

A Complete Self-Contained Audio Measurement System

This automatic, autoranging audio analyzer has the signal source, distortion analyzer, and counter to make the measurements most often needed in audio-frequency testing.

by James D. Foote

HEWLETT-PACKARD'S NEW MODEL 8903A Audio Analyzer (Fig. 1) is a complete audio measurement system for quick and accurate characterization of systems and signals in the frequency range 20 Hz to 100 kHz. The starting point for the 8903A is the classical distortion analyzer. Added to this are microprocessor control, a reciprocal frequency counter, rms detectors, and a programmable audio source. These provide accurate measurement of ac level, distortion, SINAD,* signal-to-noise ratio, and dc level. The audio source and the measurement circuits can work independently or together. The source is programmable in frequency and level and has very low distortion. The measurement circuits can monitor this internal source or any other independent input waveform. Together the source and measurement input can be used for swept response measurements.

*SINAD = ratio of signal plus noise plus distortion to noise plus distortion in a receiver's output.

All measurements are available at the push of a button. No knob adjustment or operator interaction is necessary. One simply applies the signal and selects the measurement mode. All control and processing are handled by the internal microprocessor. The microprocessor monitors the input signal and makes internal gain and frequency adjustments as required.

In automatic measurement systems, the 8903A is capable of rapid and straightforward remote control. Analyzer operations can be controlled and all measurements can be transferred via the Hewlett-Packard Interface Bus (HP-IB). Hewlett-Packard's implementation of IEEE Standard 488-1978. On the bench, the 8903A allows rapid and accurate circuit characterization when many repetitive measurements are necessary.

Major application areas for the 8903A Audio Analyzer are general audio testing, transceiver testing, and automatic systems. In general audio testing, the 8903A measures the



Fig. 1. Model 8903A Audio Analyzer makes the accurate measurements needed to characterize systems and signals in its frequency range of 20 Hz to 100 kHz. It has applications in general audio testing, transceiver testing, and automatic systems. Microprocessor control makes it automatic and easy to use.

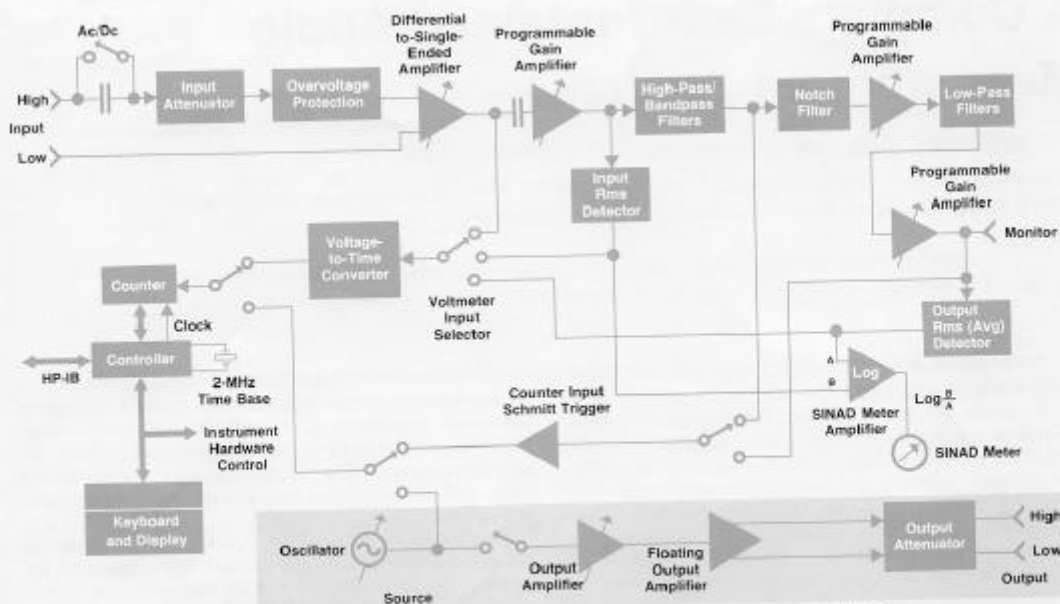


Fig. 2. The 8903A Audio Analyzer is basically a distortion analyzer, with a tunable notch filter to remove the fundamental frequency component of the signal and a detector to measure what remains, which consists of noise and distortion. Added to this are a microprocessor-based controller, a reciprocal frequency counter, a programmable audio source, and other resources, forming a complete audio measurement system.

frequency response and distortion characteristics of filters, high-quality amplifiers, audio integrated circuits, and other devices. The frequency of the internal source can be swept while making measurements in all modes. The analyzer provides recorder outputs and scaling for easy generation of plots using an X-Y recorder.

For transceiver applications the most common receiver measurements are SINAD for FM receivers and signal-to-noise ratio for AM receivers. A psophometric filter is included for making measurements to CEPT standards. Common transmitter measurements such as audio distortion and squelch tones are made using the 8903A with its companion instrument, the 8901A Modulation Analyzer.¹ In automatic systems, the 8903A provides many frequently needed audio functions, doing the work of an audio synthesizer, digital multimeter, frequency counter, and tunable notch filter. More details on specific applications are presented on page 6.

Control Philosophy

Front-panel control of the audio analyzer is simple, yet powerful. Most functions can be used and understood with very little training. The casual user can select amplitude, frequency, measurement mode, and filtering simply by reading the labels on the controls. More details are available on the instrument's pull-out card.

A great deal of measurement sophistication is built into the 8903A software. Measurement routines are structured to optimize measurement speed and accuracy. As a rule,

measurements triggered from the bus or initiated from the keyboard are accurate from the first reading. The operator needn't wait for successive measurements to verify that the reading has stabilized. The software algorithm monitors key voltages in the audio chain and waits until they stabilize before taking data. Not only does the software perform these functions much more rapidly than the operator, but it can also ensure the optimal convergence of the measurement with a repeatable, well defined technique. Distortion, SINAD, and signal-to-noise ratio, in particular, are examples of measurements that in the past required a significant amount of settling time and operator interaction. A classical distortion analyzer requires repeated adjustments to achieve an accurate distortion reading. More recent analyzers have offered semiautomatic tuning and leveling, but response time is often long, and operator interaction is required if the frequency, amplitude or relative distortion of the signal changes significantly. With the 8903A, some delay still exists, but the delay involved is minimized by careful circuit design and microprocessor control.

Special functions extend user control of the instrument beyond that normally available from the front panel. These functions are intended for the user who knows the instrument and the service technician who needs arbitrary control of the instrument functions. Automatic tuning and ranging, overvoltage protection, and error messages protect the user from invalid measurements during normal operation. When special functions are used, some of these

safeguards are removed, depending on the special function selected, and thus there is a degree of risk that the measurement may be invalid. However, there is no risk of damage to the instrument.

To enter a special function, the user enters the special function code (usually a prefix, decimal, and suffix) then presses the SPCL key. The special function code appears on the display as it is being entered. If a mistake is made during entry of the special function code, the user can press the CLEAR key and start over. When a special function is entered, the light in the SPCL key goes on if it is not already on. The readout on the display depends on the special function entered. It may be a measured quantity, an instrument setting, or a special code. In some cases the display is unaltered. Special functions can be entered from the HP-IB by issuing the special function code followed by the code SP.

Floating Input and Output

To eliminate troublesome ground loops, both the source output and the measurement input of the 8903A are floating. This is helpful in low-distortion or low-level ac measurements when it is necessary to reject potential differences between the chassis of the 8903A and the device under test. The 8903A also has EMI (electromagnetic interference) protection built into the source output and measurement input lines so that it can work in the presence of high EMI. All of the analog circuitry is shielded by an internal EMI-tight box. The output and input lines extending from this box to the front panel are shielded and terminated in BNC connectors. For user convenience, the BNC connectors are spaced so that BNC-to-banana adapters can be attached. Thus a banana or twisted-wire connection can be made to the instrument when EMI shielding is not critical.

Analyzer Architecture

The 8903A Audio Analyzer combines three instruments into one: a low-distortion audio source, a general-purpose voltmeter with a tunable notch filter at the input, and a frequency counter. Measurements are managed by the microprocessor-based controller. This combination can make most common measurements on audio circuits automatically. To add to its versatility, the analyzer also has selectable input filters, logarithmic frequency sweep, X and Y outputs for plotting measurement results versus frequency, and HP-IB programmability. Fig. 2 is a simplified block diagram.

The amplitude measurement path flows from the INPUT jacks (HIGH and LOW) to the MONITOR output on the rear panel, and includes the input and output rms detectors, the dc voltmeter (the voltage-to-time converter and counter), and the SINAD meter circuitry. Measurements are made on the difference between the signals at the HIGH jack and the LOW jack. Differential levels can be as high as 300V. Signals that are common to both the HIGH and LOW jacks are balanced out. Signals applied to the LOW jack must not exceed 4V.

The input signal is ac coupled for all measurement modes except dc level. The signal is scaled by the input attenuator to a level of 3V or less. To protect the active circuits, the overvoltage protection circuit quickly disconnects the input amplifier if its input exceeds 15V.

The differential signal is converted to a single-ended signal (referenced to ground) and amplified. The signal is further amplified by a programmable gain amplifier, which is ac coupled. The gain of this amplifier and the differential-to-single-ended amplifier are set to keep the signal level at the input rms detector between 1.7 and 3V rms to optimize its effectiveness and accuracy.

The output from the first programmable gain amplifier is converted to dc by the input rms detector and measured by the dc voltmeter. The output of the detector is used to set the gain of the input circuits and becomes the numerator of the SINAD measurement and the denominator of the distortion measurement. The gain of the input path is determined by measuring the dc level. The input rms detector also monitors the ac component (if there is one) and lowers the gain of the input path if the ac signal will overload the input amplifier. At this point either the 400-Hz high-pass filter or the psophometric filter can be inserted into the signal path. The 400-Hz high-pass filter is often used to suppress line hum or the low-frequency squelch tone used in some mobile receivers. The psophometric filter has a bandpass frequency response that simulates the "average" response of human hearing. It is often used to condition a receiver audio output when determining the receiver's input sensitivity.

During SINAD, distortion, or distortion level measurements, the fundamental of the signal is removed by the notch filter. The output from the filter is the distortion and noise of the signal. In the ac level and signal-to-noise modes the notch filter is bypassed. After amplifying and low-pass filtering, the output from the notch filter is converted to dc by the output rms detector and measured by the dc voltmeter.

During distortion or distortion level measurements, the notch filter is tuned to the frequency counted at its input. Coarse tuning is done by the controller, and internal analog circuitry fine tunes and balances the notch filter. During SINAD measurements, the controller coarse tunes the notch to the source frequency. Thus a SINAD measurement is normally made with the internal source as the stimulus; this permits measurements in the presence of large amounts of noise (where the controller would be unable to determine the input frequency). If an external source is used in the SINAD measurement mode, the source frequency must be within 5% of the frequency of the internal source.

The two programmable gain amplifiers following the notch filter amplify the low-level noise and distortion signals from the notch filter. The overall gain of the two amplifiers is normally set to maintain a signal level of 0.25 to 3V at the output detector and monitor. The 30-kHz and 80-kHz low-pass filters are selected from the keyboard. With no low-pass filtering, the bandwidth of the measurement system is 750 kHz. The filters are most often used to remove the high-frequency noise components in low-frequency distortion and signal-to-noise measurements. The output from the second programmable gain amplifier drives the rear-panel MONITOR output jack. Taking advantage of the increased amplification available at this point, the counter monitors this output in ac level and signal-to-noise modes.

The output rms detector is read by the dc voltmeter in the ac level, SINAD (the denominator), distortion (the numerator), distortion level, and signal-to-noise measure-

Audio Analyzer Applications

The 8903A Audio Analyzer's measurement capabilities reach far beyond conventional distortion analyzers. Much of this performance results from microprocessor control and HP-IB programmability. Numerous hardware features such as a fast counter, both analog meter and digital display, and switchable detector filtering allow the user to make unusual or special measurements with convenience and little auxiliary apparatus.

Consider the 8903A used at a test and calibration station in the manufacture of audio power amplifiers. A typical sequence of events might include an output offset null, frequency response check, distortion test, and noise measurement. The 8903A can perform all these measurements quickly with a single test setup. If an X-Y plotter is connected to the rear-panel outputs, the results of swept frequency measurements can be recorded on standard log/log or log/lin graph paper.

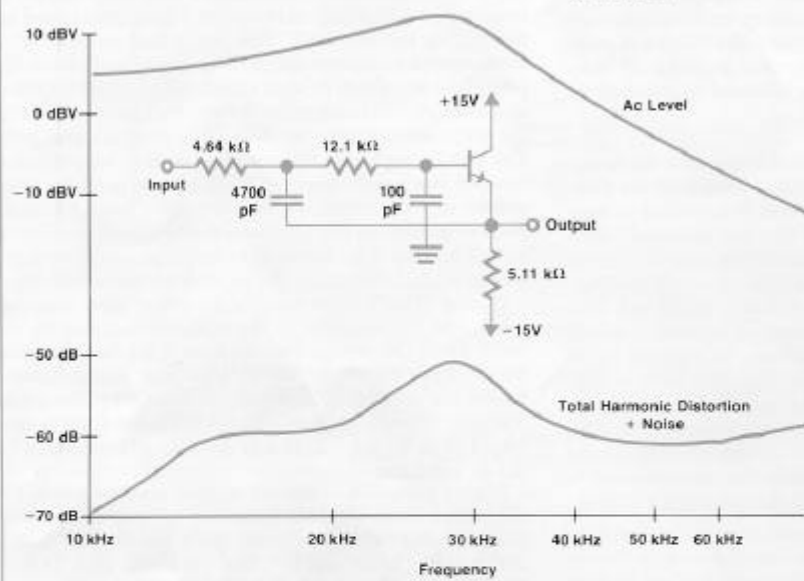


Fig. 1. Swept distortion and frequency response of a two-pole active filter, measured by the 8903A Audio Analyzer.

ment modes. It is also used to set the gain of the two programmable gain amplifiers. The detector can be configured internally to respond to the average absolute value of the signal instead of the true rms value. This option is provided because some measurement specifications for detection of distortion and noise specify the use of an average responding detector. Average responding detectors do not give an accurate indication of signal power unless the signal waveform is known. (If the waveform is Gaussian noise the reading will be approximately 1 dB less than the true rms value.)

In the SINAD mode the outputs from the input and output rms detectors are converted to logarithms, subtracted, and converted to a current by the SINAD meter amplifier to drive the SINAD panel meter. Since SINAD measurements are often made under very noisy conditions, the panel meter makes it easier to average the reading and to discern trends.

The voltage-to-time converter converts the dc inputs into a time interval, which is measured by the counter.

The 8903A uses a reciprocal counter. To measure frequency, it counts the period of one or more cycles of the signal at its input. Then the controller divides the number of cycles by the accumulated count. The reference for the counter is the 2-MHz time base, which also is the clock for the

controller. The counter has four inputs and three modes of operation:

1. Voltage measurement. The time interval from the voltage-to-time converter is counted. The accumulated count is proportional to the dc voltage. For direct measurements (ac level and distortion level), the count is processed directly by the controller and the result is displayed. For ratio measurements (SINAD, distortion, and signal-to-noise), the counts of two successive measurements are processed and displayed. For SINAD and distortion, the controller computes the ratio of the outputs of the input and output rms detectors. For signal-to-noise measurements, the output of the output rms detector is measured with the oscillator on and off and the ratio of the two measurements is computed.
2. Input frequency measurement. The signal from the last programmable gain amplifier or the high-pass/bandpass filters is conditioned by the counter input Schmitt trigger to make it compatible with the counter's input. The period of the signal is then counted, the count is processed by the controller, and the frequency is displayed.
3. Source frequency measurement. The counter measures the frequency of the oscillator during tuning and when

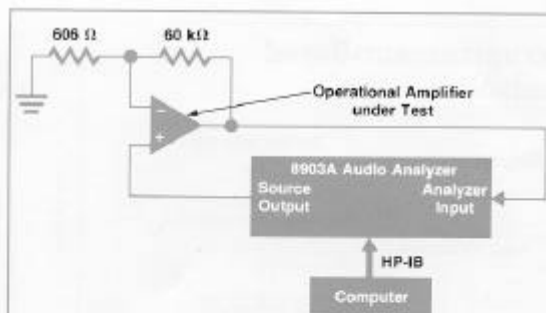


Fig. 2. Test setup for screening operational amplifiers.

The swept frequency measurement with plot capability finds many applications in the laboratory. Fig. 1 shows the swept distortion and frequency response of a two-pole active filter. The upper curve shows the magnitude response, while the lower shows distortion. Notice that the analyzer input magnitude covers a 30-dB dynamic range. During the sweep, the 8903A is automatically setting input gain before performing the distortion measurement.

If an HP-IB controller is available, there are many more applications for the 8903A. Fig. 2 shows a simple test setup for screening operational amplifiers. With no other instruments in the system, a computer-controlled 8903A can rapidly and accurately measure input offset voltage, input noise voltage, and distortion. It can also be used to measure the gain-bandwidth product of the op-amp, provided it is not greater than 30 MHz. The controller test program can be written to provide either a go/no-go output or a listing of measurement results.

Many of the special measurement modes are available through the use of special functions. For example, the 8903A can be used as a test amplifier with gain settable from -24 dB to +94 dB. Signal filtering can be added by selecting the appropriate front-panel controls, or the special functions can be used to put the instrument into a notch or bandpass filter mode. Of course, signal frequency and amplitude will be measured and displayed if the operating loads are chosen correctly. This mode of operation makes it possible to count

verifying that the oscillator frequency is within tolerance. This frequency is normally not displayed.

The source covers the frequency range of 20 Hz to 100 kHz. It is tuned by the controller to the frequency entered from the keyboard, using a tune-and-count routine. (Note that the frequency is not obtained by frequency synthesis.) The switch following the oscillator is closed except in the signal-to-noise measurement mode or when an amplitude of 0V is entered from the keyboard.

The source output amplifier and output attenuators provide 77.5 dB of attenuation in 2.5-dB steps, and 2.5 dB of attenuation in 256 steps. This gives an open-circuit output from 0.6 mV to 6V. The floating output amplifier converts the ground-referenced input to a floating output. Either output, HIGH or LOW, can be floated up to ten volts peak.

The entire operation of the instrument is under control of the microprocessor-based controller, which sets up the instrument at turn-on, interprets keyboard entries, executes changes in the mode of operation, continually monitors instrument operation, sends measurement results and errors to the front-panel displays, and interfaces with the HP-IB. Its computing capability is also used to simplify circuit operation. For example, it forms the last stage of the

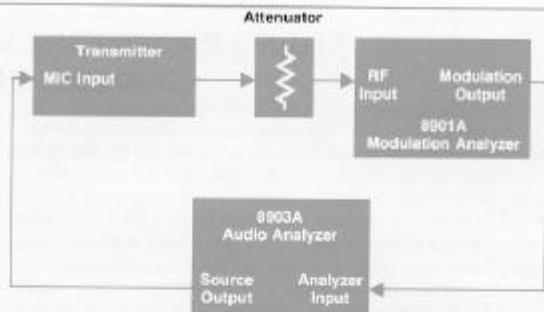


Fig. 3. Transmitter test setup using the 8901A Modulation Analyzer and the 8903A Audio Analyzer.

the frequency of a signal whose amplitude is in the low tens of microvolts.

Transceiver Testing

Much effort has gone into the design of the 8903A to facilitate the audio measurements required for automatic, programmed, transceiver testing. For example, the worst-case source frequency error of 0.3% allows the 8903A source to replace a synthesizer for squelch-tone generation. In addition, a binary programming mode is available through the HP-IB that causes the 8903A to generate a tone burst sequence that can be used to unlock a coded receiver. A related function allows the 8903A to measure burst tones generated by an external source, such as a transmitter under test. The packed, four-byte output allows the 8903A to output frequency measurements as often as every eight milliseconds.

When the 8903A is used in conjunction with the 8901A Modulation Analyzer, almost all transmitter tests can be automated. Fig. 3 shows the block diagram. With the source turned off and the transmitter keyed, the squelch-tone frequency can be counted. Then, the 400-Hz high-pass filter can be switched in to eliminate the squelch tone from the remaining measurements. With the source output turned on, the 8903A can easily be programmed to make the necessary measurements to determine distortion and microphone sensitivity.

counter, converts measurement results into ratios (in % or dB), and so on. It also executes routines for servicing the rest of the instrument as well as itself.

Input Circuits

Numerous design constraints were imposed on the input attenuator/protection/amplifier circuitry. An input impedance of 100 k Ω is necessary to prevent the input circuitry from unnecessarily loading the device under test. On the other hand, maintaining a good signal-to-noise and distortion ratio, good frequency response to 100 kHz, automatic operation and reliable performance with input signals as large as 300V is very challenging. Consider the 300V constraint. It is possible for a 300V signal to appear suddenly at the input while the instrument is measuring a 50-mV input level. Not only must no damage occur, but also the overload recovery must be quick, and the input protection circuit must not be allowed to degrade the input noise floor when measuring 50 mV. The 300V signal may also have spikes or transients that far exceed 300V, or the user may inadvertently apply a larger signal. In all these cases the circuit must recover without causing a safety hazard to the user, destroying internal components, or even blowing a fuse.

Making the Most of a Microprocessor-Based Instrument Controller

by Corydon J. Boyan

In the 8903A Audio Analyzer, most of the tasks that lead to the display of measurements are coordinated by a microprocessor. The processor (an 8080 with 18K bytes of ROM and 192 bytes of RAM) counts and tunes the internal audio source, sets the input amplifier's gain, tunes the notch filter, sets the output amplifier's gain, and measures the voltages that will be used to generate each reading. This means that the 8903A can, among other things, automatically take distortion readings without calling upon the user to turn several knobs when seeking a null. This ability alone is ample justification for basing the instrument controller on a microprocessor, but the 8903A goes far beyond this in applying the power of its controller.

"Guaranteed" Accurate Measurements

The 8903A is HP-IB programmable, and this brought some important factors into consideration during the design. The performance of the internal source when settling from one frequency or level to another, for example, is a key factor in assuring the validity of the first measurement taken after changing the frequency or level. Similarly, the settling performance of the input and output amplifiers and the notch filter under various adverse signal conditions becomes very important when the instrument is trying to deliver an accurate first measurement after a change in operating parameters (e.g., after tuning to a new frequency). Add to this the desire to make measurements as rapidly as possible and you have a very interesting problem for a microprocessor to solve.

For example, every time the internal source is tuned, the processor spends about 170 milliseconds tuning to and verifying the frequency. This translates to over 86,000 operations. Of equal complexity is the job of setting the correct gain for the output amplifier and allowing the circuits to settle before the output amplifier voltage is read, with the object of never giving the user an invalid reading while at the same time delivering the reading in as short a time as possible. To accomplish this, the routine that controls the output amplifier makes use of such techniques as slope checking and frequency dependent delays to ensure rapid, valid readings. These techniques make possible accurate readings in half the time it might otherwise take. Referring to the flow chart, Fig. 1, note that the key to this routine is the technique of measuring the rate of change of the signal on the output amplifier (the slope) before attempting to check the signal level to determine if the output amplifier gain is properly set. This is done because it is quite common, after the signal level has suddenly changed (for example, when the notch filter is suddenly switched in

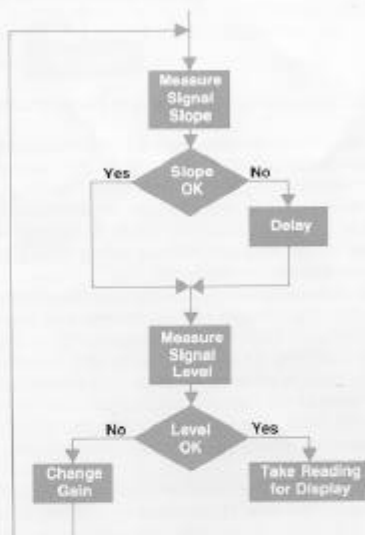


Fig. 1. 8903A leveling algorithm. The microprocessor checks the rate of change of the input signal as well as its level to make certain that the signal has settled before the output amplifier gain is set.

when going from ac level to distortion mode), to have a rapidly falling output amplifier voltage pass through the acceptable region, and thus have the gain appear to be properly set, when in fact the signal is on its way to a level that will require more gain for the proper reading. To keep the controller from being fooled by this phenomenon, we take two voltage readings in rapid succession, and from them calculate the rate of change of the signal. The time over which we measure this slope is longer at lower frequencies, and is in fact looked up in a table in ROM, based on the frequency at which we're operating. If the rate of change is too fast during this period, we delay before checking the level (see Fig. 2). This delay is also frequency dependent and table-driven. Note that if we simply delayed each cycle to keep from risking

Dc inputs and low-level differential amplification for common-mode rejection are necessary features for an instrument like the 8903A. One consequence of automatic ranging is that low-capacitance mechanical switching techniques cannot be used effectively. Needed are high-voltage reed relays, which affect high-frequency performance and require the use of compensation capacitors.

For dc operation the first part of the input signal path is dc coupled with the input blocking capacitor bypassed. The output of the differential-to-single-ended amplifier can then be monitored to obtain an accurately scaled representation of the input dc level.

Input voltages larger than 3V are attenuated by the input attenuator, a network of resistors that divide down the input signal. The appropriate tap point is selected by a reed relay network. If an overload occurs, the maximum attenuation setting is enabled.

To protect the sensitive input amplifier following the attenuator from short-term transients, an overvoltage protection network is used. For low-level signals the transfer impedance is low and signals applied to the input are coupled to the differential-input amplifier. However, for input signals large enough to damage the amplifier the output of the protection network is limited to a safe value.

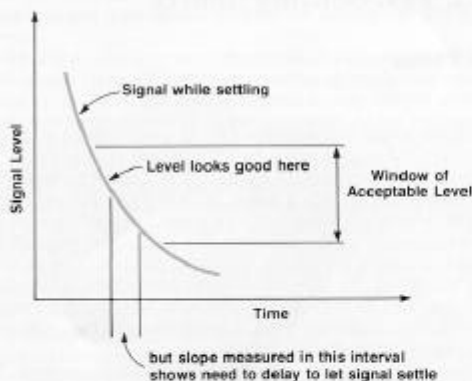


Fig. 2. An example of slope checking by the 8903A microprocessor.

displaying invalid readings, every measurement would take longer, perhaps by several hundred milliseconds. The slope checking gives us a rapid check on signal quality which eliminates the need to delay every time.

After the slope check (and possible delay) we assume the output amplifier is well settled and ready for leveling. We then measure the voltage at its output, and if too high or low, adjust the gain accordingly. After adjusting the gain, we make another pass through the leveling algorithm (unless we are now in the highest-gain range). To keep from getting caught in infinite loops, we require that after the first pass the gain never be reduced, and assume that if it is, we have an unstable signal, which causes us to start the entire measurement cycle (tune source, level input, tune notch, level output) over again. The final protection from infinite loops comes from counting the number of times we restart without displaying (each time we put a "----" pattern on the display), and after 128 times we display Error 31, which also goes out to the HP-IB.

Sweep

Because the 8903A has the ability to generate accurate and comparatively rapid measurements automatically, with no need for the user to insert delays in HP-IB routines or wait for the display to settle before taking a reading, it is possible to have the instrument sweep itself over a range of frequencies without user or computer intervention. Here the calculating power of the microprocessor is brought to bear on the problem of determining the frequency increment for each new point in the logarithmic sweep, based on the sweep range and the number of points per sweep (which can be set by the user). All the user need do to get a series of measurements spanning a frequency range is set the start and stop sweep frequencies and the number of

points per sweep, and press the SWEEP button. This makes it easy to measure the frequency response of an amplifier.

X-Y Recorder Output

The calculating power of the microprocessor-based controller is also apparent in the operation of the 8903A's X-Y recorder outputs. These outputs are driven by digital-to-analog converters controlled by the microprocessor, rather than directly from internal detectors. As a result, the user does not have to worry about the recorder output voltage abruptly changing when the analyzer autoranges. The microprocessor scales the recorder output according to the displayed reading and the plot limits entered via the keyboard. The recorder outputs are always between zero and ten volts so the recorder's zero and vernier controls need be adjusted only once. Thus, a properly scaled plot is easily generated by using the sweep and the X-Y outputs, without any need for an external controller.

Special Functions

Hidden behind the basic measurements are nearly forty special functions, which provide extended measurement capability and many service aids. For example, the analyzer can be given a load resistance in ohms and commanded to display ac level in watts. Another special function changes the number of points per decade in a sweep, and several special functions modify display operation. Normally the left and right displays indicate the frequency and level (or distortion, etc.) of the signal applied to the analyzer input. Sometimes, when using the 8903A just as a source, the user may want the analyzer to display the frequency and level of the source. Special function 10 provides this display.

Service aids provide front-panel display of many internal voltages and settings. Without microprocessor control each special function would require one or more switches on the front panel instead of one SPECIAL key, and would therefore probably not be included. Thus, the processor allows implementation of useful features the user would not otherwise get.

Corydon J. Boyan

Cory Boyan received his BSEE and MSEE degrees from Stanford University in 1974 and 1976. With HP since 1974, he's contributed to the design of the 436A Power Meter, the 8662A Synthesized Signal Generator, the 8901A Modulation Analyzer, and the 8903A Audio Analyzer. He's taught microprocessor design at Foothill College and served as chief engineer of Stanford FM station KZSU. Born in Boston, Massachusetts, Cory now lives in Mountain View, California. His interests include FM radio broadcasting, photography, backpacking, and the art of comedy—he's an avid fan of radio and television comedy groups.



The network consists of two back-to-back diodes which open up under large-signal conditions.

The differential amplifier consists of three high-performance amplifiers. These have the necessary noise, dc offset and frequency performance so that they do not degrade the signal quality. The total input amplifier chain acts as a 4-dB/step amplifier with leveling hysteresis, which maintains the post-amplifier signal level within 6 dB (3 to 1.5V rms). Should the output level change with time and deviate from this range, the gains of the attenuator, the differential-to-single-ended amplifier, and the programmable gain amplifier are adjusted to compensate.

To summarize, the gain of the input amplifiers is modified in three ways. First, the microprocessor monitors the input detector. If the detector voltage is too high or too low, the microprocessor varies the gain of the attenuator/amplifier chain to bring the level within bounds. Second, the overvoltage protection network limits if the input signal exceeds $\pm 15V$. Third, any sustained input overload trips the input overload detector. This detector monitors the input rms detector and the differential amplifier, and if either exceeds a certain absolute voltage limit, the overload detector trips, resetting the gain of the entire amplifier chain to its minimum value (maximum attenuation). This

Design for a Low-Distortion, Fast-Settling Source

by George D. Pontis

To fulfill the requirements of the 8903A Audio Analyzer, the built-in source must have good performance in certain key areas. First of all, for swept measurements, it must be readily programmable and fast settling. It must also have low distortion and noise, and it must have very good amplitude accuracy over the entire frequency range of 20 Hz to 100 kHz.

The combination of these conflicting requirements suggested the use of a high-performance RC oscillator instead of a synthesizer. The synthesizer is easy to program and settles quickly, but it is difficult to build a synthesizer with noise and distortion more than 80 dB below the fundamental. Synthesized designs also do not have sufficient absolute level accuracy or flatness without leveling, and a leveling loop that does not unduly degrade the distortion and settling time would be very difficult to design.

Unfortunately, none of the common RC audio oscillator designs looked suitable either. Usually the amount of distortion is inversely proportional to the settling time. Also, the tuning elements usually float, making the circuit difficult to interface with programming lines.

For these reasons a state-variable oscillator was chosen, similar to that proposed by Smith and Vannerson in 1975.¹ Since this oscillator is built around a state-variable filter structure, inexpensive JFET switches can be used easily to switch the tuning elements. More important is that the ALC design permits very rapid settling without trading off good distortion performance. Although the programmed frequency does not have the accuracy of a synthesizer, sufficient resolution is available to permit firmware tuning to within $\pm 0.3\%$ of the programmed value.

This oscillator design can be described as a state-variable filter in

which the resistor that determines the Q is replaced by an analog multiplier. The control signal for the multiplier is provided by an automatic level control (ALC) circuit. The ALC circuit compares the oscillator amplitude with a stable dc reference obtained from a temperature-compensated reference diode. The resulting error signal is processed through two paths. One path carries the cycle-by-cycle proportional error to the controlling multiplier. This provides very fast settling. The other path includes an integrator in the loop to eliminate nearly all of the steady-state error. This design is theoretically capable of settling the output amplitude within two cycles after small-signal amplitude disturbances.

There are two important refinements in the 8903A oscillator. The first is the use of a special two-stage peak detector. This consists of track/hold and sample/hold amplifiers to eliminate any distortion-causing ripple on the detected peak output. The second refinement is the addition of an ALC loop gain control to compensate the leveling loop, cycle by cycle, for changes in oscillator amplitude. This greatly decreases the large-signal settling time of the oscillator, which is important when switching from one range to another.

Fig. 1 shows the oscillator integrators. The gain constant of these integrators is changed in three-octave steps by selecting the feedback capacitor. This gives us range switching. Coarse tuning within each range is done by switching the input resistors. The eight binary-weighted resistors provide 255 usable steps. Placing the switching devices at the virtual ground point permits the use of JFETs with low drain-to-source on resistance ($R_{DS(on)}$). An individual transistor switch conducts when its gate is allowed to rise to ground potential, and turns off when its gate is pulled to the negative supply, $-15V$.

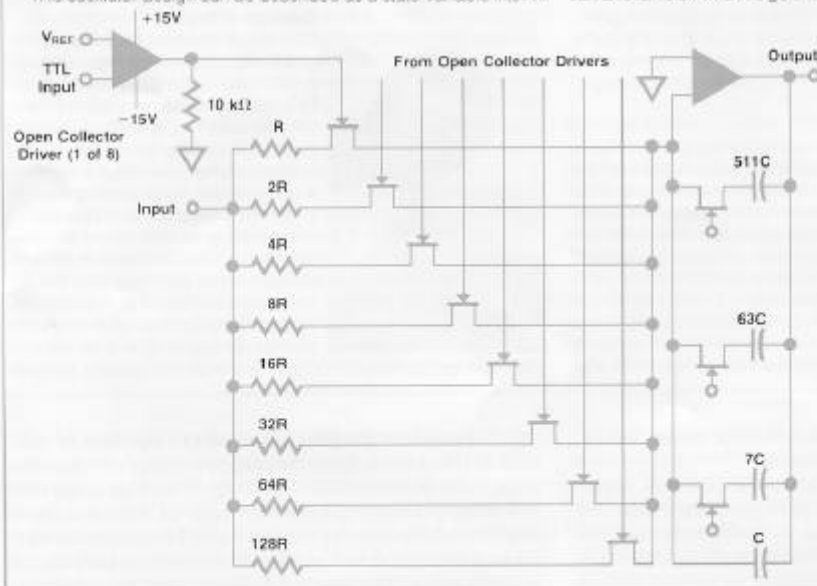


Fig. 1. The gain-switched integrator used in the 8903A's internal oscillator. Capacitors are switched to change ranges. Resistor switching provides coarse tuning within each range.

reset occurs within ten milliseconds if the overload is severe. This protects the input from burning out or blowing a fuse, and allows for rapid overload recovery.

Notch Filter

The notch filter design challenge was twofold. First, it was necessary to design a low-distortion, low-noise filter that was also programmable. Second, to minimize overall

The switching scheme is economical because multiple resistor packages and quad comparators can be used to interface to TTL levels.

Fig. 2 is a block diagram of the oscillator. Integrators U1 and U2 and inverter U4 form the state-variable filter structure. Fine tuning is done by U4. This stage uses a switched resistor network similar to that used for coarse tuning in the integrators. As the resistors are switched the gain of U4 changes, effectively altering the amount of signal transmitted from the output of U2. The gain is proportional to $\sqrt{A-B/R}$, where R is the parallel combination of the selected input resistors and A and B are constants that provide a $\pm 5\%$ fine tuning range. This gives the oscillator enough resolution to tune within $\pm 0.2\%$ of any frequency within the range of the instrument.

It is generally true for sinusoidal oscillators that purity and settling time are chiefly functions of the ALC circuits or mechanisms used. The two-stage peak-detector circuits are the key to the performance of the 8903A oscillator. Oscillator amplitude data is obtained by the track/hold and sample/hold amplifiers in the following manner. Switch S1 closes during the time the output is at its negative peak. Capacitor C1 rapidly charges, following the sine-wave amplitude up to its positive peak. At this time S1 opens, holding the peak voltage on C1. Switch S2 then quickly closes and opens again, thus updating the sampled peak level held by capacitor C2.

The two-stage scheme has several advantages. For one, the first stage may be optimized for fast data acquisition, while the second stage is optimized for long hold time. Fast acquisition is essential for

good amplitude accuracy at high frequencies. Low droop is important to maintain low distortion at low frequencies. Also, this scheme has no steady-state ripple in the sampled output which would cause distortion to appear on the multiplier output and in turn on the main oscillator output.

In practice, the two-stage scheme is also easier to implement than a very fast single-stage sample/hold circuit. The required circuit functions are accomplished with simple JFET switches and unity-gain buffers as shown in Fig. 2.

For sampled data systems in general, settling time is strongly dependent upon loop gain. For this circuit the ideal integrator gain is linearly proportional to frequency. This effect is achieved by switching the integrator resistor on once per cycle for a duration of $35 \mu\text{s}$. This process increases the integral error signal with frequency. The duty cycle increases with frequency until the integrator switch S3 is left closed continuously for frequencies greater than 25 kHz. Above 25 kHz, the oscillator easily settles in less than one millisecond.

8903A oscillator performance is largely limited by the quality of commercially available, reasonably priced analog multipliers. The multiplier used is decoupled slightly (from optimum) to reduce its contribution to THD and noise. This extends the oscillator settling time to a period of four to five cycles.

Reference

1. E. Vannerson and K. Smith, "A Low-Distortion Oscillator with Fast Amplitude Stabilization," *International Journal of Electronics*, Vol. 39, No. 4, 1975, pp. 465-472.

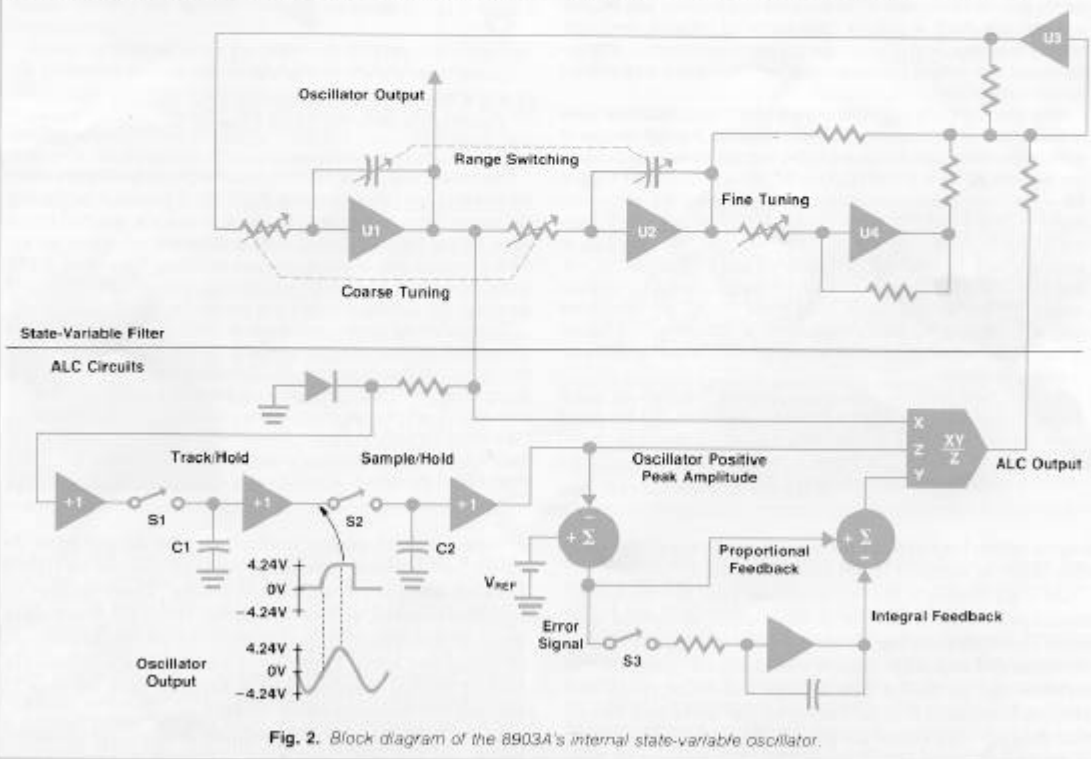


Fig. 2. Block diagram of the 8903A's internal state-variable oscillator.

measurement time, it was necessary to develop an accurate, rapid fine-tune mechanism with quick recovery from overloads and mistuning. To achieve good Q over the frequency range, an active RC filter is necessary. To tune this device, a

variable resistance or conductance device is needed, since achieving the tuning range with variable capacitors or inductors is impractical. Many resistively tunable active configurations are feasible. However, determining the op-

Floating a Source Output

by George D. Pontis

To provide the greatest versatility in both benchtop and systems applications, the 8903A Audio Analyzer's built-in source is floating. This lets the user eliminate ground-loop errors, sum signals, and add dc offsets to the source output.

Previous designs used a separate, isolated power supply for the source circuits. This method is straightforward and offers very good low-frequency common-mode rejection. However, there are several reasons why this arrangement is not used in the 8903A.

The biggest problem with the floating power supply approach is interfacing with the digital programming lines. Since only three of the thirteen attenuator lines have relay isolation and none of the nineteen oscillator lines can be floated, over thirty lines must be coupled in some manner. One solution is to float only the final output stage. This eliminates the need for couplers, but requires a high-performance differential-input amplifier to reject the common-mode signal that appears at the input of the floating stage. Since a floating power supply is still required, the cost of this approach is relatively high.

The 8903A solves this problem with a single-ended-to-differential output converter. This circuit, shown in Fig. 1, operates on the instrument's ground-referenced $\pm 15V$ supplies and requires only two operational amplifiers. A precise combination of negative feedback, positive feedback, and cross-coupling yields a symmetrical differential output with infinite common-mode rejection and a well-defined output impedance.

An analysis of this circuit is generally a tedious procedure because of the number of components involved. However, the high degree of symmetry in the circuit can be exploited to great advantage by using the relations $R2/R1 = R12/R11$, $R3 = R7$, $R6 = R10$, and $R4 = R5 = R8 = R9$. From these relationships, one can derive the expression $R2/R1 = (2R6 + R4)/2R3$, which is a necessary condition for achieving an infinite common-mode output impedance. Then it is easy to calculate the differential output impedance and the open-circuit voltage gain. The resulting equations can be manipulated to find suitable values. For the resistor values used in the 8903A, the associated gain is 1.125 and the output impedance is 480 ohms. The output is further padded with a 120-ohm resistor to yield the desired 600-ohm output impedance.

It would have been possible to use resistors that gave an output impedance of exactly 600 ohms instead of 480 ohms, but this would have required setting up and stocking a supply of several extra odd resistor values. As it is, the circuit is realized using 0.1%, 25-ppm resistors that are also used elsewhere in the instrument.

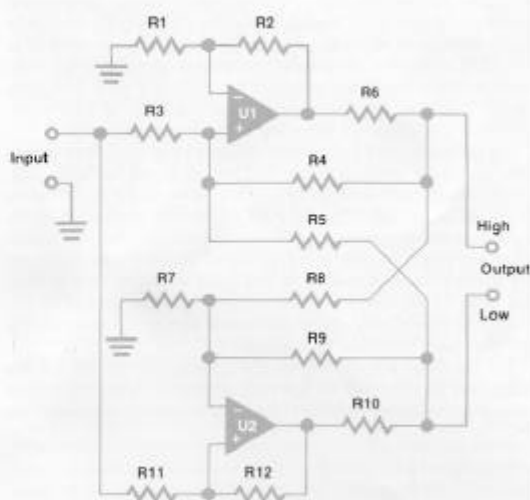


Fig. 1. Single-ended-to-differential output converter provides a floating output for the 8903A's internal oscillator.

The easiest way to see how this circuit works is to eliminate either the inverting (low) or noninverting (high) half of the circuit by shunting the respective output to ground. Fig. 2 shows the reduced circuit when the low half is grounded. If R4 is disconnected, the circuit will have a forward gain of about one, and an output impedance of 276 ohms. R4 works in conjunction with R6 to provide voltage and current feedback that causes the gain and the output impedance to rise.

To demonstrate that the output is truly floating, we ground the input and apply a test source to both outputs. Ideally, the current flow from the test source should be zero. Fig. 3 shows a block diagram and the reduced circuit for this test. Here it can be quickly calculated that the output of U1 will rise just enough over that of the test source to make the current through R6 cancel the current through R4 and R8. Note that the current flowing through sources V1 and V2 is supplied by the other half of the circuit, which is not shown.

timum match between the filter configuration and the variable resistive element is not straightforward.

Let's go through the alternatives and the tradeoffs. Switchable resistor networks have good distortion and noise characteristics but do not provide continuous tuning coverage and require extensive switching circuitry. Photoresistors can be driven over a large resistance range and provide continuous tuning. However, the noise and distortion they add to a signal are greater than the required level of 90 dB below the signal level. They can be used as fine-tuning elements if coupled only partially into the circuit. These devices can also be slow and are awkward to control rapidly, reducing the tuning speed. Finally, they tend to vary significantly from device to device and with time and temperature, making compensation difficult.

Four-quadrant analog multipliers also do not have the 90-dB performance necessary, but they too can be lightly coupled into the circuit for fine tuning. These devices are fast, inexpensive, and easy to drive. There are many variations on this type of circuit, some of which can be obtained in integrated form. Those most suitable use a differential pair of bipolar transistors as a variable gain element by varying the common-mode current.

Light bulbs as variable resistive devices are relatively linear but are slow and have a limited dynamic range. Thermistors, diodes and other nonlinear devices would all be useful only for fine-tune applications. The drive and compensation circuitry for all of these alternatives would be complex and the overall performance marginal.

The tuning elements selected were switchable resistor

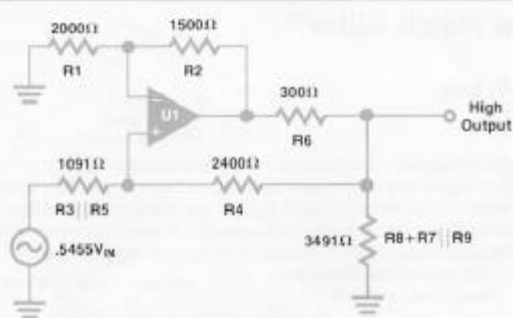
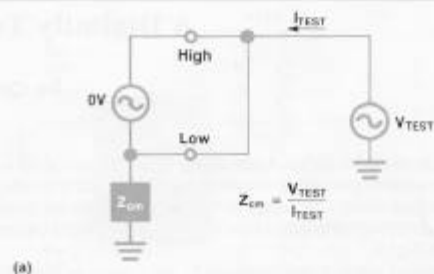


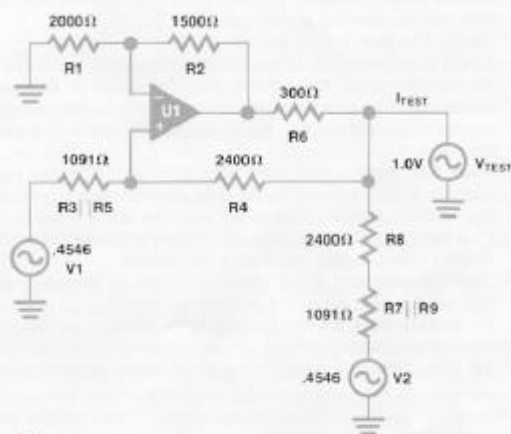
Fig. 2. Eliminating the inverting side of the circuit of Fig. 1 by shorting the low output to ground results in this reduced circuit.

In practice it was found that parasitic effects and the electromagnetic-interference (EM) filters degrade the circuit balance when the frequency approaches 100 kHz. However, each board is tested for a minimum of 50 dB common-mode rejection at 1 kHz. A typical unit has greater than 40 dB rejection at 100 kHz. Also, a 10-k Ω resistor internally ties the low output to the chassis ground. This provides a reference for the output when no external load is connected. Without this resistor, the common-mode output voltage is indeterminate.

One initial concern about the circuit was difficulty of troubleshooting problems; such as one of the resistors drifting out of tolerance, causing poor common-mode rejection. This problem was solved by implementing the following test procedure. First, the input to the floating amplifier is set to exactly 1.00V rms using the special functions built into the 8903A. Then, the technician shorts the low side to



(a)



(b)

Fig. 3. (a) To demonstrate that the output of the circuit is floating, the input is grounded and a voltage source is applied to both outputs. (b) Reduced circuit for this test (only half of the circuit is shown). Current through R6 cancels the current through R4 and R8. Current through V1 and V2 is supplied by the other half of the circuit (not shown). Thus no current is drawn from the test source.

ground and measures the potentials at several circuit nodes. The measured values can be compared to the calculated values published in a table in the 8903A manual. A second set of measurements can be made, if necessary, with the high side grounded. By observing the deviations between the measured and calculated values, it is easy to locate the faulty component.

George D. Pontis

Born in Los Angeles, California, George Pontis attended the University of California at Santa Barbara, receiving his BSEE degree in 1975. He joined HP in 1978 and contributed to the design of the 8903A Audio Analyzer. Formerly a contributing editor of Audio Magazine, he's a member of the Audio Engineering Society. He lives in Palo Alto, California and enjoys outdoor sports, especially tennis, skiing, and swimming.



networks for low-distortion coarse tuning, and a four-quadrant multiplier for fine tuning between the discrete steps of the switchable resistors. Since the four-quadrant multiplier is coupled into the circuit only enough for $\pm 7\%$ tuning, it does not contribute significantly to the overall noise and distortion. A resistor switching network is best implemented if one end of the network is dynamically kept at ground potential; this relieves many constraints on the switching network. To this end, a state-variable* notch configuration is used (see page 14). With this design, low-distortion tuning over a 10:1 frequency range is achieved.

*The name "state-variable filter" is a consequence of the fact that the equations that describe the system can be written in a form that fits the classical state-space description of a linear system, i.e., $\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu}$, where \mathbf{x} is the response (or state variable) of the system to an input u .

Capacitors are switched into the network to change frequency in three-octave bands and provide complete coverage of the frequency range of the analyzer.

To complete the fine-tuning path, a synchronous detector mixes the filter output waveform with the fundamental waveform. If any fundamental exists in the notch output, a dc current is generated to fine tune the notch. The critical parameters here are 100-dB dynamic range and rapid operation. A FET double-balanced mixer was selected. The input mixing signal that drives the FET is a square wave, rich in odd harmonics, so the circuit responds to odd-order harmonics as well as to the fundamental. This is the classical solution. A complete null may not be achieved if third, fifth, or higher-order odd harmonics are present. Total error can

A Digitally Tuned Notch Filter

by Chung Y. Lau

The notch filter in the 8903A Audio Analyzer rejects the fundamental frequency component of the incoming signal. This filter consists of a state-variable active filter and fine-tune and fine-balance control circuits. A simplified schematic of the filter minus the control circuitry is shown in Fig. 1.

Conventional notch filters (bridged-T, etc.) used in distortion analyzers are not well suited for digital control because they require expensive relays or analog switches. The state-variable filter approach has the following advantages:

- The Q of the filter is fixed and independent of frequency.
- Tuning is accomplished by switching resistors and capacitors in the two integrators. Inexpensive JFETs are used as switching elements and the switch drivers are simple because both the gate and the channel of the JFET are virtually at ground potential when the switch is on.
- Three filter outputs (low-pass, bandpass, and high-pass) are available simultaneously. The low-pass and high-pass outputs are phase-shifted from the input by 90° when the filter is tuned to the input frequency. These two outputs are used in the control circuits, so extra phase-shifting circuits are not needed.
- Distortion generated in the fine-tune circuitry is filtered by two integrators before appearing at the notch output.

In distortion mode, the filter is tuned in the following manner. The microprocessor counts the input frequency and tunes the notch filter to the same frequency by switching in the proper capacitors and

resistors. When the filter is tuned to the fundamental frequency of the input signal, the bandpass output (V_{BP} in Fig. 1) inverts the fundamental component of the input signal. U5 sums V_{IN} and V_{BP} to cancel the fundamental component of V_{IN} exactly. However, because of the phase and amplitude characteristics of V_{BP} , the harmonics present in V_{IN} are relatively unattenuated.

Analytically, we have

$$\frac{V_{BP}}{V_{IN}}(s) = \frac{-s\omega_c/Q}{s^2 + s\omega_c/Q + \omega_c^2}$$

and

$$\frac{V_{BP} + V_{IN}}{V_{IN}}(s) = \frac{s^2 + \omega_c^2}{s^2 + s\omega_c/Q + \omega_c^2}$$

where ω_c is the center frequency in rad/s. Therefore, there is zero transmission at $s = \pm j\omega_c$.

The state-variable filter alone can provide only about 15 dB of fundamental rejection because of the coarseness of the frequency tuning (use of discrete resistor values, resistor mismatches, etc.). Fine-tune and fine-balance control circuits are necessary to achieve a notch depth greater than 90 dB. The fine-tune circuit insures that the filter is exactly tuned to the input frequency (no phase error) and the

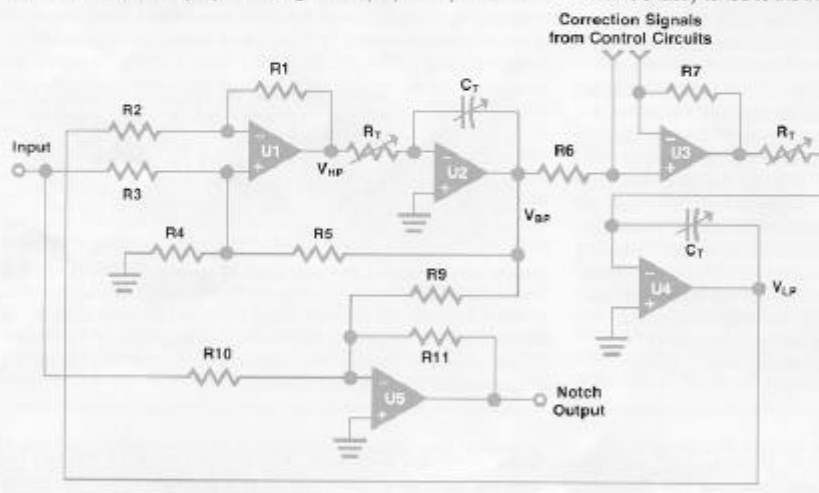


Fig. 1. 8903A notch filter. Correction signals from the control circuits of Fig. 2 provide fine tuning and fine balance.

be as much as 0.46 dB, but this is considered reasonable.

Rapid response is achieved in two ways. First, when an overload occurs, the circuit gain is quickly reduced by the output amplifier circuit, minimizing the transient and thus minimizing the recovery time of the synchronous detector from the overload. Second, the response time of each frequency band has been optimized. Theoretically, it is impossible to fine tune a filter as rapidly at 20 Hz as at 1 kHz. It simply takes a much longer time to detect a null at 20 Hz, but even more important is that the time response of a 20-Hz notch filter is much greater. In fact, a fundamental burst

tone applied at the input of the notch will not immediately be nulled at the output. The entire input will first appear at the output and then decay away in proportion to $(\text{notch}) \cdot Q$ where Q is the Q of the notch. Thus, as the notch frequency is increased, so is the speed of the fine-tuning circuitry. The bottom frequency band (20-200 Hz) is relatively slow but nulls optimally at 20 Hz. For higher frequencies the instrument response time improves.

Output Amplifier

For the amplifier following the notch filter, gain accuracy

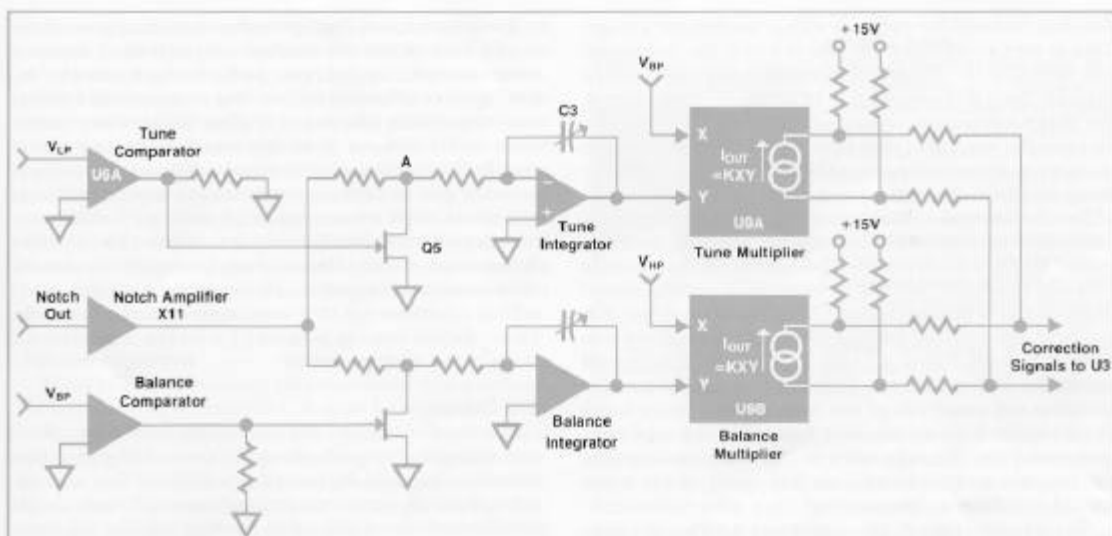


Fig. 2. 8903A notch filter fine tuning and balance circuits.

fine-balance circuit insures that $|V_{BP}/V_{IN}| = 1$ at the input frequency (no amplitude error). The fine-tune circuit is described in some detail here; the fine-balance circuit operates in a similar manner. Fig. 2 is a simplified schematic of both circuits.

The fine-tune circuit operates as follows (see Fig. 2). The low-pass filter voltage V_{LP} , which is phase-shifted 90° from the input, drives the tune comparator U6A, which turns JFET switch Q5 on and off. When Q5 is on, point A is essentially grounded and no current flows into the integrator capacitor C3. When Q5 is off, the notch amplifier feeds the tune integrator. Because of this chopping action, the dc current that flows into the integrator is caused by the notch output component that is synchronous with V_{LP} .

The tune integrator output is a dc voltage that can be changed only by the dc current flowing into the integrator. This voltage feeds one input of the multiplier U9A. The other input to the multiplier is V_{BP} . The product of the two inputs is summed into the state-variable filter via U3 (see Fig. 1). The net result is that the tune integrator voltage can change the effective resistance (R_f in Fig. 1) and hence the center frequency of the notch filter.

The direction of change is such that any notch output components in synchronism with V_{LP} are reduced. At steady state, the

tune voltage is stable, although not necessarily zero, and no dc current flows into the tune integrator. Therefore, there is no notch output component synchronous with V_{LP} .

The fine-balance circuit insures that there are no components synchronous with V_{BP} in the notch output. V_{BP} and V_{LP} are both at the fundamental input frequency and are in quadrature with each other. Thus, at steady state, there is no fundamental frequency in the notch output.

*Actually, third harmonics in the notch output can also cause dc current to flow in the integrator. This error causes a small amount of the fundamental frequency component to pass through the notch filter. However, the maximum error contribution to distortion measurements from this effect is only about 0.46 dB.



Chung Y. Lau

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and good frequency response are important. In ac level measurements the signal travels through the output amplifier chain and is detected by the output rms detector. Any error degrades the 8903A's performance. The output amplifier allows the low-level distortion products leaving the notch to be accurately detected by limiting the required dynamic range of the rms-to-dc converter. The rms detector is accurate over only a 30-dB range, and the output amplifier boosts these signals into the range of the detector. As with the input circuitry, rapid recovery from overload conditions is crucial. If the notch becomes mistuned because of a disturbance at the input, the output suddenly

increases dramatically, sending the amplifier into overload. This in turn generates dc offsets in the amplifier chain that can take seconds to decay, even if the amplifier may start operating again much sooner. Thus a large low-frequency impulse appears at the amplifier output along with the signal being amplified. The composite signal is transferred to the output rms detector, which responds to a time-weighted average of the total. If the impulse is a significant fraction of the signal, the rms detector will not give a true indication of the signal amplitude. This can cause problems when the instrument ranges automatically. In effect, it forces the leveling algorithm to wait much longer to con-

firm that the amplifier output is within the detector's range. This in turn slows down the rate at which the instrument can determine the proper measurement range and display a reading. Even if the impulse is 20 dB less than the signal, the detector error can be as much as 0.5%. This will also increase the amount of time required by the instrument to make an accurate reading from the detector once the proper range has been obtained.

To alleviate these problems, the size of the transient is minimized in three ways. First, operational amplifiers and circuit configurations are used that have better than average overload immunity. Second, an overload detector is placed at the output of the rms detector. If the signal level becomes too large, the overload detector trips and the amplifier gain is reset to 0 dB. Third, a 13-Hz high-pass filter is placed before the output detector. This significantly reduces the duration and amplitude of any transient and hence keeps the transient from significantly increasing the total measurement time. The only delay factors that remain are the controllable and predictable settling times of the notch circuit and the rms detector.

The response time of the output rms detector is a compromise between rapid settling and low-frequency accuracy. A configuration was selected that settles to within 1% of a 10:1 step in 350 milliseconds, and has a steady-state error of 0.2% at 20 Hz. This includes filtering in the detector and additional filtering following the detector to reduce excess ripple. For leveling purposes the ripple is not significant, so the microprocessor uses the detector output when leveling and avoids the extra delay contributed by the additional filter. The output detector and filters could have been designed with switchable time constants to respond more rapidly for higher-frequency signals. However, the penalties would have been additional circuit complexity and the ambiguity of not knowing when to invoke the longer time constant. A 20-kHz signal, for example, might still have a significant low-frequency component, which would cause excessive error with a more rapid time constant.

Oscillator

Many of the design considerations for the notch filter apply equally to the oscillator. In both cases tuning considerations were the same, with switchable resistor networks used as the decade tuning elements and four-quadrant multipliers used for amplitude control. In many ways the oscillator and notch circuits can be seen as duals. The notch generates a pair of zeros on the $j\omega$ axis that reject the fundamental component, while the oscillator generates a pair of poles on the $j\omega$ axis that generate sustained oscillations. The trick in the oscillator is to keep the poles exactly on the axis to maintain constant output amplitude. This must be done continuously by the automatic leveling circuit. If the frequency of the circuit deviates from the desired frequency, the circuit can be fine tuned by the microprocessor, which monitors the output frequency on a sampled basis. The major performance goals of the oscillator were low noise and distortion, rapid amplitude and frequency settling, and digitally programmable frequency control. Again the state-variable filter configuration along with a special leveling circuit offered the flexibility and performance required. The oscillator design is described on page 10.

It was determined during development that the oscillator would have to run at a constant output level to maintain reasonable settling and noise performance. It was also desired to have a floating output. The attenuator and output amplifier circuit (see page 12) takes the oscillator output level and translates it to the selected floating output amplitude. To minimize cost and achieve overall output accuracy goals the attenuation is done in two stages. Coarse amplitude steps are implemented with a 2.5-dB/step attenuator network. Smaller steps are provided by a resistive ladder network that adjusts the amplitude linearly in small discrete steps. The combination can adjust the amplitude within a nominal $\pm 0.15\%$ worst case. Computation of the proper switch settings is an easy job for the computational skills of the microprocessor.

EMI Design

Meeting the required electromagnetic interference (EMI) and susceptibility goals was a bit more challenging than initially expected. Large-amplitude RF fields tend to generate voltages on exposed cabling and circuits. These voltages overdrive many of the active circuits, causing nonlinear operation and distortion. To avoid direct exposure to these fields, the analog circuits are housed in an internal EMI-tight box. The box has an aluminum frame around the sides. The bottom cover is the ground plane of a printed circuit board and the top cover is a removable EMI-tight lid. Removal of the lid, which is held in place by only two screws, makes all the circuits available for service. The microprocessor boards are sufficiently shielded by the instrument cabinet and do not require the extra shielding. To keep the RF fields from developing voltages on the cabling feeding the circuits, special precautions were taken. First, from the inner box to the front panel, shielded cable is used. Second, BNC connectors are provided on the front panel. The BNC connectors allow the attachment of shielded cables directly to the instrument if desired, thus preventing EMI pickup. The instrument's digital circuitry also generates EMI related to harmonics of its 2-MHz clock. This problem was minimized by means of RF gaskets on some of the cabinet seams and by installing an EMI suppressing filter on the power line input. As a result, the instrument will not disturb sensitive receivers operating nearby, and yet will perform well near a powerful transmitter.

Frequency Measurement

A key feature of the 8903A is its ability to measure frequency automatically, even when the input waveform may have a significant amount of noise and distortion and the amplitude may vary from 6 mV to 300V. Part of this problem is solved because the instrument is autoranging and keeps the leveled waveform within 6 dB over most of the input amplitude range. But before the signal can be accurately counted it must first be converted into a binary signal having the same period as the major frequency component in the waveform, and herein lies a problem. If a zero-crossing circuit is used, noise may cause multiple crossings and a false indication of the frequency. Hysteresis in the detector will help, but if the hysteresis is too large, smaller-amplitude waveforms may not trigger the detector at all while large-amplitude waveforms will have relatively little

hysteresis protection when large noise components are present. To alleviate this problem, the 8903A employs variable hysteresis. As the peak amplitude of the signal varies, so does the hysteresis level, which is maintained at approximately one-half the positive peak for the positive portion of the waveform and approximately one-half the negative peak for the negative portion of the waveform. Hence noise immunity remains constant regardless of the incoming waveform. Hysteresis is implemented with a bipolar peak detector followed by a dual comparator. The waveform is transferred to a reciprocal counter, which measures the period of the signal, and the microprocessor inverts this period to get frequency.

Acknowledgments

Many people throughout HP contributed to the successful introduction of the 8903A. It must be stressed that the instrument's success is the result of the total contribution of many people from early investigation through production. So first and foremost may I thank all of those who contributed their time, enthusiasm, and support. On the R&D team I would like to thank Allen Edwards for contributing to the original project concept and leading the project through early development. Chung Lau for his overall technical support and especially for his efforts in developing the notch and input circuits and verifying overall instrument performance. Cory Boyan for the initial oscillator circuit investigation, software development and coordination, and digital circuit development, Bob DeVries for product design, Derrick Kikuchi for overall software development and latch board development, George Pontis for developing the oscillator, attenuator, output amplifier and power supply circuits as well as various special test fixtures to verify

James D. Foote

A native of Madison, Wisconsin, Jim Foote earned his BSEE degree at the University of Wisconsin in 1972 and joined HP in 1973. In 1975 he obtained his MSEE at Stanford University. He has served as a design engineer on both the 8901A Modulation Analyzer and the 8903A Audio Analyzer, and as project manager for the 8903A. Jim has just joined HP's Disc Memory Division and has moved to Boise, Idaho with his wife and daughter. Among his interests are reading, walking, racquetball, skiing, chess, and doing odd jobs around the house.



instrument performance, Peter Lee and Jim Stewart for industrial design, and Bruce Creedy for initial product design. Special thanks also to Ray Shannon and Jim Stinehelfer who were instrumental in the early product definition. Other key contributors include Rick Pinger and Jim Harmon in providing service and operating documentation, Bob Stern and Bob Rands in product marketing support, Bob Cirner and Ken McFarland in parts scheduling and procurement, Greg Hoberg, Bob Shatara, Phillis Nakano, Dana Kreitter, and Rich Mills for production support, and Charlie Sallberg and Chuck Clavell for reliability engineering test and support.

Reference

1. A.P. Edwards, "Precise, Convenient Analysis of Modulated Signals," Hewlett-Packard Journal, November 1979.

SPECIFICATIONS

HP Model 8903A Audio Analyzer

SOURCE

FREQUENCY RANGE: 30 Hz to 100 kHz
 FREQUENCY RESOLUTION: 0.5%
 FREQUENCY ACCURACY: $\pm 0.01\%$ (500 Hz)
 OUTPUT LEVEL RANGE: 0.05 mV to 6V open circuit
 OUTPUT LEVEL RESOLUTION: 0.5% on meter
 OUTPUT LEVEL ACCURACY: open circuit: 2% of reading
 80 mV to 4 V: $\pm 0.1\%$ to $\pm 0.5\%$, 5V: $\pm 0.1\%$ to $\pm 0.5\%$, 6V: $\pm 0.1\%$ to $\pm 0.5\%$
 20 mV to 6 mV: $\pm 0.5\%$ to $\pm 1.0\%$ (typical)
 FLATNESS: $\pm 1\%$ (typical) $\pm 0.5\%$ to $\pm 1.0\%$ (typical)
 DISTORTION & NOISE: The higher of ± 80 dB or 30 Hz to 20 kHz, 30 mV RMS
 $\pm 0.02\%$ to $\pm 0.1\%$ (typical) 20 kHz to 100 kHz, 100 mV RMS
 100 kHz RMS
 IMPEDANCE: 600 Ω - 1 Ω
 SWEPT MODE: Logarithmic sweep with up to 500 points/decade or 200 points
 between selected start and stop frequencies, whichever is greater.

AC LEVEL

FULL-RANGE DISPLAY: 300 mV, 300 mV, 3.000 V, 30.00 V, 300.0 mV, 3.000 mV
 OVERRANGE: 3% except on 300 V range
 ACCURACY: $\pm 0.5\%$ of reading 30 V to 300 V, 30 Hz to 100 Hz, $\pm 0.5\%$ of reading, 50 mV to 300 V, 20 Hz to 20 kHz, $\pm 0.5\%$ of reading, 0.5 mV to 300 V, 20 Hz to 100 kHz

AC CONVERTER: True rms measurement capability (standard) or average to rms converter (optional) $\pm 0.5\%$ (typical) $\pm 0.5\%$ (typical)

DC LEVEL

FULL-RANGE DISPLAY: 300 mV, 40.0 mV, 10.00 V, 100.0 V
 OVERRANGE: 3% except on 100 V range
 ACCURACY: $\pm 0.7\%$ of reading 100 mV to 300 V, $\pm 0.5\%$ (typical)

NOISE

FUNDAMENTAL FREQUENCY RANGE: 20 Hz to 100 kHz
 DISPLAY RANGE: 0 dB to 90 dB (dB)
 ACCURACY: ± 0.5 dB to ± 2 dB, ± 0.5 dB to ± 100 dB
 INPUT VOLTAGE RANGE: 20 mV to 100 mV
 DETECTION: True rms (average detection) or peak (peak-to-peak)

RESOLUTION: 0.01 dB for 50 dB to 100 dB, 0.1 dB below -20 dB the display is measured to the nearest 0.1 dB to indicate 0.1 dB of noise signals. Full resolution is available by disabling this feature using special function 15.1.

ANALOG METER: Active in SMD mode and for SMD values ± 10 dB on 20 dB using meter function 7.1
 SWEPT mode: Meter is locked to analyzer source frequency.
 DISPOSITION
 FUNDAMENTAL FREQUENCY RANGE: 20 Hz to 100 kHz
 DISPLAY RANGE: 0.01 Hz to 100 Hz, 0.01 Hz to 100 Hz (typical)
 ACCURACY: ± 1 dB, 20 Hz to 20 kHz, ± 0.5 dB, 20 Hz to 100 kHz
 DISPLAY RESOLUTION: 0.001 Hz, ± 0.1 Hz, 0.1 Hz to 100 kHz
 INPUT VOLTAGE RANGE: 20 mV to 100 mV
 REGIONAL NOISE AND DISTORTION: The higher of 0.01% to 30 dB or 20 Hz to 20 kHz, 80 mV RMS, 0.02% to 0.1% (typical) 20 kHz to 100 kHz, 100 mV RMS
 0.02% to 0.1% (typical) 20 kHz to 100 kHz, 100 mV RMS
 DISPLAY RESOLUTION: 0.001%, ± 0.1 Hz, 0.1 Hz to 100 kHz
 0.01%, 2% to 30% (typical), 0.1%, ± 0.1 Hz, 0.1 Hz to 100 kHz
 DETECTION: True rms

SIGNAL TO NOISE

FREQUENCY RANGE: 20 Hz to 100 kHz
 DISPLAY RANGE: 0 dB to 20 dB (dB)
 ACCURACY: ± 1 dB
 INPUT VOLTAGE RANGE: 20 mV to 100 mV
 REGIONAL NOISE: The higher of ± 80 dB or 20 Hz to 100 kHz, 100 mV RMS
 100 mV RMS
 RESOLUTION: Same as SMD
 DISPOSITION: The analyzer displays the ratio of the input voltage to the selected noise to automatically calculate on read out

FREQUENCY COUNTER

RANGE: 20 Hz to 100 kHz
 RESOLUTION: 0.5 Hz
 ACCURACY: $\pm 0.5\%$ to $\pm 1\%$
 INPUT SENSITIVITY: 20 mV in standard and SMD modes, 5.0 mV in ac line and trigger modes
 COMPARED TECHNIQUE: Response with 2 MHz reference

ANALOG FILTERS

40 Hz HIGH PASS FILTER: 20 dB/oct frequency 40 Hz to 40 Hz; Roll-off 40 dB/decade
 PROPHETIC FILTERS: COITT (Compensation) Filter, Deviation from Ideal
 Response: ± 0.5 dB at 100 Hz, ± 0.5 dB at 100 Hz; ± 0.5 dB at 100 Hz, ± 0.5 dB at 100 Hz
 35 Hz to 5 kHz
 30 kHz LOW PASS FILTER: 20 dB/oct frequency 20 kHz to 20 kHz; Roll-off 40 dB/decade
 30 kHz LOW PASS FILTER: 20 dB/oct frequency 20 kHz to 20 kHz; Roll-off 40 dB/decade

REAR PANEL INPUTS/OUTPUTS

RECEIVER OUTPUT: 0 to 100 Hz typical corresponding to top of available frequency
 0 to 100 Hz typical corresponding to top of available frequency
 RECEPTOR OUTPUT RESISTANCE: 1 k Ω nominal
 MODULATOR OUTPUT: In an SMD mode provides scaled output of measured input signal
 in SMD, detector, and detector level modes provides scaled output of input signal
 with the fundamental removed

General

INPUT IMPEDANCE: 1000 Ω to 1 Ω (switchable by ± 100 pF with low terminal grounded)
 CMRR: 20 Hz to 40 dB for differential input, 20 Hz to 40 dB for differential input ± 100
 10 dB for differential input ± 100
 REMOTE OPERATION: HP-IB, all functions except the low carrier, low terminal ground
 switches, and the ± 10 to 100 pF terminal legs
 TEMPERATURE: Operation 0 to 50°C, storage -10 to 70°C
 POWER REQUIREMENTS: 100, 120, 220 or 240 VAC $\pm 5\%$, 1.5, 1.5, 4.0 or 4.0 VA
 WEIGHT: See 103 mg 103 mg
 SIZE: 148 H x 420 W x 148 D mm (5.8 x 16.5 x 5.8 in)
 EMI: Conducted and radiated emissions in accordance with requirements of methods
 C100 and R102 or MIL-STD 461A, VCE 401, and CISPR publication 11. Conducted
 and radiated susceptibility meet the requirements of methods C101, C102, and R103.
 (For information on MIL-STD-461A, see MIL-STD-461A)

PRICES IN U.S.A.

8903A Audio Analyzer SMD
 Option 011: Rear panel connectors instead of front panel for remote output and input
 type input, 800
 MANUFACTURING DIVISION: STAMFORD PARK DIVISION
 1300 Page Mill Road
 Palo Alto, California 94304 U.S.A.

1. SMD is a proprietary measurement computer from the ratio of digital data (noise and signal) to noise (dB) (SMD).
2. Resolution and accuracy same as for standard.
3. 20 Hz to 100 kHz in SMD and distortion.
4. Resolution is limited to 0.01 Hz for noise frequencies ± 100 Hz.
5. In dB mode noise measurement is 0.01 dB.
6. Input capacitance is 20 pF for 500 Hz.