IF Filters for the 8901A Modulation Analyzer

by Andrew H. Naegeli

The 8901A Modulation Analyzer has two alternate intermediate frequencies. The primary IF is 1.5 MHz, chosen to accommodate an FM signal with 400-kHz deviation at a 200-kHz rate. The secondary IF is 455 kHz, used to avoid spurious conversion problems in the mixer for input frequencies below 10 MHz. This IF is limited to rates of 10 kHz and deviations of 40 kHz.

Low-Pass Filter and Phase Compensation

The primary IF uses a low-pass filter with a bandwidth of 2.5 MHz. The low-pass filter is used because of the very wide bandwidth requirements of the worst-case FM signal described above. The filter must reject IF signals of 6 MHz and higher by at least 60 dB, and must have a very flat amplitude response to preserve the FM-rejection performance of the AM detector. These requirements led to a 6-pole Cauer-Chebyshev filter design with two zeros of transmission at 6.1 and 8.4 MHz. This filter was synthesized as a passive network using standard inductors and capacitors as the filter elements.

To prevent distortion of FM signals, a phase compensation filter is cascaded with the 2.5-MHz low-pass filter. For an FM signal with 100-kHz deviation at a 100-kHz rate, the phase compensation filter reduces the distortion from 2% to less than 0.1%. This filter is an all-pass network with four poles and four zeros. It is realized as an active filter, as shown in Fig. 1.

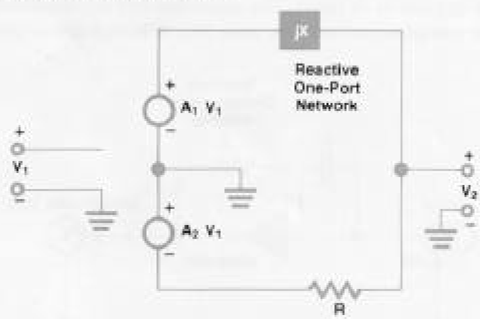


Fig. 1. *All-pass network acts as a phase compensation filter to prevent distortion of FM signals.*

The two voltage sources are models for two amplifiers that deliver the input signal to two points, 180 degrees out of phase. Assume for the moment that A1 = A2 = 1, that is, both channels have the same unity gain. The circuit may be analyzed using superposition. First, with the lower voltage source shorted, the output voltage is:

$$V1\,\frac{R}{R+jX} \qquad (1)$$

Next the contribution from the lower voltage source is found by shorting the upper voltage source:

$$-V1\,\frac{jX}{R+jX} \qquad (2)$$

The sum of (1) and (2) is the output voltage, V2. Therefore, the transfer function is:

$$\frac{V2}{V1}=\frac{R-jX}{R+jX} \qquad (3)$$

This transfer function is an all-pass function because the zeros o[f] transmission (the roots of the numerator) are a mirror image of the poles (the roots of the denominator) across the frequency axis of the s-plane.

The amplitude and phase responses of this network must be adjusted to give optimum performance. Component variations in the reactive elements can change the locations of the poles and zeros, leading to phase changes. The phase response is adjusted to the desired shape by adjusting the value of R.

The amplitude response of the network is theoretically flat, bu[t] losses in the reactive elements can affect the amplitude flatness. Fo[r] example, a small resistance, r, in series with the reactive network changes the transfer function to:

$$\frac{V1}{V2}=\frac{(R-r)-jX}{(R+r)+jX} \qquad (4)$$

Now, if the reactance term, jX, is zero at some frequency, the amplitude is no longer = 1, but is less than 1. This causes ripples that depend on the location of the poles and zeros of the reactance, jX. The reactance network for this filter has a zero at 2.2 MHz and poles at 1.0 and 4.3 MHz. These can cause substantial ripples in the amplitude response. To adjust for this problem, the two gains are adjusted such that A1 is not equal to A2. This changes the amplitude response primarily, so the two adjustments are only marginally interactive.

The transistor circuit realization of this filter is shown in Fig. 2.

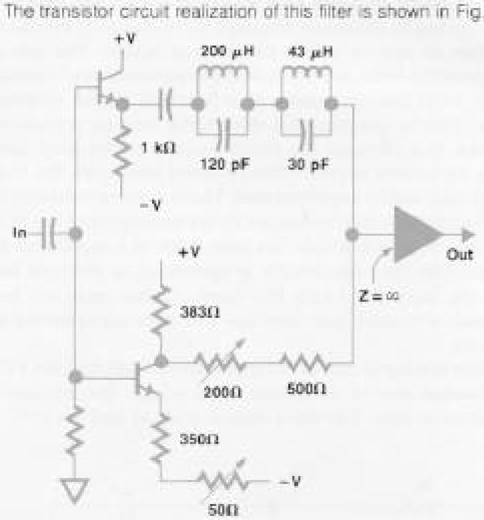


Fig. 2. *All-pass network is a four-pole, four-zero active filter.*

Bandpass Filters

The secondary IF uses a bandpass filter at a 455-kHz center frequency. The 3-dB bandwidth of this filter is 200 kHz. The wide bandwidth is necessary because of the wide deviation of permissible FM signals. The phase response of this filter is very important, because of FM distortion. The filter was designed to add less than 0.03% distortion to an FM signal with 10-kHz deviation at a 10-kHz rate. To accomplish this, the phase response was made very linear, while sacrificing a sharp cutoff in the amplitude response. This required the filter to be designed as an arithmetically symmetrical bandpass filter (same response at $f_0 - \Delta f$ as at $f_0 + \Delta f$, where f_0 is center frequency), as opposed to the standard geometrically symmetrical design (same response at $\frac{1}{2}f_0$ as at $2f_0$).

In standard bandpass filter design, a low-pass prototype filter is transformed geometrically up to the desired center frequency. Unfortunately, this transformation distorts the phase response in the filter passband. To avoid this problem, the desired bandpass function was approximated directly by analyzing the response of pole-zero patterns using a computer. A transitional Butterworth-Thompson design was used as the low-pass prototype, to make the group delay response flat across most of the passband. When the desired constellation of poles and zeros was achieved, the seven-pole bandpass filter was synthesized directly using standard insertion loss techniques.[1]

The filter bandwidth is wide enough so that standard inductors and capacitors are used as the filter elements. The amplitude and phase responses are adjusted using two slightly interactive adjustments, adjusting the value of two inductors. The variable elements were chosen from the results of a sensitivity analysis performed with a computer-aided-design program.

The secondary IF filter may also be used at input frequencies above 10 MHz to make the 8901A more frequency-selective. This can be useful when measuring the level of signals at various frequencies in the input spectrum, using the TUNED RF LEVEL function.

References

1. E.A. Guillemin, "Synthesis of Passive Networks," Wiley, 1957.

Andrew H. Naegeli

Born in Arcadia, California, Andy Naegeli received his BSEE degree from Stanford University in 1975 and his MSEE in 1979, also from Stanford. An HP employee since 1975, Andy worked on the RF input circuits, IF filters and microprocessor software for the 8901A Modulation Analyzer. He is currently a project manager with the HP Stanford Park Division. Married to a professional musician and living in Menlo Park, Andy spends much of his spare time fixing up his 30-year-old house. A reborn Christian, he plays string bass with a folk singing group at his church. He also works with stained glass and enjoys camping, water-skiing and volleyball.

cisely locating signals that are moving (FM), only there part of the time (AM), and that have harmonics only 10 dB down (or 30 dB down if higher than third). In the process of finding the fundamental of the incoming signal the octave oscillator is swept rapidly across its band about 20 times. To do this, a current source is fed into the integrator input to the VCO (Fig. 8). The output of the integrator is a ramp that causes the VCO to move down in frequency. Another current source retraces the VCO at the end of a band if no signal is found.

The microprocessor sweeps each octave until a signal is found. A fast detector in the IF chain senses the presence of a signal in the IF and switches off the current source driving the integrator. When a sweep is successful in generating an IF signal, the local oscillator is counted to determine where it was when it stopped.

The sweep rate is very fast, up to 1 MHz/μs. This presents the problem that any time involved in recognizing the presence of an IF signal and turning the sweep source off causes the VCO to go too far. To compensate for this, a step-back resistor is put in the integrator, such that removing the sweep current removes $V = IR$ from the output voltage of the integrator. R is chosen so that the frequency represented by the sweep current times the resistor is equal to the amount the output frequency moved during the turnoff delay. This allows the output of the VCO to be counted to determine where it was when the signal first entered the IF.

Demodulators

The modulation analyzer's two demodulators operate on the IF signal and produce a pair of audio signals. One demodulator extracts the frequency modulation components of the IF signal and the other extracts the amplitude modulation components. The two demodulators are designed for low susceptibility to incidental modulation, or conversion from one type of modulation to the other. They are also designed not to add significantly to the modulation noise output so that they do not degrade the system residual modulation characteristics. They are also both very linear and have flat frequency responses so that the overall 1% accuracy specification can be maintained. Both detectors do a fine job of meeting their design goals.

The job of removing the AM from the IF signal before it is sent to the FM demodulator is done by a series of three limiter stages. Considerable care was taken to minimize the effect of amplitude variations on the phase of the limiter output (minimum AM to ΦM conversion). It is most important that both the positive half cycle and negative half cycle be treated the same and always symmetrically. It is obvious that if the limiting does not take place exactly at the zero crossing, changes in amplitude will affect the zero crossing and, in turn, the FM output. (However, the 8901A's full-wave FM demodulator averages the zero crossings so that some symmetrical movement can be tolerated.) To minimize this effect, we treat the signal within the limiter in a dual-ended differential arrangement (Fig. 9).

Each limiter stage consists of a differential pair with the emitters tied together and to a current source. The collector loads are individual resistors with emitter followers to take the signal differentially to the next stage or to the output. By maintaining strict symmetry, both phases of the signal are treated the same. Since the zero crossing information is carried in the difference between the two outputs any asymmetry that might exist at one output will not exist in the difference between the two. Symmetry was further assured by having all of the transistors for a given limiter on the same transistor array integrated circuit chip.

Perhaps the most important consideration involves maintaining all of the operating conditions independent of input level. The current source on the emitters keeps the operating current and consequently the output level independent of base voltage. The differential arrangement keeps the transistors out of saturation (a necessity) and the emitter output stages buffer the loads from changes in the input

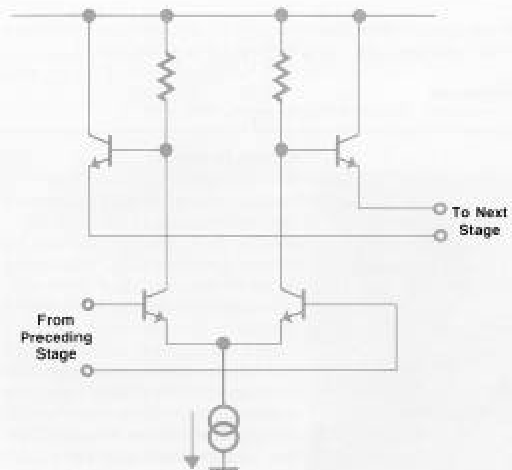


Fig. 9. *Limiters (one stage shown here) remove the AM from the IF signal before it goes to the FM demodulator. The design minimizes AM-to-ΦM conversion.*

impedance of the following stage as it begins to limit.

Some care was taken to make the stage delay independent of input level. Normally the delay through a differential stage gets smaller as the stage is driven harder. In this design, the small-signal bandwidth is extended by applying negative feedback. (When the stage is limiting, the feedback is not effective, of course.) The feedback equalizes the delay through the stage so that the small-signal delay is equal to the large-signal delay.

With all of these steps, phase shift at the limiter output is typically less than ½ degree for a 3-to-1 change in input level.

FM Demodulator

The most difficult task in selecting an FM demodulator was to achieve both the noise specification and the distortion specification we needed for our various potential users. To satisfy the distortion requirements of broadcast FM applications we are providing FM distortion that is typically 70 dB down at rates and deviations up to 100 kHz. This is very good, and suggests a pulse-count type of discrimination. On the other end of our target applications are the high-quality, low-deviation FM mobile radio applications. For these we are providing discriminator noise in a 3-kHz bandwidth that is small compared with 1 Hz (around 1/3 Hz, typically). This is also very good, and suggests a tuned-circuit type of discriminator. The problem is that a pulse-count discriminator is too noisy and a tuned discriminator is not linear enough. We were able to develop a new type of discriminator that is neither pulse-count or tuned-circuit, yet meets all our requirements. The beauty of this circuit is that it is inherently linear and yet does not have a significant noise mechanism. It is described in detail on page 13.

AM Demodulator and AGC

The AM demodulator detects the amplitude modulation on the IF signal and produces an audio-frequency output. This detector has a very wide IF bandwidth and therefore very low conversion of FM to AM. At most operating frequencies, conversion of FM to AM is negligible, approaching the published specification only at the lowest input frequencies when the RF input circuit and the mixer begin to roll off slightly.

The demodulator is a half-wave average-responding rectifier with sufficient bandwidth to work well with an IF above 2.5 MHz. It is similar to a circuit sometimes used at lower frequencies, with diodes in the feedback loop of an amplifier. The difference is that the amplifier is a very high-gain amplifier capable of the high slew rates required to turn the diodes on during the IF signal's transition through zero.

A simplified schematic is shown in Fig. 10. (Some biasing and compensation elements have been omitted.) Gain is provided by a cascade of three transistors arranged for a high gain-bandwidth product. The current source (grounded base) output allows the fast voltage changes required to switch the diodes as the sign of the current reverses. The audio output is taken from one of the half-wave outputs by way of a constant-resistance network. The constant-R network filters the IF signal so that very little of it reaches the buffer amplifier that follows the demodulator. Without the constant-R filter, the detection diode would be back-biased at each IF zero crossing by the average value of the half-wave-rectified IF signal, and this would degrade performance because the amplifier would have to slew farther.

The IF gain of the circuit is R_2/R_1. To have reasonable gain and yet keep R_2 small enough to get the required bandwidth, R_1 has to be small. Its value is actually 100Ω, so

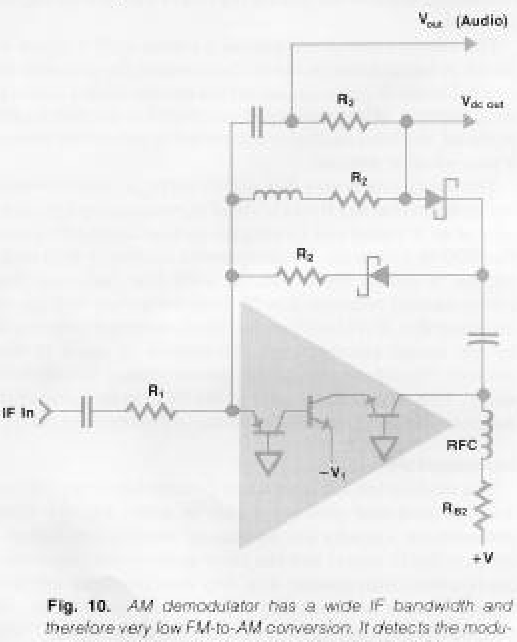


Fig. 10. *AM demodulator has a wide IF bandwidth and therefore very low FM-to-AM conversion. It detects the modulating signal (V_{out}) and the average carrier level ($V_{dc\ out}$).*

it requires a good stout IF amplifier to drive it.

The amplitude modulation percentage is the amplitude of the AM detector audio output divided by the average carrier level times 100%. The average carrier level is also detected by the AM detector; it is labeled $V_{dc\,out}$ in Fig. 10.

An AGC circuit ahead of the AM demodulator (Fig. 11) provides a constant-level IF signal to the demodulator. $V_{dc\,out}$ is compared with a stable reference and an error

A New Type of FM Demodulator

by Russell B. Riley

We have called the wideband low-noise FM demodulator used in the 8901A Modulation Analyzer a charge-count discriminator. The basic idea is to form pulses of constant charge at a rate proportional to frequency and then average these pulses to produce an output voltage proportional to the input frequency.

The operation of the charge-count discriminator is similar to that of the more familiar pulse-count discriminator, the basic difference being the pulse shapes involved. The typical pulse-count circuit forms a pulse of constant amplitude and constant duration once per cycle of the signal to be demodulated. Note that both the amplitude and the duration have to be controlled with great accuracy and stability for linear, low-noise performance. In practice it is usually jitter in the pulse duration that limits noise performance. In contrast, the charge-count circuit requires only that a dc voltage be accurate and stable.

In the circuit diagram (Fig. 1), diodes D1 and D2 clamp the left end of capacitor C1 to a voltage swing of 15V plus two diode drops. Charge-steering diodes D3 and D4 limit the right end of C1 to a voltage excursion of two diode drops. Thus the charge that flows back and forth in C1 is well defined and is given by

$$Q = C1\ \Delta V$$

where ΔV is 15V in this example. Since current is given by the rate of change of charge with respect to time the average current I_{avg} flowing in resistor R, through the action of the charge-steering diodes and the operational amplifier U, is given by

$$I_{avg} = Qf = C1f\ \Delta V$$

where f is the number of cycles per second (the frequency of the input signal). The average output voltage V_{avg} is simply

$$V_{avg} = RI_{avg} = RC1f\ \Delta V$$

Thus the demodulator linearity depends on passive components and a stable voltage ΔV. The principal source of noise is the op-amp U, with resistor R and noise on ΔV making somewhat smaller contributions.

The circuit actually used in the modulation analyzer includes some refinements. For example, transistor Q1 is also used, driving circuitry similar to that connected to the collector of Q2. The two outputs are added, with the result that the demodulator transfer function (sensitivity) is doubled and the output ripple frequency is doubled, making the filtering job easier. A damped inductor is added in series with C1 to introduce a controlled amount of overshoot so that the steering diodes cut off cleanly.

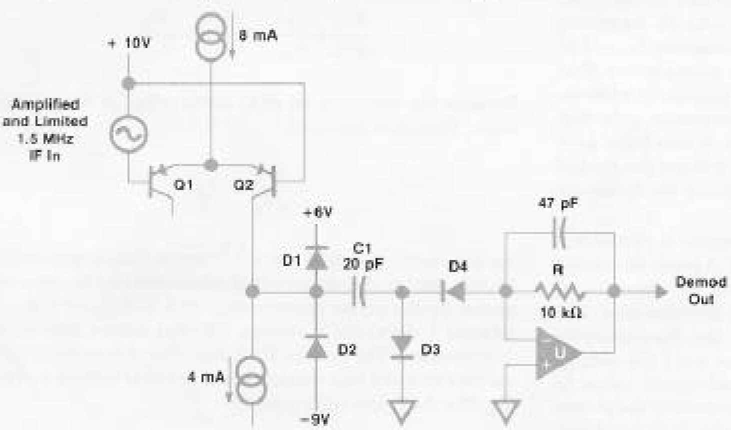
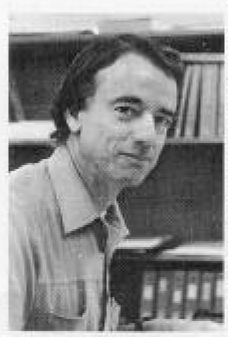


Fig. 1. *Simplified schematic of the charge-count discriminator.*



Russell B. Riley

A 1959 graduate of the University of Colorado, Russ Riley completed his studies for the PhD degree at Stanford University in 1961. His responsibilities at HP have included the 938A and 940A Frequency Doubler Sets, waveguide thermistor mounts, the 423A and 424A Crystal Detectors, the 415E SWR Meter, the 432A Power Meter, and parts of the 8558A Spectrum Analyzer. He is named as inventor on several patents on these products. Most recently he contributed to the design of the 8901A Modulation Analyzer, especially the FM limiter-discriminator; he's now with the optoelectronics section of HP Laboratories. Born in Kansas City, Missouri, Russ is married and lives in Portola Valley. His daughter attends high school, and his two sons are in college. He spends some of his leisure time gardening and singing in his church choir, and he played clarinet in the Peninsula Symphony for three years.