

### Amplifier Keying.

"AMP PTT", on pin 13 of ACC3 at the rear of the transceiver, provides an active-ground signal which can sink up to 250 mA to key an external amplifier. It is the open collector of a grounded-emitter NPN transistor switch which can tolerate up to 32 VDC in the off condition. Most linear amplifiers are designed for this type of keying signal. Connect it to the amplifier using a shielded cable. Connect the shield to transceiver ground at ACC3, pin 1.

### Band Information.

Six band signals are provided on ACC3 for selection of harmonic filters or operating bands in external power amplifiers. These signals are active-ground, and can sink up to 80 mA when on. Only one is active at a time. The signals cover the HF range as follows:

BAND .....	FREQUENCY RANGE (MHz)
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A .....	1.6 - 3.0
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B .....	3.0 - 5.0
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C .....	5.0 - 8.0
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D .....	8.0 - 13.0
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E .....	13.0 - 20.0
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F .....	20.0 - 29.9
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The frequency ranges, which correspond to half-octave harmonic filters in the transceiver, are valid when the transmitter is activated.

When the receiver is manually tuned, a channel or "snapshot key" recall is executed, the band sweep display is initiated, or during scanning operation, the "F" filter is automatically selected to prevent continual switching of filter relays. This eliminates their noise, and prolongs their lives. As soon as the transmitter is again activated, the proper band signal is asserted at least 20 ms prior to the onset of R.F. output. This signal remains active until the receiver frequency is changed.

### Setting Amplifier Drive Levels.

Most amplifiers will require a drive level between 30 and 75 Watts to achieve full output power. If external ALC is not used, the output of the transmitter must be reduced using the power output control to avoid damage to or destruction of the amplifier.

### Applying External ALC and VSWR Cutback.

External ALC voltage may be applied to control the output power of the combined transceiver/amplifier system. The "sense" of this voltage must be positive; i.e., the voltage increases as the square root of the output power. Further, the voltage value applied depends on the drive requirements of the external amplifier. The external detector's output should be adjustable for precise setting of drive power. Transmitter output power is related to ALC voltage by the following equation:

$$V_{alc} \sim \text{SQRT}(P_{output} / 25 \text{ W})$$

E.g., for a required drive level of 50 Watts, drive cutback begins at around 1.4 Volts.

Very little filtering should be used at the external detector's output. Enough to eliminate R.F. is adequate, as further peak detection and loop lag compensation are performed inside the transceiver.

External ALC must be coupled to the transceiver through a diode so that no current is drawn from the transceiver

pin when the external amplifier is off. This prevents the internal ALC from being affected.

An external signal proportional to the square root of the reflected power may also be applied to ACC3. This is normally done to protect the external amplifier against high VSWR loads. The sense of this signal is exactly the same as that of the ALC above for the forward power case. Note that for a load VSWR of 2:1, the reflected power is about 10% of the forward power; the reflected detector voltage will be 33% of the forward detector voltage.

## AN101: 505 DSP - On RTTY

### Interfacing External H.F. Data MODEMs.

External H.F. data MODEMs or "TNC's" may be interfaced to the 505 DSP using the transceiver's ACC1 connector. Provision is made for audio input and output (I/O) at a nominal level of 0 dBm (1 mW), and at an impedance of 600 Ohms. This means the receiver audio appears at pin 16 with an available level of approximately 2 Volts peak-to-peak, or 0.707 Volts RMS. This level is independent of the volume control setting of the transceiver. The external device must provide the equivalent level into pin 17 to drive the transmitter to full output power.

If selectable, the external MODEM should be set to its "high-level" output mode. If the MODEM isn't capable of providing audio drive over 1 Volt peak-to-peak, it should be set to around 10 mV; the output should then be connected directly to the microphone audio input on the control head. Refer to the 505 DSP Installation Manual for pin assignments.

### Transmitter Keying and Push-to-talk (PTT) Requirements.

The 505 DSP PTT line assumes a voltage near the main supply voltage when not active. To activate the transmitter, this line must be grounded. Approximately 12 mA of current must be "sunk" by the external device to bring on the transmitter.

### Transmit-Receive Switching Delays.

A delay between the activation of transceiver PTT and the onset of transmitter R.F. energy is necessary to prevent "hot-switching" of relays both internally and in any external amplifier. This delay is slightly more than 20 ms, so the external MODEM must be set to accommodate it.

A delay between the de-assertion of PTT and the onset of receiver audio occurs because the transceiver must reset itself for receiver operation by toggling relays, re-programming filters, and so forth. This delay is slightly less than 20 ms.

Note that the total "turn-around" time between the transmission of data and the reception of any acknowledgement includes not only the transceiver delays, but also the propagation time of the actual radio signals in both directions. I.e., the time required for a radio signal to travel half way round the world must be doubled and added to the total time. The great-circle transit time around the world is approximately 70 ms.

### RTTY Filters and Frequency Offsets.

Two special filters are provided in the 505 DSP receiver to accommodate audio frequency-shift keying (AFSK) RTTY operation. The "Data Med" filter has a passband centered on 1700 Hz, and the "Data Hi" filter is centered at 2210 Hz. The bandwidth of each of these filters is 500 Hz. The transmitter always has a passband from 180 Hz to around 3 kHz. The use of tone pairs above 1600 Hz is mandatory to avoid transmitting audio second harmonics appearing the transmit passband.

To calculate the actual transmitted frequencies, the audio mark and space tone frequencies must be known. Most amateur H.F. MODEMs use the high tones, at mark frequency 2,295 Hz, and space frequency 2,125 Hz. With this arrangement, amateurs must use LSB mode to transmit and receive signals that are "right-way up". It is possible to command the MODEM to reverse its tones, then to use USB mode and still be right-way up.

To compute the transmitted mark frequency using LSB mode, subtract the audio mark frequency from the transceiver's dial frequency. In USB mode, add the two frequencies together. Inversely, to find the dial frequency from a given mark frequency, add the AFSK mark frequency to the R.F. mark frequency using LSB. Subtract them for USB operation.

## CW Features.

Many features of the 505 DSP are designed especially for the CW operator. Narrow IF-DSP bandpass filters are available for the utmost in receiver selectivity on crowded portions of the spectrum. Offset frequency can be selected to suit the tastes of the operator, and a CW spotting tone is provided for "zero-beating" received signals.

In transmit, the keyer has been designed with flexibility in mind. Keyer speed and weighting are continuously adjustable, and the rise and fall times of the output waveform can be altered using a dynamics control. Full break-in, or "QSK", operation is possible beyond 30 word per minute (WPM). Alternatively, semi-break-in operation may be selected with a continuously adjustable hang time. Details of these features are discussed below.

### CW Spotting Tone.

In CW mode, the spotting tone may be activated by depressing the "space bar" on the PC keyboard. A tone will be audible at the level set by the "sidetone" level control, and at the frequency selected by the "CW offset" control. The tone represents the transceiver's actual transmit frequency, and is extremely useful in setting the frequency equal to that of a received signal. After some practice, it's easy to tune the transceiver to within several Hz of the desired signal.

Many operators we've spoken with find it's very handy to first select the 1 kHz bandpass filter (BPF), find a signal of interest, and use the spotting tone to zero-beat it. Then, a narrower filter may be selected to eliminate QRM. This procedure is quite fast, once one gets used to the controls.

### Selecting an Offset Frequency.

While the "dial" frequency of the transceiver always indicates the actual transmitted or received frequency, the CW offset changes the audio tone coming out of the loudspeaker or headphones. The tone frequency may be altered from 300-800 Hz in 100 Hz steps to suit the liking of individual operators. The received signal will appear at the same frequency as the transmitter's sidetone.

### Setting the Keyer Parameters.

When used with a "paddle", the keyer generates self-completing "dits" and "dahs" according to closures of the two switch contacts. The lengths of these elements is inversely proportional to the keyer's speed setting. Speed is calibrated in WPM, based on the average length of a word in ordinary language, and the frequency of appearance of particular letters in words.

With normal weighting, the dit length is defined to be exactly one third of the dah length; i.e., their length ratio is 1:3. The spacing between the elements is one dit length. As the weighting control is adjusted toward the "heavy" end, the ratio increases toward 1:1. To avoid altering the speed as this is done, time is borrowed from the space in between the elements. The effect is that the elements tend to sound closer together. Setting the weighting control toward the "light" end effectively shortens the dits in relation to the dahs. This results in a more "staccato" sound.

The keyer's dynamics control can be used to alter the rise and fall times of the transmitted waveform. The longest rise and fall times are obtained at the "soft" end of the range. At the other end, the times are short enough that the amount of ALC comes into play. If the rise time is faster than the ALC attack time, some "overshoot" will occur. When operating with "harder" settings of the dynamics control, it is best to reduce the TX Gain control while monitoring ALC on the TX meter; reduce the gain until 2 to 4 dB of ALC is indicated. This will minimize the overshoot.

Note that the CW sidetone reflects the setting of the dynamics and other controls. The change in rise and fall times will be audible in the sidetone.

### Auto-repeating and Memory Features.

If the dit key is held on continuously, the auto-repeating feature of the keyer causes a continuous string of dits to be transmitted. This means a minimum of keystrokes is necessary during sending. E.g., to send the numeral "4", only two key depressions are required.

The keyer also incorporates "2-key roll-over" memory. This means that while the dit key is being held down, if the

dah key is then pressed, the keyer will finish sending the dit, then will send the dah with the correct spacing. The memory feature makes sending much easier and reduces errors.

Left-handed operators can reverse the dit and dah keys without re-wiring their paddles by simply selecting the "left-hand" mode from the CW Keyer menu in the control software. A "straight key" mode can also be selected.

#### Straight Key Mode.

Separate jacks are provided on the control head for the paddle and a straight key. In straight key mode, the dit key of the paddle will remain active.

All keyer controls are also active in the straight key mode. The self-completing feature of the keyer is useful in smoothing out sending, but means that the speed setting must be adjusted slightly higher than otherwise. A value approximately 1.5 times the actual sending speed is recommended. Increase the speed if you notice the keyer is missing some of your key depressions.

This feature can effectively be disabled by setting the speed to its maximum setting. The keyer will then behave as if the self-completing feature were not present. This setting is recommended when using the transceiver with an external keyer, such as those found in external HF MODEMs.

#### QSK Operation.

When operating with QSK on, the transmitter stays on just long enough to complete the sending of the element corresponding to the last key depression. If no further keystrokes are detected, the transceiver will immediately switch to receive mode. The memory and auto-repeating features affect QSK operation such that, during normal sending, the transmitter remains on during entire words. This is necessary to maintain the correct element spacing.

When QSK is off, the transmitter will "hang" on for a time after the last key depression. The hang time is continuously adjustable using the "VOX delay" control.

#### CW TX Buffer.

Nine 60-character buffers are provided for quick, error-free sending of a common message, such as a call sign or a "CQ". The message is typed ahead of time, and can be sent using a "smart key" combination. The message will be sent using the current keyer settings.

#### CW Type-ahead Buffer.

In version 2.xy and higher of the control software, a multi-line type-ahead buffer is provided for keyboard sending of CW. The buffer window may be activated and left on-screen next to the control window.

## 505 DSP on Phone.

The 505 DSP's advanced IF-DSP processing is used in transmit mode as well as in receive. This makes it possible to alter the transmitter's frequency response by changing the characteristics of the digital, bandwidth-limiting filters. It also means that speech processing equivalent to RF compression can be employed, as discussed in AN104.

The nominal frequency response of the transmitter in SSB modes is approximately 180-3,000 Hz at the 6 dB points. The response between these points is symmetrical, and between 400 and 2,800 Hz, and flat to within  $\pm 0.75$  dB. See Figure 1. The flat audio response and the low-distortion phasing-method modulator mean that the actual transmitted audio quality is dependent mostly on the quality of microphone used.

The Kachina hand microphone is designed for the highest voice intelligibility, and so has a response which matches the "psophometric" response of the human ear. The response has a midrange peak corresponding to the peak sensitivity of the human auditory system. This characteristic does not necessarily produce the most natural sound with all voices.

Other microphones have similar responses, yet others are flatter. We find that transmitter audio quality can be highly variable, and depends on the nature of a person's voice, the dynamics of speech, the proximity to the microphone, and propagation conditions; in short, it is a complex function of many factors which are difficult to measure and control.

## The Transmit Audio Equalizer.

To allow the operator to alter his or her transmitted audio attributes, the 505 DSP has a transmit audio equalizer. This control allows up to 12 dB of bass or treble boost, just as is found in the "tone" controls of other audio devices. The equalizer can be quite effective in compensating for different voice traits under varying conditions. See Figure 2.

As an example, in a DX "pile-up", more treble is effective in producing a "piercing" signal. Alternatively, a setting on the "bass" side can be more pleasing for "rag-chewing".

## The Speech Monitor.

The 505 DSP provides a speech monitor to allow operators to get a better idea of the sound of their transmitted signal. The monitor output is a replica of the actual filtered, processed, and equalized audio applied to the modulator. It's useful in setting the relative levels of bass, treble, and speech processing applied.

Microphone audio in the 505 DSP is amplified and then digitized using a 10-bit analog-to-digital converter (ADC). This signal is applied to a 16-bit IF-DSP modulator. The dynamic range of the system exceeds 60 dB with respect to the audio input. Because the speech monitor output is generated using 8-bit pulse width modulation (PWM) techniques, its dynamic range lags the actual transmitted signal by at least 12 dB. This means that while the tonal quality of the speech monitor accurately reflects that of the modulator's output, quantization noise and other inaccuracies limit the ultimate sidetone quality.

Also note that the signal-to-noise ratio, distortion, and frequency response of a signal depend heavily on the characteristics of the receiver used to evaluate it. E.g., most receivers don't have a frequency response down to 180 Hz unless the IF shift is used. And quite often, the intermodulation distortion (IMD) introduced in the receiver is actually greater than that produced by the transmitter on the other end.

## Transmitter Gain Control.

The microphone gain setting in the 505 DSP isn't as critical to audio quality as in other transceivers. The digital ALC compensates effectively for audio inputs which would otherwise drive the transmitter above the set power limit. Little difference will be noticed in quality as the "mic gain" is adjusted above the point where ALC action occurs.

The transmitter also employs a slow-acting transmitter gain control (TGC). This control, transparent to the operator, reduces ALC "overshoot" by keeping the total transmitter gain within 6 dB of that necessary to achieve full power output. It means that, in the long run, the excess transmitter gain cannot be more than twice what's required for a given power output setting.

Transmitter gain is precisely set using the "mic gain" control. Excess gain is registered on the ALC meter in dB. A mic gain setting of 50% provides exactly 6 dB less gain than at full. As the power output setting is reduced, the transmit gain is automatically reduced in proportion; the IF-DSP system regulates its output precisely to keep transmit gain within reason, according to power output.

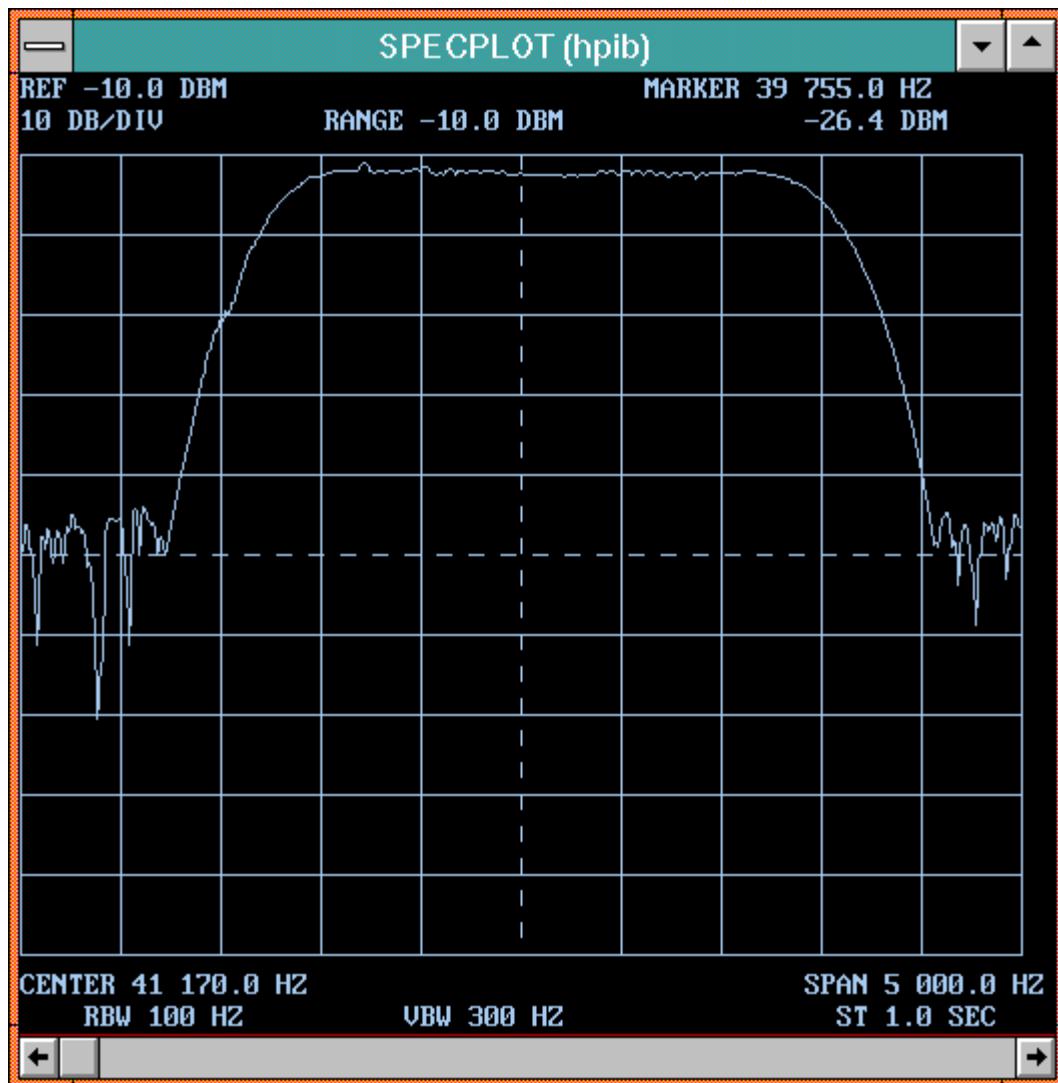


Figure 1. Flat Equalization

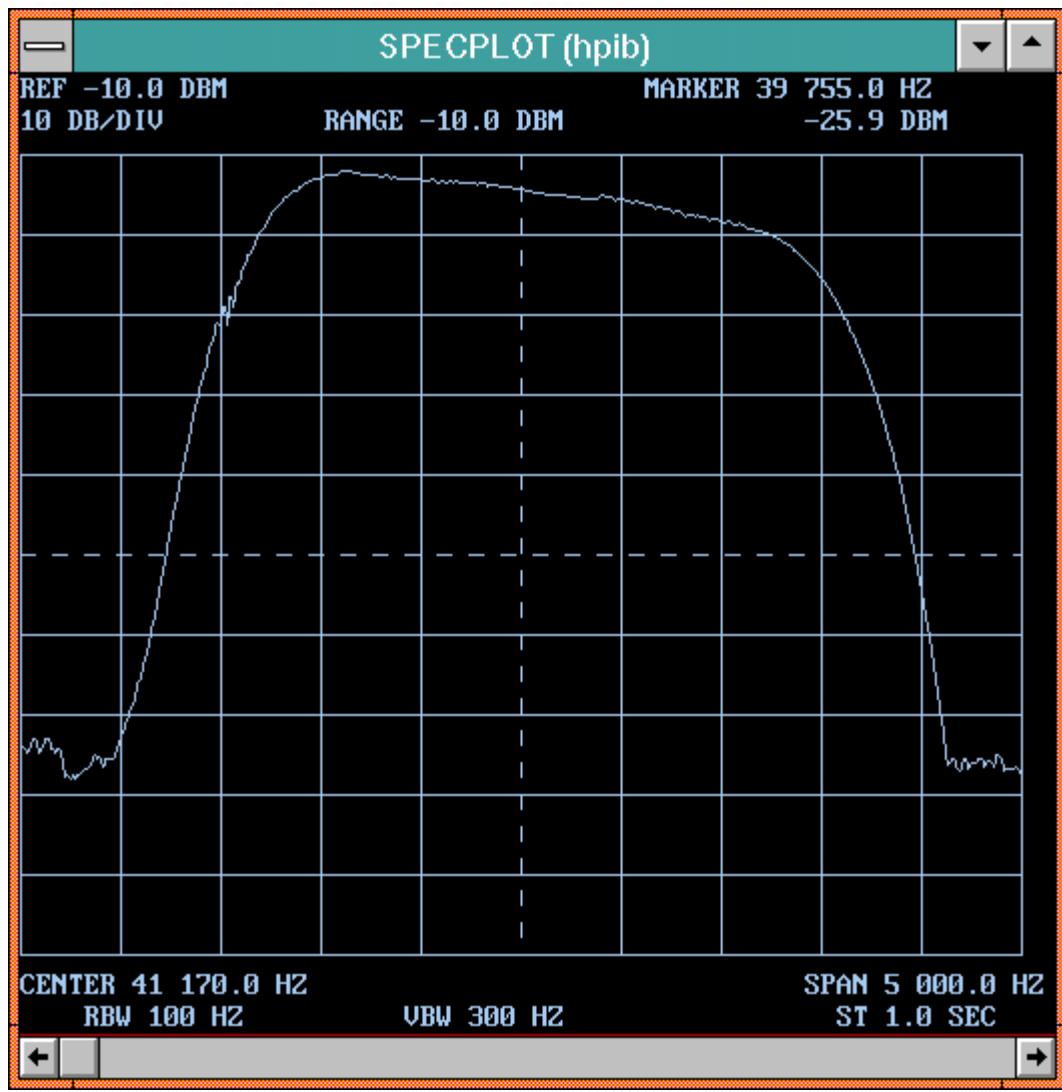


Figure 2a. 12 dB Base Equalization

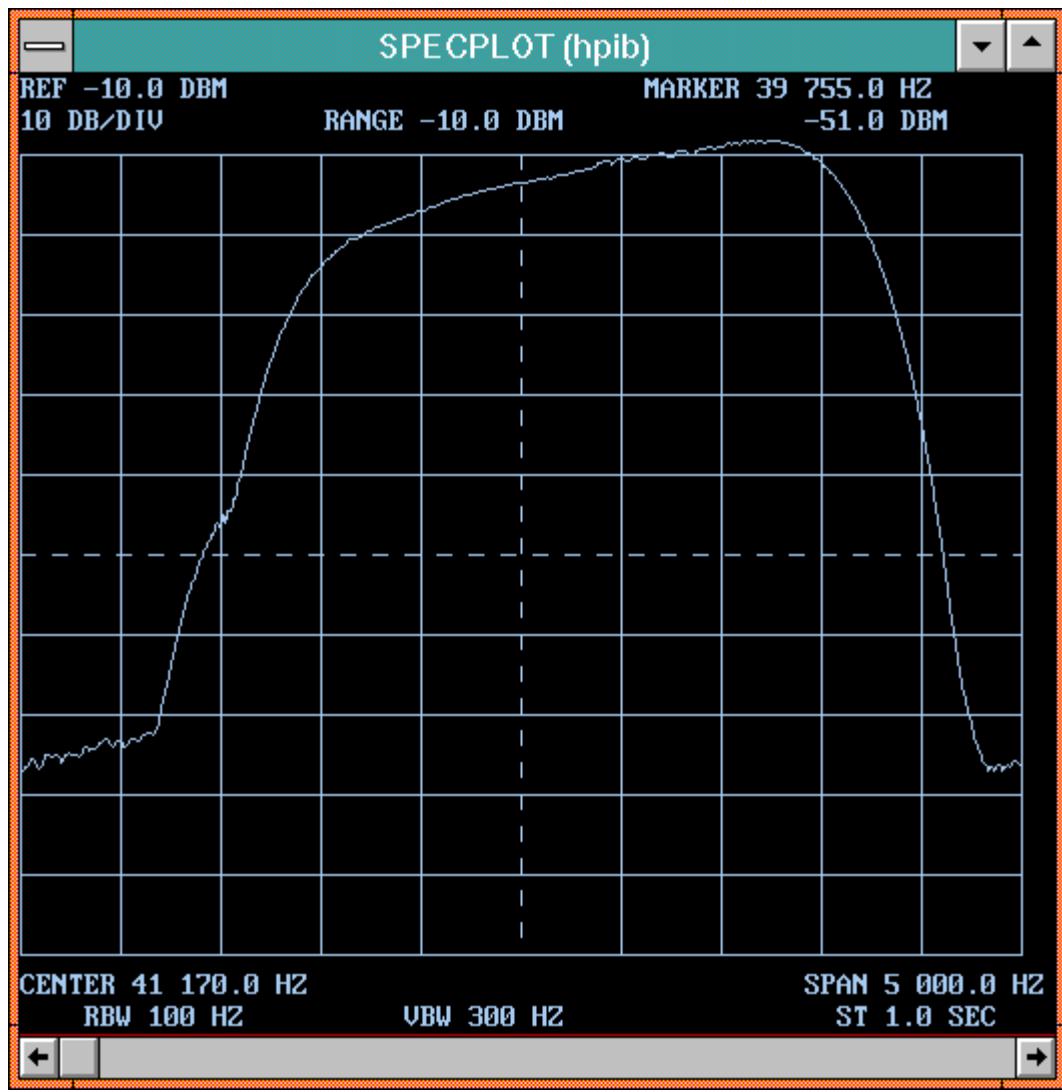


Fig 12b. 12 dB Treble Equalization

## AN104: 505DSP - Speech Processing using the R.F. Compressor

### R.F. Speech Processing: More Bang per Buck.

A distinct advantage in a digital, phasing-method SSB modulator is that the R.F. envelope can be calculated before the modulation is performed! This allows us to employ R.F. compression methods on the baseband signal prior to filtering, where they can be effective without adding to "splatter".

Refer to Figure 1. In the digital R.F. compressor, the transmitter audio is processed to produce in-phase and quadrature signals I and Q. All signals in Q are phase-shifted 90 degrees from the signals in I. The envelope of the SSB output is then computed as:

$$E_{ssb} = \text{SQRT} (I^2 + Q^2)$$

To avoid the time-consuming square root calculation, we can use an approximation:

$$\text{For: } |I| > |Q|, \text{SQRT}(I^2 + Q^2) \sim |I| + 0.4|Q|$$

$$\text{For: } |Q| \geq |I|, \text{SQRT}(I^2 + Q^2) \sim |Q| + 0.4|I|$$

This envelope amplitude is used to compress the baseband levels so that the peak-to-average ratio of the transmitted signal is reduced; i.e., the average power is increased. The effect is the same as that produced by R.F. processing. This naturally involves the introduction of distortion, since the transmitter will no longer be linear. But this type of distortion enhances the syllabic and formant energy in speech without introducing the "mushy" sound caused by audio clipping.

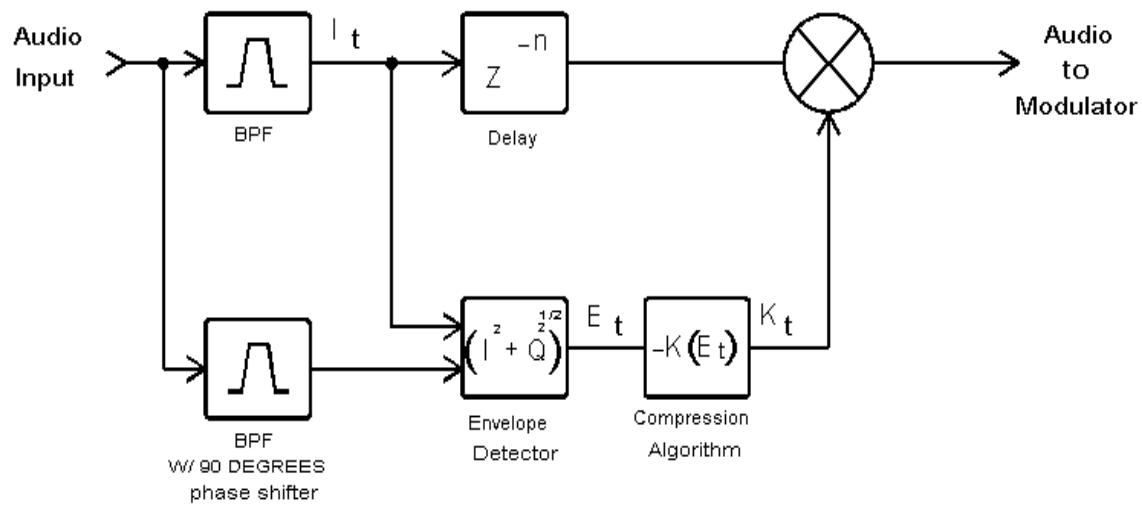
To elaborate, consider that the human voice has a peak-to-average ratio as high as 15 dB; this doesn't utilize a peak-limited transmitter very well, and at the 100-Watt PEP level, the average output power might be as little as 3 Watts! R.F. compression enhances the weaker parts of human speech such that intelligibility is improved. As shown by studies in the literature, 15 dB of R.F. compression can produce up to 6 dB of intelligibility improvement on the receiving end. This is equivalent to a quadrupling of output power!

The compressor attack and decay times can be varied to change the amount of processing introduced. As they are made faster, the compression approaches the effects of R.F. clipping. It's widely known that this is the most effective form of speech processing.

### Acknowledgement.

Some of this material appears in "Smith, D. T., Signals, Samples, and Stuff", a series of articles published in 1998 issues of QEX, the ARRL's Forum For Communications Experimenters. For subscription information, contact: ARRL, 225 Main Street, Newington, CT, 06111-1494; or via E-mail, use "qex@arrl.org"

Figure 1. Digital R.F. Compressor



DIGITAL R.F. COMPRESSOR

### Automatic Antenna Tuner Unit (ATU) Functions.

Most transmitters are designed to operate into a load which looks electrically like a 50 Ohm resistor. When the antenna isn't exactly 50 Ohms, some of the power is reflected back to the transmitter, where it may produce deleterious effects. The reflected energy may increase the heating in the output devices to dangerous levels, or it may cause voltages there to exceed breakdown limits. Also, operating an amplifier into a load very different from the design value may destroy its linearity because of alteration of the output load line. Finally, some of the transmitter's output power will obviously be lost because of the mismatch.

The goal of the ATU is to provide the transmitter with a constant 50 Ohm load. It does this by inserting a network of reactance values which transforms the antenna impedance. We shall see below what the typical network is, and why the technique works. First, let's examine how the antenna impedance is measured, and define a way of plotting it on a graph.

#### Antenna Impedance Measurement.

A directional coupler is a form of impedance bridge which produces outputs proportional to the forward and reflected power. If we arrange to detect the voltages at the forward and reflected ports, the reflection coefficient is defined as:

$$\rho = V_{\text{refl}} / V_{\text{fwd}}$$

and the voltage standing wave ratio (VSWR) is computed as:

$$\text{VSWR} = (1 + \rho) / (1 - \rho)$$

To find the antenna impedance, we also need to know the phase of the reflection coefficient. This is just the relative phase of the signals at the forward and reflected ports. Several types of phase detectors are suitable. We represent the reflection coefficient's phase using Greek letter phi, or  $\phi$ .

#### Complex Impedances and the Smith Chart.

The Smith chart, invented by Philip H. Smith more than half a century ago, is simply a graph for plotting complex impedances. The chart consists of a circle, the center point of which represents the characteristic resistance of interest, usually 50 Ohms. A zero-reactance line is drawn horizontally through the center. All points on this line are pure resistances. The point at the right end represents infinity, or an open circuit; the point at the left end represents zero, or a short circuit.

Other circles or arcs can be drawn on the chart to describe all points having either the same resistance or the same admittance. Constant-resistance circles always pass through the infinity point, and have diameters smaller than the chart diameter. Constant-admittance arcs pass through the zero point, and also have diameters smaller than the chart diameter. The beauty of the Smith chart is that as circuit elements are adjusted, the plotted impedance follows a predictable circular curve. This property is extremely useful in planning antenna tuner strategies.

Our antenna impedance is represented in polar form by the parameters we measured above:

$$Z = (\rho, \phi)$$

The point can be plotted on the chart by interpreting the magnitude of the reflection coefficient as the distance or radius from the center. The phase is the angle between a line from the center to the point and the zero-reactance line.

The point can also be expressed in Cartesian or rectilinear coordinates using the transforms:

$$x = \rho \cos \phi$$

$$y = \rho \sin \phi$$

The origin (0,0) is the center of the chart. The normalized complex impedance  $R+jX$  can then be computed using:

$$R = (1 - x^2 - y^2) / ((1 - x)^2 + y^2)$$

$$X = 2y / ((1 - x)^2 + y^2)$$

For a 50-Ohm system, we'd just multiply R and X by 50.

Any impedance can be expressed in the form  $R+jX$ , where R is the resistance and X is the reactance, both in Ohms. In the equivalent circuit, the resistance and reactance are in series. At a single frequency, even the most elaborate network looks electrically like such a circuit. The reactance of an inductor is positive; a capacitor has negative reactance.

The reactance of a capacitance C, in Farads, is:

$$X_C = 1 / 2\pi f C$$

and that of an inductance L, in Henries, is:

$$X_L = 2\pi f L$$

where f is the frequency in Hertz, and  $\pi$  is the constant pi, 3.14159.

Getting Around the Chart.

From any starting impedance point, the addition of series reactance transforms the impedance along a constant-resistance circle. A series inductance takes us clockwise along such a circle. A shunt capacitance takes us counter-clockwise along a constant-admittance circle. These two element types can be used to build an ATU which provides a conjugate match between any impedance and the characteristic resistance. Two examples will illustrate the concept.

In the first example, the antenna data at 2 MHz are:

$$Z = 100 + j50$$

$$\rho = 0.45$$

$$\phi = 26.6^\circ$$

$$\text{VSWR} = 2.62 : 1$$

The first step is the addition of a shunt capacitor at the antenna, transforming us to the 50 Ohm constant resistance circle, and an impedance of 50-j61. The capacitance is 1100 pF.

The second step adds a series inductance of 4.9  $\mu$ H, bringing us to 50+j0.

In the second example, also at 2 MHz, the start point resistance is less than 50 Ohms. The antenna data are:

$$Z = 20 - j20$$

$$\rho = 0.50$$

$$\phi = -130^\circ$$

$$\text{VSWR} = 3.00$$

The first step adds series inductance of 3.5  $\mu$ H, getting us to 20+j25 on the constant admittance circle passing through 50 Ohms. The second step adds shunt capacitance of 1950 pF, bringing us to 50+j0.

Now that we know how to get around the Smith chart, we need an algorithm, under microprocessor control, which steers the network toward a match.

## "Fuzzy-reasoning" ATU Algorithms.

Fuzzy reasoning is a process, like those in the human mind, which assesses a situation in relative terms. E.g., if we see the antenna is capacitive, we know inductance must be added; if inductive, capacitance must be used. Further, if the antenna is very capacitive, more inductance must be inserted. A fuzzy-reasoning system employs transfer functions which describe how much adjustment to make based on detector inputs. The transfer functions can represent not only the theoretical requirements of the system, but can also incorporate any predictable errors from the detectors and other sources. Fuzzy reasoning tends to overcome errors in systems which can provide only roughly-accurate absolute measurements, but which produce good relative resolution.

In the case of the ATU, the accuracy of the phase detector is likely to degrade rapidly below VSWR=1.3:1, and we must rely solely on the reflection coefficient to guide us. A certain amount of "thrashing about" must be employed to find the minimum VSWR. Above this level, the phase information is useful in steering toward the goal. Transfer functions are developed which embody the matching rules, and a fuzzy-reasoning engine is created which adjusts circuit elements on a step-by-step basis until minimum VSWR is reached.

Step size must be determined by the degree of correlation between the transfer functions and the actual performance of the circuit. Tuning speed is the parameter which suffers because of inaccuracies. In actual practice, with measure-and-adjust cycles of around 25 ms, tuning times well under one second are obtained. The use of adaptive, memory-tuning techniques enhances performance. Refer to Kachina AN106 for further information.

## General Characteristics.

The 505 DSP internal automatic antenna tuner unit (ATU) provides a conjugate match between the transmitter and the antenna. It automatically inserts reactive elements so that the transmitter sees a load very close to 50 Ohms. When the ATU is activated, the transmitter comes on and CW power at a constant 20 Watts is routed to the selected transmit antenna port. Detectors provide information about antenna reflection coefficient magnitude and phase, which the "fuzzy reasoning" algorithm uses to steer the ATU network to a 50-Ohm match.

If the detectors indicate the antenna voltage standing wave ratio (VSWR) is greater than 3:1, the transmitter immediately shuts off, and the ATU is placed in the "straight through", or bypassed position. This action protects the ATU against high voltages or currents which might otherwise damage it. An ATU capable of matching load VSWR's much higher than 3:1 at 100 Watts would have to be too physically large and costly for inclusion inside any HF transceiver.

## ATU Memory Operation.

The frequency, and either the antenna impedance or ATU network data are stored in one of 64 non-volatile memories for later use. For initial antenna VSWR less than 3:1, the ATU will always reduce the VSWR to 1.5:1 or less; 1.1:1 is typical, since the algorithm always searches for the lowest VSWR possible. In this case, ATU network data are stored along with the frequency. When this frequency, or one near it, is selected in future, the correct network is inserted. Further tune cycles are eliminated.

For initial antenna VSWR greater than 3:1, the ATU won't attempt to find a match, and the raw detector information is stored in one of the memories. These data, along with any network data stored as above, are used to plot the antenna impedance on the transceiver's Smith chart display. Refer to AN105 for a discussion of the Smith chart.

## Frequency Resolution and Memory Allocation.

Frequency resolution for ATU memory points is approximately 30 kHz. Only one data point per antenna port may exist in memory for each frequency. When a new tune cycle is initiated, memory is searched for the stored point nearest the current frequency. If they are network data, and the frequency is close enough, and the stored network corresponds to the correct antenna port, it is inserted as a starting point for the new cycle. This reduces cycle time.

If the new frequency, when truncated to the 30 kHz resolution, is identical to that of the previously stored data for the appropriate antenna port, the old data are overwritten in memory with the new data. If no stored data exist within 30 kHz of the new frequency, the next available storage location is used to save the new data.

When shipped from the factory, all memories are cleared. Storage locations are henceforth allocated by frequency, by antenna port, and in sequential order. I.e., for each truncated frequency at which the ATU is activated on each antenna port, an empty storage location is used, until the memory is full. E.g., if antenna port B is never used, 64 locations are available for port A data storage. This is equivalent to:

$$F \text{ span} = 64 ( 30 \text{ kHz} ) = 1920 \text{ kHz}$$

of spectrum coverage. If port A is used for the first 30 storage points, then port B is selected and 14 points are stored, 44 locations are occupied. 20 remain for further use.

After all 64 locations are occupied, new points which do not match previously stored frequencies and ports will overwrite the oldest stored data, regardless of port or frequency. The method boils down to allocation on a "first-come, first-served" basis. Most hams will operate in a total spectrum less than half that of the memory capacity.

## ATU Memory Recall and Clearing.

When the transmitter is activated on or near a frequency and antenna port for which a memory is stored, the correct network is automatically inserted when the ATU is "ON". If the antenna system hasn't changed since memory point storage, the ATU doesn't need to go through another tune cycle for that frequency and port. When the antenna system has changed only slightly, the ATU can always be re-activated to achieve the lowest VSWR. The new data will automatically be stored in the proper location.

Whenever a new antenna is connected to one of the antenna ports, the memory data should be cleared for that port. This allows more tune cycles to be achieved on the new antenna before any old data associated with the other port are overwritten.

## AN107: 505DSP - On Eliminating Electromagnetic Interference

The 505DSP radio is well shielded and in our experience, noise problems are usually relatively minor and easily correctable. But as with any other radio operated near a computer, the 505DSP may pick up computer or monitor noise via the antenna. To see whether the noise is entering the radio through the antenna, disconnect the antenna from the rear panel of the radio. Experience has shown that most radiated noise is emitted by the monitor. To see whether the monitor is the cause, disconnect the monitor power cable from its source of voltage (it is not sufficient to just turn the on/off switch off), and unplug the monitor data cable from the computer.

These checks should be made at different bands and different frequencies, and with one or the other, or both cables disconnected. Often the power cable may cause one noise, the data cable another. Most monitors come with ferrite beads on the data cable where the cable exits the monitor. However, a considerable length of cable may remain between the bead and the printed circuit board assembly. In such cases, it may be necessary to take the monitor from its case and fit a clip-on ferrite bead over the cable where it enters the pc board.

If the power cable is the culprit and a clip-on ferrite bead does not cure the problem, noise may be entering the house wiring system via the power cable, which then acts as an antenna. It may be necessary to filter the noise from the power cable. First try a shielded power cable.

Line filters may be purchased from many electronics outlets. The same filter may be used to suppress noise conducted via the power cable from the computer to the power lines. Leads to printers and auxiliary equipment should be fitted with clip-on beads as a matter of course.

In some cases, particularly when using the 505DSP with portable computers, the transceiver may need to be physically separated from the computer and/or monitor. The 505DSP is provided with control cables approximately 8' (2.4m) long, but 25' (8m) and 75' (23m) cables are available optionally from Kachina. If using your own cables, they must be 100% shielded and not longer than 75' (23m).

## DSP Considerations.

At some stage in the receiver, we're going to digitize some signals and do filtering and other processing on them; we want this point to be as close to the antenna as possible, so in choosing I.F.'s, we have to look at available components to see what frequencies, bandwidths, and dynamic ranges we can utilize. We can eliminate many traditional analog signal processing stages if we can digitize the signals ahead of the point that expensive crystal or mechanical filters previously occupied. Our first tradeoff is between high-speed analog-to-digital converters (ADC's) and the costly filters they would replace. This decision is driven mainly by cost, although issues of current consumption and processing power definitely come into play.

The final compromise will also depend on the performance levels we expect to achieve. For example, many excellent ADC's exist which are quite capable of digitizing signals directly from the antenna; their sampling rates are fast enough for the job, but their dynamic ranges fall far short. As we'll see below, H.F. receivers must handle a tremendous range of input signal levels without flinching! So before we can make even this first decision about the receiver's conversion scheme, we have to think about what dynamic range is, and how much we want.

## Receiver Dynamic Range.

Every receiver's job is to produce a useful replica of what was sent from the other end, and to reject all other signals. In today's crowded H.F. bands, this is an increasingly difficult task! The desired signal might be quite weak, so we need good sensitivity-- this means lots of gain without the introduction of excess circuit noise.

Sensitivity must be specified as a function of the bandwidth (BW) of interest, since we're trying to copy a narrow-BW signal in the presence of noise, which exists at every frequency! In the specified BW, a signal received at the antenna terminals has a certain signal-to-noise ratio (SNR). We fight to preserve this SNR throughout the receiver. Electronic circuits introduce some noise, though, because the electrons are constantly bashing into conductor atoms and other electrons as they move. The ratio of the input SNR to the output SNR of a receiver is referred to as noise figure (NF), and is expressed in dB.

Originally explained by Einstein in 1905, Brownian motion of atoms and free electrons in any conductor produces an available noise power in Watts of:

$$P_{\text{noise}} = kTB$$

where  $k$  is Boltzmann's constant,  $1.38 \times 10^{-23}$ ,  $T$  is the absolute temperature in degrees Kelvin, and  $B$  is the BW in Hertz. "Plug and chug" on these numbers, and you'll find that at room temperature, or 293 K, and in a BW of 3 kHz, this power is -139 dBm or 12.1 attoWatts!  $12.1 \times 10^{-18}$  Watts represents the minimum discernable signal (MDS) in a perfect receiver using voice BW. Note that as the temperature is lowered, the possibilities increase linearly; a receiver operating in a liquid nitrogen bath is a real gem! However, atmospheric and cosmic noise are usually much greater than this theoretical limit.

The best H.F. receivers today have NF's around 7 dB, so at the point where the noise power is just equal to the signal power, the output SNR is 0 dB, and the input signal level is:

$$P_{\text{in 0dbsnr}} = -139 + 7 \text{ dBm} = -132 \text{ dBm}$$

We define this MDS level as the lower limit of the receiver's dynamic range. Now let's describe how to find the upper limit of the dynamic range. Because of the manifold ways receivers degrade at high input levels, we'll define dynamic ranges based on each of these.

## Receiver Overload-- Let Me Count The Ways.

Normally, overload phenomena involve large, off-channel signals. Of course, it's also possible to overload on a very strong desired signal; for most modern receivers, this level would be so high that radio communications wouldn't be necessary-- just shout out the window!

Large-signal performance is typically characterized by measuring the following effects:

3rd-order intermodulation distortion (IMD)

2nd-order IMD

"blocking", or de-sensitization

in-band IMD

Let's examine how each of these measurements is made, and the strengths and weaknesses of current methods.

#### IMD Dynamic Range and Intercept Point.

To measure IMD dynamic range, we inject two off-channel signals of equal amplitude, then measure the degradation in receiver performance. In this case, degradation comes in the form of an undesired, on-channel signal produced by the mixing of the off-channel signals. We increase the interfering levels until the on-channel signal produces a power just equal to the noise power. This is the definition of MDS given above.

We define the IMD dynamic range to be the ratio of this off-channel signal power to the MDS power, expressed in dB. In the ARRL method for 3rd-order IMD, one interfering signal is placed 20 kHz from center channel, the other 40 kHz from center. The 3rd-order intercept point (IP3) is calculated by assuming the receiver distortion obeys a perfect cube law; i.e., for every dB the interference (QRM) is increased, the 3rd-order IMD product will increase 3 dB, and the difference will increase by 2 dB.

IP3 is extrapolated, therefore, by adding half the 3rd-order IMD dynamic range to the QRM level obtained in the measurement above:

$$IP\ 3 = ( IMD\ D.R.\ / 2 ) + P\ QRM$$

This is supposed to be the level at which the 3rd-order IMD product is equal in amplitude to the QRM. Were we to actually inject QRM of this level, however, we might find a real IP3 much higher; receivers seldom obey perfect cube laws as they're predicted to do! This normalized procedure is a good basis for comparison, though.

In the 2nd-order test, we inject two non-harmonically-related signals and look for the undesired product at the sum or difference of the QRM. IMD dynamic range is measured as above, and IP2 is extrapolated by assuming the receiver obeys a perfect square law. For every dB of increase in the QRM, the 2nd-order product increases 2 dB, and the difference increases by 1 dB:

$$IP\ 2 = ( IMD\ D.R.\ ) + P\ QRM$$

But how can the receiver obey two apparently conflicting laws at the same time?! It's because in the 2nd-order case, we're mixing the two fundamentals of the QRM; whereas, in the 3rd-order case, we're mixing the fundamental of one with the internally-generated 2nd harmonic of the other.

Note that when we add two fundamental signals, the result is always greater than twice the frequency of one of the signals. For this reason, our 2nd-order performance can be improved by using half-octave bandpass filters ahead of the receiver front end. Such filters, switched or tuned as the receiver changes frequency, always provide attenuation to one of the interfering signals, reducing the deleterious effects.

#### "Blocking" Dynamic Range.

In this measurement, we inject a single off-channel source, and look for some degradation in the on-channel performance. In the ARRL method, the output power from a single, on-channel input signal is monitored. The QRM, 20 kHz away, is increased until the desired output power either increases or decreases by 1 dB.

A decrease is supposed to indicate that some stage or other is saturating, while an increase results in a "noise-limited" measurement. The blocking dynamic range (BDR) is calculated as the ratio of the QRM power in the measurement above to the MDS power, expressed in dB.

In reality, saturation seldom occurs in modern receivers before the noise takes over. This noise is the result of reciprocal mixing, wherein the QRM mixes with the phase noise sidebands of the local oscillator to produce in-band noise. A state-of-the-art synthesized local oscillator has phase noise in a 3 kHz BW, and at a 20 kHz offset, of around 100 dB below its injection level. Were the BDR measured using the SNR instead of the average output power, we could call it de-sensitization or "de-sense"; it would be on the order of 100 dB, and would be solely a measure of the

synthesizer phase noise. This number is quite a bit lower than that usually obtained with the ARRL method.

The difference becomes evident when trying to measure an IF-DSP receiver with a digital AGC system. Such a system holds the peak desired output level constant, and as the SNR degrades, the average output power will go down! In a conventional receiver, all other things being the same, the SNR would be identical, but the output power would increase because of the added noise. The peak-to-average ratio of noise is high, so monitoring the average or RMS output power wouldn't indicate an increase until much more QRM power were added.

To correlate the SNR method with the ARRL method, we might consider using a degradation in the output SNR as our criterion, as in the EIA standard; the amount of degradation could be chosen to equate the new measurements to BDR measurements already made on known receivers. Let's face it, it's the reciprocal mixing that gives the most trouble these days. If a blocking measurement is still desired, we ought to use the peak output level, not the RMS.

#### In-band IMD.

This is a measure of the distortion produced by the receiver when the only signals present are inside the desired passband. Current ARRL methods call for a two-tone input with a frequency separation of 100 Hz. This is excellent-- it's roughly the natural impulse frequency of the human voice system. The IMD product levels are examined relative to one of the tones. The AGC speed, if adjustable, is set to its fastest setting.

Digital AGC systems can again cause problems here, since they are capable of very fast attack and decay times. If the decay time can be set fast enough, clearly the two-tone will be subject to extreme distortion, since it'll begin "flat-topping". It doesn't make sense to defeat the very system designed to prevent the thing being measured!

#### Pre-amplifiers and Dynamic Range.

It's obvious that to achieve the best sensitivity, some gain ahead of the first mixer is required. If this gain stage has a low NF, we can improve the sensitivity by almost the amount of the gain. This extends the receiver's dynamic range on the low end.

It's difficult, though, to make up the difference on the high end. The large-signal handling will degrade by at least the amount of the pre-amplifier gain, so the dynamic range is generally reduced. Also note that dynamic range is just the ratio of maximum and minimum signals that can be handled, and says almost nothing about the actual large-signal handling capability! A receiver might have a greater dynamic range than another, and still have poor IP. Its sensitivity may be excellent, but it might not be a good large-signal performer.

## AN110: 505DSP - The AGC System

AGC in an IF-DSP Receiver.

H.F. receivers must handle signals as little as  $0.05 \mu\text{V}$ , and strong signals may approach 1 Volt, near the IP3. Expressed this way, the dynamic range can exceed 150 dB! As we expect the audio output level to remain relatively constant, and the distortion to stay within limits, a gain-control system has to be utilized. We must keep our analog stages linear, so an analog AGC system is mandatory.

The final selectivity in the 505DSP is provided using digital filters, as this eliminates the need for expensive crystal or mechanical filters. So it follows that some of the signals we digitize are going to be undesired-- this raises a problem: the digital filters are going to remove the QRM, all right, but the analog AGC will still be acting on the total bandwidth! A strong interfering signal will reduce the analog gain, as it must, and the level of our desired signal will fall as well. This is where the digital AGC system comes in.

### Digital AGC Algorithms.

To keep the desired signal's output level constant, we use a system which measures the ratio of total digitized signal energy to desired signal energy. When the QRM increases, this system compensates for the reduction in gain caused by the analog AGC. The effect is to hold the desired signal's peak level constant. Now we have to determine how to measure the critical ratio, and how and when to make adjustments in the gain.

Clearly, the digital gain compensation algorithm must use two data as inputs: the ratio of total signal level to desired signal level, and the actuation or amount of analog AGC. The ratio of the amplitudes is easily calculated by the DSP system, since it need only compare the peak digitized input level with the peak output level after filtering. This isn't quite the whole solution, however, since when the desired signal decreases, the system can't tell if it was because of QRM-caused analog gain reduction, or because the other station just stopped transmitting!

So we arrange to monitor the analog AGC voltage in order to find out what it is doing. It turns out we don't need to know the amount of analog gain reduction if we can adjust the digital gain fast enough. We examine the analog AGC to detect when the gain is decreasing rapidly, and when the amplitude ratio is also rapidly increasing, quickly boost the digital gain until the desired output level is maintained. Note that both the analog and digital AGC systems are fast-attack in all situations.

In practice, this system works quite well; the digital AGC decay time can be continuously adjusted as desired. On-channel signals are digitally boosted by the amount necessary to keep the peak output constant. The main drawback is that the dynamic range of the DSP system places restraints on how much digital gain can be applied. Using a 16-bit "sigma-delta" analog-to-digital converter, the DSP system's dynamic range approaches 100 dB.

It turns out this is quite adequate for the task, since IMD, reciprocal mixing, and other effects limit adjacent channel rejection performance before the DSP system comes into play. Refer to AN108 for more information on dynamic range issues.

### Restoring an AGC Threshold or "Knee".

The system described above is "kneeless", since the audio output remains constant regardless of input signal level. This arrangement can result in operator fatigue, because receiver and atmospheric noise are boosted in the absence of strong input signals. SSB receivers normally have AGC knees around  $2$  or  $3 \mu\text{V}$ , so that input signals under the knee produce audio output directly proportional to signal strength; i.e., they are in a linear region of the transfer function. Background noise is much quieter, and fatigue is reduced. To restore the knee in our digital AGC, we must re-examine the dynamics of the system, and redefine the goal in terms of actual system behavior.

In the kneeless system, the goal was to hold the desired signal's peak output level,  $n$ , constant. We measure the peak level of the total digitized signal as the sum of the desired,  $n$ , and the undesired,  $m$ :  $A_{\text{tot}} = m + n$ . Note that when the analog AGC is operating, this level  $m + n$  is almost constant; that's the job of any analog AGC: to adjust the gain to hold the output steady. When  $m + n$  is not strong enough to actuate analog AGC, then its value is proportional to input level, down to a minimum established by the receiver's noise floor.

To hold the desired signal's output level constant, therefore, we compute the ratio of the appropriate output level,  $A_{\text{out}}$ , to the actual level,  $n$ :

$$(1) k = A_{out} / n$$

$A_{out}$  is a constant determined by what output level is required by the receiver's audio processing system.  $k$  is the digital gain boost factor applied to the desired signal, such that the audio output always has peak level  $k n$ .

The desired signal's audio is peak detected in firmware to produce  $n$ . The decay time of this detector is continuously adjustable as AGC speed. As defined so far, the system works as long as the levels of undesired signals  $m$  aren't changing.

When strong QRM comes along,  $m$  increases rapidly; if the QRM is strong enough to actuate the analog AGC,  $n$  decreases rapidly. The DSP system monitors the analog AGC voltage to detect this rapid gain reduction, and temporarily sets the  $n$  detector's decay time constant to a fast value. As the rate of analog gain reduction slows, the decay time constant is returned to the value set by the AGC speed control. In this way, digital gain boost factor  $k$  adjusts itself rapidly upward at the onset of QRM, and settles at the proper value after  $n$ 's decay time constant returns to normal. This works well as long as the analog AGC slope detector can catch reductions in analog gain significant enough to affect  $n$ .

A servo loop is formed which does not need to have information about the actual amount of analog gain reduction, since the rapid adjustment of  $k$  is too fast to hear. When  $m+n$  isn't strong enough to actuate analog AGC,  $k$  remains inversely proportional only to  $n$ .

To re-introduce a knee to this system, however, does require information about the actual amount of analog gain reduction, as we'll see below. To begin the analysis, we'll consider 3 cases: In case 1,  $m$  and  $n$  are both under the analog knee; in case 2,  $m$  and  $n$  are both over the analog knee; and in case 3,  $m$  is over the analog knee, but  $n$  is under it. We'll have to determine what amount of analog gain reduction was caused by  $m$  and  $n$  separately to make it work.

A "Case" History.

In case 1, no analog gain reduction occurs. To make the digital AGC knee equal to the analog AGC's, we'd set  $k=1$ . In case 2, since the desired and undesired signals are each large enough to actuate analog AGC by themselves, the correct value for  $k$  is almost the same as in the kneeless system; it's the ratio of the total signal to the desired:

$$(2) k = (m + n) / n$$

In this case we know  $m+n$  is nearly constant, and we're tempted to just use Eq. 1.

Case 3 is where the headaches start.  $m$  alone is responsible for the gain reduction, and we have to know its actual amount. The most straightforward method is to build a table of the gain reduction factor versus analog AGC voltage. As each receiver may have a slightly different characteristic, we may want to adaptively alter the table, or we may calibrate the table on a unit-by-unit basis. The intention is to formulate the 3 cases into a uniform system. As each of the cases is likely to occur in our receiver, and because rapid switching between states is possible, we need a case-detection algorithm which operates in "real time".

Let's try a comparison between the analog gain reduction value  $k$  from the table and the wanted-to-actual output ratio (Eq.1). Is the following true:

$$(3) k_{table} \leq A_{out} / n$$

For case 1, this inequality will always be true, because  $k_{table}=1$ , and  $n \leq A_{out}$ . For case 3, it will also be true, since  $n \leq A_{out}$  and  $m \leq m+n$ , where  $m$  is solely responsible for the gain reduction factor  $k_{table}$ . For case 2, it will be false, because:

$$(4) k_{table} > (m + n) / n$$

and:

$$(5) (m + n) \sim A_{out}$$

The analog gain reduction must be more than that caused by either  $m$  or  $n$  separately, each of which is strong

enough to actuate analog AGC by themselves.

Although case 3 was the standout above, it is paired with case 1 in the final algorithm here, leaving case 2 by itself. We'll use a digital gain boost factor  $k=k_{table}$  in cases 1 and 3, and use Eq. 1 for case 2. Note that the resulting transfer function is continuous, because the segments corresponding to the truth or falseness of the inequality in Eq. 3 above must meet where  $k_{table}=A_{out}/n$ . See Fig. 1.

#### Making the Knee Continuously Adjustable.

Although the transfer function is produced in 2 segments, adjusting the threshold requires only a modification of the test made in Eq. 3. If we apply a scaled value of  $k_{table}$  to the inequality, the knee will move downward according to the scaling factor. The test becomes:

$$(6) p(k_{table}) \leq A_{out} / n$$

where  $p >= 1$ . The variable  $p$  has the effect of narrowing the width of cases 1 and 3 on the transfer curve. At some point,  $p$  gets very large, and the inequality is never satisfied; the knee is infinitely low, and the system is exactly equivalent to the kneeless system initially described. When  $p=1$ , the digital knee is equal to the analog knee. For values of  $p$  between 1 and the maximum value of  $A_{out}/n$ , the knee is continuously adjustable.

For a receiver with a noise floor of  $0.05 \mu V$  in a  $500 \text{ Hz BW}$ , and with an analog AGC knee of  $3 \mu V$ , the linear portion of the transfer function would be almost  $36 \text{ dB}$  deep. Without the adjustable threshold, this may be a bit much. While listening to a very weak signal with the volume cranked up, a strong signal may come along and blow the operator off their chair! Adjustment of the threshold allows maximum flexibility in operation. The control is actually used much as an RF gain control would be. Gain is reduced (threshold raised) while copying stronger signals to keep the noise floor quiet, and gain is increased (threshold lowered) during periods of low-level signal operation.

## AN112: 505DSP VFO Functions: Splits, Offsets, and Manual Tuning

### Simplex versus Split.

Two VFO modes are available on the 505DSP: simplex and split. In simplex operation, the transmitter operates on the same frequency as the receiver; in split operation, the transmitter and receiver frequencies are different. The mode may be changed by either clicking the mouse on the "VFO" button, or pressing the "S" key.

Simplex mode is used for most communications. Split mode is usually reserved for working DX stations, or for working cross-band during certain contests. We'll see that the 505DSP system offers several ways of obtaining receiver and transmitter frequency offsets for various operating situations. Note that in LSB, USB, and AM, the displayed frequency is that of the carrier, which is suppressed in the SSB modes. In CW, the display indicates the frequency of the actual RF signal. When changing to or from CW mode, the receiver will shift by the amount of the CW offset selected. See AN102 for more detail.

### Receiver Incremental Tuning (RIT).

The RIT control, when set to a non-zero value, is active in both simplex and split modes. It functions as a "clarifier" or receiver fine tuning control, and is most useful in correcting small frequency errors in stations being worked as part of a net, or in following other stations which are slowly drifting. It affects the receiver frequency only. Its total range of +/-10 kHz is large enough to allow DX splits to be worked, but split mode is easier to use for this purpose, as we'll see below.

During a round-table or net operation, not all stations may be transmitting on exactly the same frequency. The RIT can be used to correct for the error of any one station. When this off-frequency station quits transmitting and an on-frequency station begins, the "Z" key can be used to zero the RIT. When the off-frequency station returns, the "BACKSPACE" key will "undo" the zeroing command, restoring the previous RIT offset. "SHIFT BACKSPACE" will "redo" the command, again zeroing the RIT. The command stack is 10 deep, and so many different offsets can be stored and recalled in this way through multiple "undo" and "redo" operations.

Alternatively, the "snapshot" keys F1-F8 may be used to store and recall the proper RIT settings; or, split mode may be used. In split mode, manually tuning the receiver doesn't alter the transmit frequency. A tuning knob accessory, 505TK, is also available for those wishing to use a rotary control for fine tuning in the split mode. A trackball or second mouse can also be used instead of the knob, using compatible software versions 2.x3 or higher.

### Operating in Split Mode.

As mentioned above, this mode is useful for implementing receiver fine tuning without using the RIT control. But most often, it's used to work DX stations operating with a large frequency split, or during cross-band contests. The general procedure is described below. While operating split, it's necessary to listen on your transmit frequency before actually transmitting. It's illegal (47 CFR 97.101[d]) and discourteous to knowingly interfere with another transmission, and that's the danger during the excitement of a DX pileup. The 505DSP makes it easy to comply.

#### To set up a DX split:

1. Tune the receiver to the DX station's frequency. Listen for the announcement about where they are listening, usually 5 to 10 kHz upward. You have to hear 'em to work 'em!
2. Select split mode by either clicking the VFO button or pressing the "S" key.
3. Select the TX VFO by either clicking on the TX frequency window, or pressing the "V" key. You're now listening on your transmit frequency.
4. Tune the transmit frequency until the pileup is heard. At this point, you may wait for a break and go ahead and transmit. When the PTT is released, the receiver will automatically return to the frequency of the RX VFO, where the DX is found. Alternatively, you may press "V" or use the mouse to listen on the RX VFO.
5. The DX will respond with a call sign and report. If it's yours, great! If not, switch rapidly to the TX VFO ("V" key), and tune the transmit frequency until you hear the station which was called.

6. Return to the RX VFO ("V" key again) and wait for your chance to transmit. You'll now be transmitting on the frequency of the station last worked-- your chances are increased.

Repeat steps 5 and 6 until you are successful. After a little practice, this procedure will become easy.

Note that, when going from split mode to simplex, the TX frequency is forced equal to the RX frequency. Also, from split mode, the "=" button or the "=" key will set them equal. In this case, however, the frequency used will be the one which was active at the time the "=" was activated.

#### Cross-band Operation.

The 505DSP is capable of cross-band operation. In fact, the limit to the difference between the RX and TX frequencies is only set by the TX range of the transceiver. By treaty, the world is divided into 3 regions by the International Telecommunications Union (ITU), an arm of the United Nations. The United States of America resides in Region 2. Band limits for the 80 and 40 metre bands vary with region. E.g., European stations (Region 1) cannot transmit above 7100 kHz. U.S. stations cannot transmit phone below 7150 kHz. To work Europe on phone on 40 metres, therefore, means a large split or cross-band operation must be used.

#### Selecting Tuning Step Size.

Each VFO may be tuned with a step size selectable in decades. I.e., you can tune in 1 Hz steps, 10 Hz steps, 100 Hz steps, and so on, up to 10 MHz steps. The step size is selected using either the mouse or the "left arrow" and "right arrow" keys on the keyboard. The decade being tuned is highlighted on the display, and "+" and "-" areas may be used with the mouse for incrementing or decrementing the frequency. Alternatively, the "up arrow" and "down arrow" keys may be used for tuning.

When the mouse is being used, the first click and release on a frequency digit will leave the mouse pointer on either the "+" or "-" area, depending on whether the top or bottom half of the digit was targeted. The next click will either increment or decrement the frequency. If the mouse button is held down, the frequency will auto-repeat at a rate of approximately 20 steps per second. Likewise, when the keyboard is used, the tuning will auto-repeat at a rate of up to 20 steps per second. This maximum rate may be reduced-- and the delay until auto-repeating begins adjusted-- by changing the keyboard parameters under the operating system's "Control Panel" window. The rate associated with the external tuning knob accessory, 505TK, is not affected by these settings.

#### Automatic Tuning Rate Acceleration.

When either the 1 Hz or 10 Hz decade is auto-repeating during tuning, the rate of frequency change automatically increases linearly from its initial value of up to 20 steps per second to a maximum rate of 200 steps per second during the first 10 seconds. This rate change is reflected on the display, and in the steady state, the next higher digit is the least-significant one changing. E.g., with the 10 Hz digit selected, the "up arrow" key is depressed and held down. Tuning begins at a maximum rate of 200 Hz per second, and increases to a maximum of 2 kHz per second after 10 seconds. Tuning with the mouse and 505TK also experience this rate acceleration. This feature makes it easier to tune across the band while keeping the 10 Hz digit selected for fine-tuning purposes later.

#### Tuning in Steps of 5.

Using "SHIFT up arrow" and "SHIFT down arrow" will achieve tuning in increments of 5 times the currently-selected decade. The "x5" function can be locked on by setting the "CAPS LOCK" on the keyboard when the feature is enabled under the "Special" menu. "x5" will appear on the display near the frequency readout. This feature is convenient for tuning international broadcast stations, which are usually spaced every 5 kHz. Or it can be used to temporarily speed up manual tuning when using otherwise small step sizes.

## Overview.

Based on feedback from users, and in keeping with our philosophy of continual product improvement, several firmware upgrades have been developed for the 505DSP. The firmware is the set of instructions executed by the internal microprocessor, governing the performance of basic transceiver features. While many advanced features can be implemented by changing only the control software running on the P.C., all the digital signal processing (DSP) is performed inside the transceiver.

Firmware upgrades can be performed on all units returned to the factory, for a nominal charge, and within the limitations of any hardware changes. As of this writing, there are no backward hardware incompatibilities; all transceivers may use the major firmware upgrade described below. For field upgrades, a ROM Option Card is available. This credit-card-sized PCMCIA unit slides into one of the 2 slots provided on PC401, the front-most module in the 505DSP.

## Compatibility Issues.

Full exploitation of the features of the new firmware requires the upgrade of the control software running on your P.C. New software is provided on diskette with each ROM Option or factory upgrade. Alternatively, the new software may be downloaded from the Kachina web-site at "kachina-az.com".

Older V1.xy, V2.xy or V3.xy software will not be able to control features added in V4.x1 firmware, where x and y can be any integer. Inversely, V4.x1 software will attempt to control features lacking in V1.xy, V2.xy or V3.xy firmware. The key is that versions of firmware and software having the same major revision level, indicated by the first digit of the version number, are guaranteed compatible. To be safe, do not attempt to use V4.xy software with V1.xy, V2.xy or V3.xy firmware.

## New and Enhanced Firmware Features.

The new characteristics of V4.x1 firmware are described below. We developed the enhancements through careful deliberation; they came about because of valuable feedback from both potential and existing customers. Our design, attentive to the evolving needs of radio amateurs and professionals alike, is a dynamic one which is hardly cast in concrete.

In addition to the enhancements added by the previous version of firmware (as detailed below), V4.xy firmware and software add support for the following:

505TVR Transverter Option - 505TVR is factory-fitted only and requires upgrade to V4.xy software and firmware.

FM Option support is included in software only. FM will be a ROM-card option and may be fitted later. CTCSS tone selection is included.

Switchable TX bandwidth (BW) for 505ETX Option only. The two selections are 3.1 and 4.0 kHz. This command is found in the Tx menu. Exercising the BW control with non-ETX firmware results in 3.1 kHz BW at all times.

Improved speech-monitor circuit (PC401 Rev. C and higher only). Version 4 firmware may be used on older transceivers but the improved speech monitor requires a new PC401 module.

Version 4 is not backward compatible with previous firmware and software. Customers with V3.xy firmware or older should not attempt to use V4.x1 or later software. Firmware and software must be upgraded simultaneously to version 4.

## Features Previously Added by V3.xy Firmware:

1. Continuously-adjustable AGC threshold. This control acts much as an RF gain control would do, only better; it allows you to set the digital AGC "knee" to any level from 0 to 36 dB above the receiver noise floor.
2. Amplifier tune-up. A feature has been added to allow the output of a continuous CW signal for the tuning of external R.F. power amplifiers and other accessories. It allows you to set the power output of the signal to optimize your performance at actual transmit levels.

V3.x1 firmware includes enhancements to existing features:

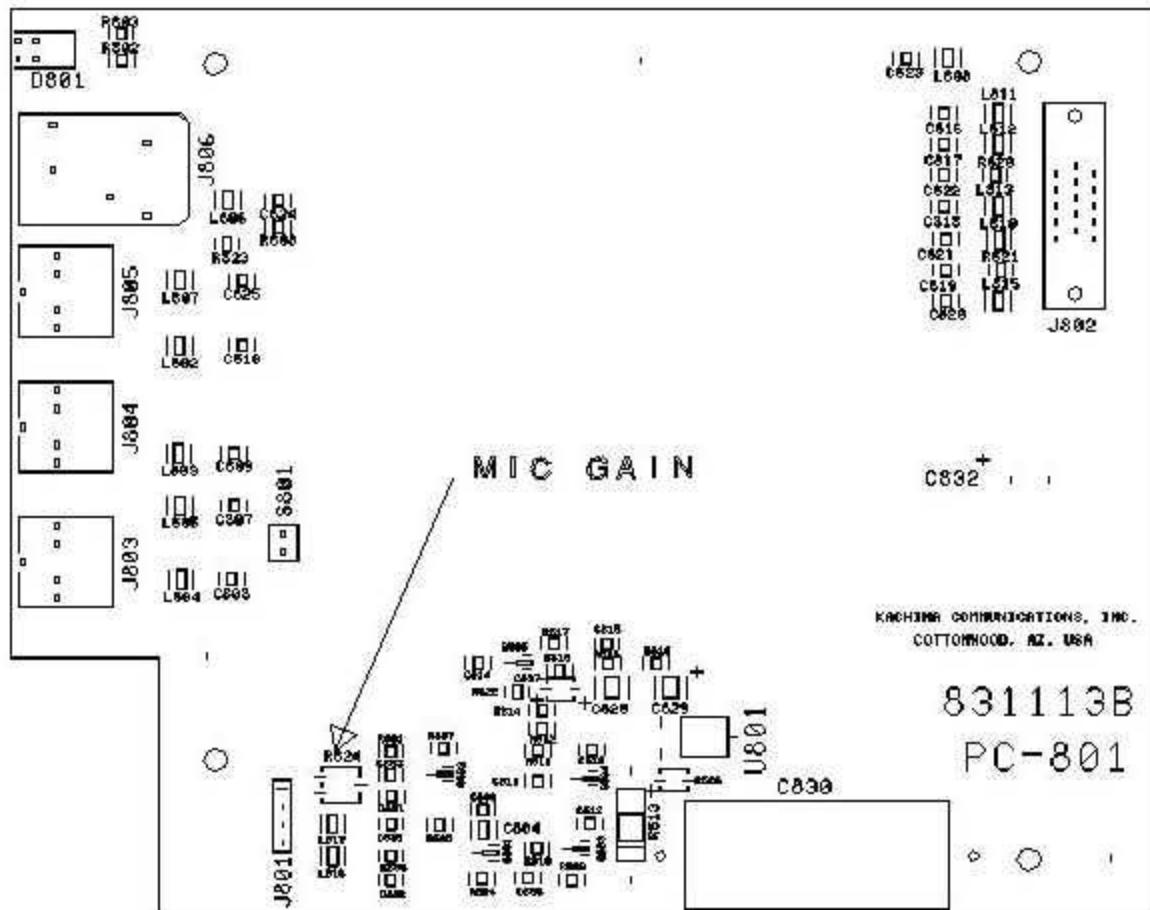
3. Improved S-meter performance below S-6. This enhancement increases the resolution and accuracy of the S-meter at low signal levels, and vastly improves the presentation of the Band Sweep Display and the performance of the level-sensitive squelch.
4. Speech processor linearization. The R.F. compression level control is now calibrated in dB, and the control is continuously adjustable for maximum effectiveness and minimum distortion.
5. Multi-tone Auto-notch. The Auto-notch feature is greatly enhanced to provide much better performance under noisy conditions. In addition, the Auto-notch will now eliminate multiple tones and other QRM.
6. Anti-vox sensitivity. The Anti-vox sensitivity has been improved to allow operation in higher ambient-noise environments.

## AN 116: Using the 505HG (Heil Goldline Microphone) with the 505DSP

In addition to following the instructions included with the 505HG, it is sometimes necessary to increase the gain of the 505DSP's microphone pre-amplifier to drive the transmitter into ALC. The 505HG does not put out quite as much level as some other dynamic microphones when used 5-8 cm (2-3") from your mouth, as recommended.

Gain adjustment may be accomplished by opening the Control Head. Remove the cover's six fasteners. Using a small blade screwdriver, rotate R824 counter-clockwise 1/8 to 1/4 turn to increase the gain (see Fig 1 below). 1/4 turn will increase gain about 6 dB.

The 505DSP's microphone pre-amplifier is designed to "soft-clip" at input levels that would overdrive the modulator. Operate the transceiver's on-screen MIC GAIN control only as high as it takes to get 2-4 dB of ALC indication on voice peaks when speaking in a normal tone.



## AN 117: Control of 505DSP By Third-Party Logging or Packet-Cluster Programs.

### Introduction

Beginning with V3.x5, where "x" is the ITU region number, Kachina 505DSP control software employs a file-based interface for use by third-party programs to set or retrieve frequency and mode. The details of this interface are described below.

### Activating the Interface

After the Kachina software is installed and running, select Station Settings from the Special menu, then select Interface To Other Logbooks, then select On. A file named "kc505\_nn.dat" will be created, where "nn" is the "com" port number. This file holds frequency and mode data as discussed below. The user may turn this function off when not using the interface.

### Accessing Frequency and Mode Data

Frequency and mode data may be read from the file by third-party software. The data in the file will be updated whenever the user makes a frequency or mode change. Each file I/O should take under 50 msec. The data could change 10-20 times a second while the user manually tunes the radio. If the user doesn't change the frequency or mode, then the data could go unchanged for minutes or even hours.

### Commanding the Transceiver

Third-party software may request a frequency or mode change by writing the new data to the file. The control software will check for requests once a second. If the user is commanding the radio when a request is received, the request will be deferred until the user stops commanding the radio.

### Notes on Operation

Every user has always had the ability to operate multiple 505DSPs from a single PC. A separate copy of the control software is needed for each radio, and obviously a separate communication port. You will need a separate copy of the shared interface file for each copy of control software running.

The file lock data are included to prevent data corruption caused by simultaneous I/O. If the lock is set to 1, don't read/write the file. If the lock is 0, set the lock to 1, do your read/write, then set the lock to 0.

If the radio is set to simplex and a TX frequency request is received, it will be ignored. The user must manually place the transceiver into split mode if split operation is anticipated.

### File Format

File lock (0 or 1) - 2 byte integer  
Current RX frequency in Hz - 4 byte integer  
Current TX frequency in Hz - 4 byte integer  
Current mode (1 - 5) - 2 byte integer  
Requested RX frequency in Hz - 4 byte integer  
Requested TX frequency in Hz - 4 byte integer  
Requested mode (1 - 5) - 2 byte integer

If more data are needed in the future, we will append it to the current format, so the file will always be backward-compatible with earlier versions of software.