you should have placed code to recover the state of the TNC when the TRAP occurred. TRAP is a RESTART interrupt (upon occurrence the program counter is saved on the stack and a branch occurs to the desired address) and is non-maskable.

2) RST 7.5 is a RESTART interrupt that is maskable (ignored if the interrupt mask is not set). It is currently coded to a receive interrupt from the 8273. A packet frame has been received by the protocol controller and is ready for processing. The routine jumped to is RXINT.

3) RST 6.5 is a RESTART interrupt that is currently coded to jump to the transmit interrupt (TXINT) routine for the 8273. It is third in priority in the TNC scheme.

4) RST 5.5 is a RESTART interrupt that is coded to jump to the Terminal Interface Program (TIP) jump table and is fourth in the TNC priority scheme.

5) INTR is identical in function to the 8Ø8ØA interrupt, but is not used on the TNC board in serial mode (825Ø USART vice the parallel 8255). It is of the lowest priority, if enabled.

The two new software instructions introduced by Intel for the 8Ø85 were:

1) SIM - Set Interrupt Mask

2) RIM - Read Interrupt Mask

*Route 2, Box 233B, Sterling, VA 2217Ø, 7Ø3-45Ø-5284.

The state of the interrupt masks for the above listed interrupts can be affected only by these two instructions and RESET IN NOT(lower the reset line on the microprocessor). Each of these five interrupts is wired to a specific pin of the 8Ø85 microprocessor and the LIP deals with them.

Forgetting the TRAP interrupt, which is really not implemented in the TNC, the first priority is then to receive packet frames. When the 8273 correctly receives a packet frame, it raises to a logic one pin 11 which is wired to the 8Ø85 RST 7.5 interrupt line, pin 7. This causes the 8Ø85 to save the program counter and execute a jump to the RST 7.5 address, ØØ3CH where a jump to the RXINT routine resides.

The RXINT routine begins by saving PSW/H and then checking the 8273 status register to see if the interrupt result is available. If the interrupt result is available, the packet is complete and we need to go process it. If the result is not available, we need to receive one byte of the packet and bump our Line Buffer pointer (LBIP) after first checking for wrap. If the packet is done, we need to get the results of the frame (the length received) and store the length in the header. This work is done in the RXRESULT routine we will study at a later date.

The transmit interrupt is a signal from the 8273 protocol controller pin 2, wired to the RST 6.5 8Ø85 interrupt pin. Upon receipt of a RST 6.5, we first save PSW/H and then check the 8273 status to see if we are sending a cw i-d. This special American requirement is a special transmit case and must use a special routine located at the top of prom memory (actually at the top of the TIP prom). If it is not a cw i-d at issue at the moment, we check for results available, meaning that we are done with a packet frame and ready to transmit it. If no results are available now, we increment the terminal buffer output pointer, TBOP (discussed last issue), and check for wrap around. If wraparound has not occurred, we update the output pointer in memory, grab the next byte to be sent out the modem and send it out the TXDATA port. If a packet is done and results are available, we have more work to do. We read the 8273 transmit results (TXIR73) and check for early interrupt result. This requires some further explanation.

The HDLC protocol allows for stringing packets together, separated by only one flag byte. Remember each single packet begins and ends with a flag byte. So, without some ability to string packets together, we need a minimum of three flag characters to separate two data frames. We would need the first flag to follow the frame check sequence. The 8273 would require one flag character time as a minimum to start the next frame and the third flag character would actually begin the frame, illustrated as follows:

D D D FCS FLAG1 FLAG2 FLAG3 A C D D D D D D

But, the 8273 is smarter than that. When you begin a frame transmission, one of the things you specify is the number of frame data bytes, so the chip knows when the last character of a frame has arrived. An early end-of-frame interrupt option is provided so that as the frame byte count decrements to 0, the 8273 device generates an interrupt (and of course reports this in the result status byte). With the use of this interrupt then, a single flag character may separate two frames thusly:

D D D D D FCS FLAG A C D D D D D D D D D D

Returning to the code in TXINT, if we read the transmit result (TXIR73) and discover an early interrupt has occurred, we check the last control field send for poll or not I frame. We exit if either condition was noted. Otherwise, we point to start of next entry, check to see if its the last entry to send. If it not, we send this one. Sending is done by calling TXFRAME which will be covered at some future date.

If in our check of the transmit result (TXIR73) earlier, we found that we had a normal completion result and not an early interrupt, then we have to turn off the transmitter. This is the point we have the 8273 lower the request-to-send line which is connected to our transmitter push-to-talk line. We then check the final bit to see if it has been set and if it has, set the line timeout and exit from the interrupt routine. If the final bit is not set, we just exit. Upon exit we restore registers before returning from the interrupt routine.

The last interrupt to study is a line from the 8250 serial chip connected to our serial terminal. If our terminal requires some attention (someone banged on the keyboard for example), the 8250 generates an interrupt, wired to RST 5.5 on the 8085 microprocessor. When this occurs, the LIP vectors us to the Terminal Interface Program (TIP) which handles the 8250. There, we examine the interrupt from the 8250 to see if the keyboard required attention (placing the character hit into the terminal buffer) or the CRT (getting a character from the line buffer and putting it on the tube).

There is of course much more to the LIP program and next time we will continue this study.

QEX: The ARRL Experimenters' Exchange is published by the

American Radio Relay League
225 Main Street
Newington, Connecticut 06111
telephone 203-666-1541

Harry J. Dannals, W2HD
President

Richard L. Baldwin, W1RU
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Paul L. Rinaldo, W4RI
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2) document advanced technical work in the Amateur Radio field, and

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

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