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A Versatile Wave Analyzer for the 1 kc to 1.5 Mc Range

THE transistorized wave analyzer (precision tunable voltmeter) introduced by Hewlett-Packard in 1958 greatly simplified the measurement of harmonic and intermodulation components in the audio frequency range. Automatic frequency control in this instrument, the Model 302A, made possible the use of an extremely narrow passband in the measurement circuits without requiring constant retuning to prevent loss of a drifting signal.

SEE ALSO:
Measuring Loop Gain, p. 5

Other features contributing to the versatility of the 302A included three output signals. One of these is a signal source controlled by the instrument's tuning, making the analyzer a self-contained test set for frequency response measurements. Another output is a dc current, proportional to meter deflection, for operation of graphic recorders. The third output is an ac voltage identical in frequency to the signal component being measured, enabling precision measurement of the component's frequency with a digital counter. Requiring no lengthy warm-up procedure, the 302A has a wide dynamic range and an excellent signal-to-noise ratio (input noise $<10\ \mu v$) enabling the measurement of weak signals in the presence of noise.

Now, a new -hp- wave analyzer (Fig. 1) provides the same ease of operation and convenience features in the 1 kc to 1.5 Mc range, mak-

Fig. 1. *New -hp- Model 310A Wave Analyzer measures amplitude of individual frequency components in 1 kc-1.5 Mc region over a dynamic range of more than 75 db. SSB carrier system testing in this frequency range is facilitated by carrier reinsertion provision contained in analyzer.*

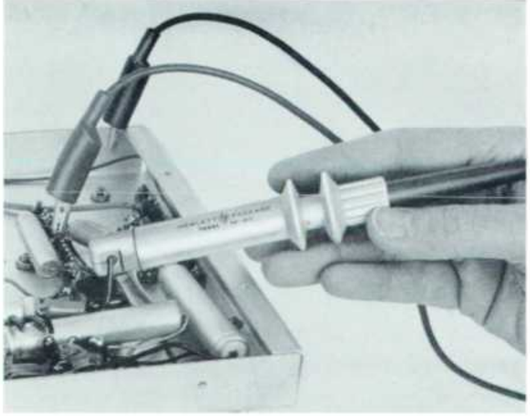

Fig. 2. *Using new analyzer with -hp- current-sensing probe, loop gains of feedback amplifiers can be measured without breaking loop. See article on p. 5.*

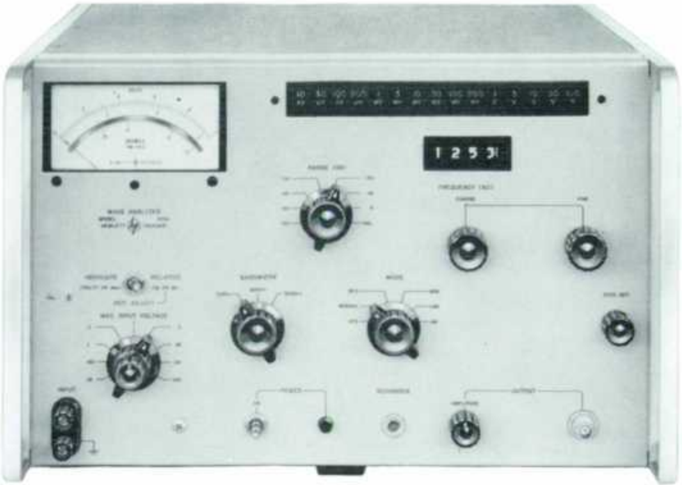

Fig. 3. *In new analyzer full scale value of voltage range in use is automatically indicated by illuminated readout system at right top of panel. Frequency tuning is indicated by digital indicator.*

ing these measurement techniques available for use in ultrasonic and low-frequency RF systems. The new model 310A analyzer has additional features for carrier telephone service, such as choice of bandwidths, AM demodulation, and carrier reinsertion for demodulating SSB signals. The three output signals available in the audio range 302A analyzer also are available in the 310A, as shown in Fig. 4.

BASIC OPERATION

Fundamentally, a wave analyzer resembles a superheterodyne receiver, as shown in Fig. 5. The signal being measured is translated to a higher frequency band by the local oscillator. The IF amplifier passes only the 3 Mc component of the translated band of frequencies, and this component is passed to the meter circuit for amplitude measurement. Tuning the local oscillator shifts the translated signal components along the frequency spectrum so that each individual harmonic component can be moved into the IF "slot" for measurement. In this way, the wave analyzer identifies the frequency and measures the amplitude of each harmonic component in a complex wave.

Actually, a precision wave analyzer such as the 310A has many added features to extend the usefulness of the instrument as well as to facilitate measurements and insure measurement accuracy. The 310A has three selectable bandwidths. The narrow 200 cps bandwidth has high selectivity for separating closely-spaced signal components and the 1000 cps bandwidth makes signal searching easier, as well as simplifying calculations of noise power per cycle bandwidth. The 3000 cps passband enables monitoring of radio or carrier telephone channels during set-up. To complement this latter capability, the 310A has an AM demodulator as well as a carrier reinsertion oscillator for demodulation of single-sideband signals.

The inherent noise level of the 310A is less than 1 μv even with the wide 3 kc passband. This permits higher sensitivity, the instrument having 15 sensitivity ranges from 10 μv full scale to 100 v full scale in steps of 3 to 1.

CIRCUIT ARRANGEMENT

The complete block diagram of the *-hp-* 310A Wave Analyzer is shown in Fig. 6. The signal to be measured passes through the high-impedance input attenuator, which prevents large signals from driving the low distortion preamp into non-linear operation. The low-pass filter rejects frequencies above 1.5 Mc, providing a high degree of attenuation of signals in the 6 to 7.5 Mc band, which otherwise would cause troublesome image responses unless removed. For making measurements on a relative basis, the preamplifier may be switched so that the gain is adjustable, enabling the signal amplitude to be set at a convenient reference level.

The input signal is multiplied by the highly stable local oscillator signal in the balanced mixer. The lower "sideband" (difference frequency) is coupled through the low-pass filter to one section of the Range attenuator which enables the measurement of weak signal com-

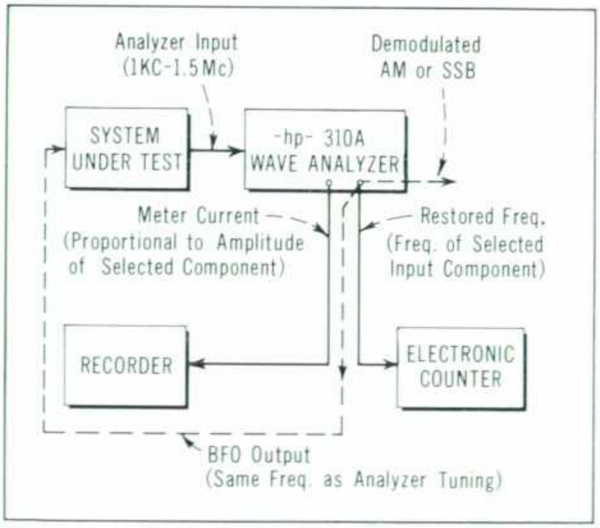

Fig. 4. *New Model 310A Wave Analyzer provides outputs for driving system under test, for operating dc recorder, for counter, for carrier insertion, and a detected AM output. Outputs are not all concurrent.*

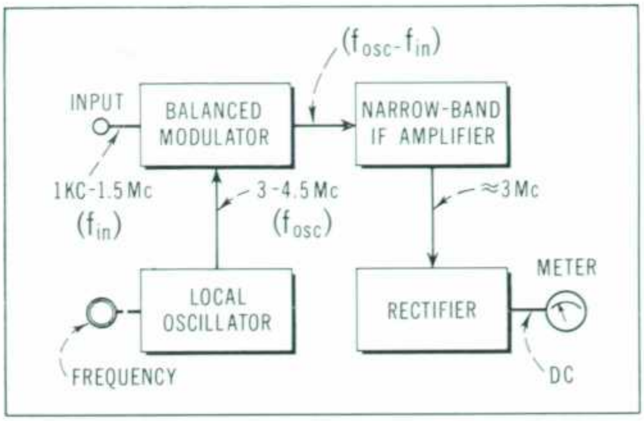

Fig. 5. *Basic circuit arrangement of new Model 310A Wave Analyzer. Local oscillator can be operated under AFC control to simplify acquiring and tracking input signal.*

ponents in the presence of larger components. The IF bandpass filter passes only those IF signal components lying adjacent to the 3 Mc IF center frequency. The filtered IF is subsequently rectified for amplitude measurement by the front panel meter. The unfiltered rectifier output, which follows signal fluctuations, serves as the AM output.

The 3 Mc IF, with the instrument in the *AFC* or *Normal* mode, is mixed with the local oscillator signal in the output mixer to obtain the "restored" frequency output. This procedure translates any signal component passing through the IF channel back to its original frequency.

When the instrument is in the *BFO* mode, a crystal-controlled 3 Mc signal is mixed with the local oscillator signal to derive a signal which corresponds to the frequency to which the analyzer is tuned, but which is unaffected by the IF channel. This signal is useful as a source for making frequency response measurements, such as loop gain characteristics or filter responses.

The same BFO is used as a carrier reinsertion oscillator in the *USB* and *LSB* modes of operation. The BFO frequency is raised or lowered 1.5 kc so that it lies at either the upper or lower edge of the IF 3000 cps passband.

The discriminator supplies a dc voltage to voltage-variable capacitors in the local oscillator, when the instrument is in the *AFC* mode, to slave the oscillator to any signal lying near the center of the IF passband.

THE BANDPASS FILTER

The 310A uses a bandpass filtering technique based on "polyphase modulation" which translates the IF frequency down to the audio range, where it is relatively easy to design filters with the desired flat passband and sharp cut-off characteristics, and then restores the filtered signal to the original IF frequency.

The IF bandpass filter, shown in the block diagram of Fig. 7, works as follows. An IF signal, adjacent to 3 Mc, is multiplied by a 3 Mc carrier in balanced modulators in both channels. (This essentially translates the IF band of frequencies down to a new band centered at 0 cps, with

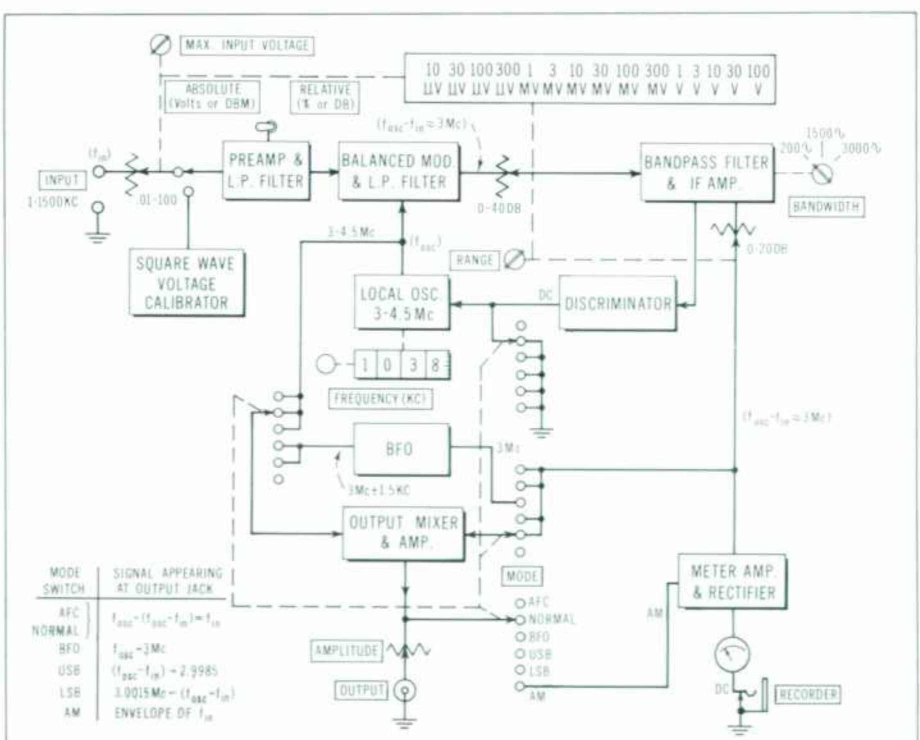

Fig. 6. *Circuit arrangement of Model 310A Wave Analyzer.*

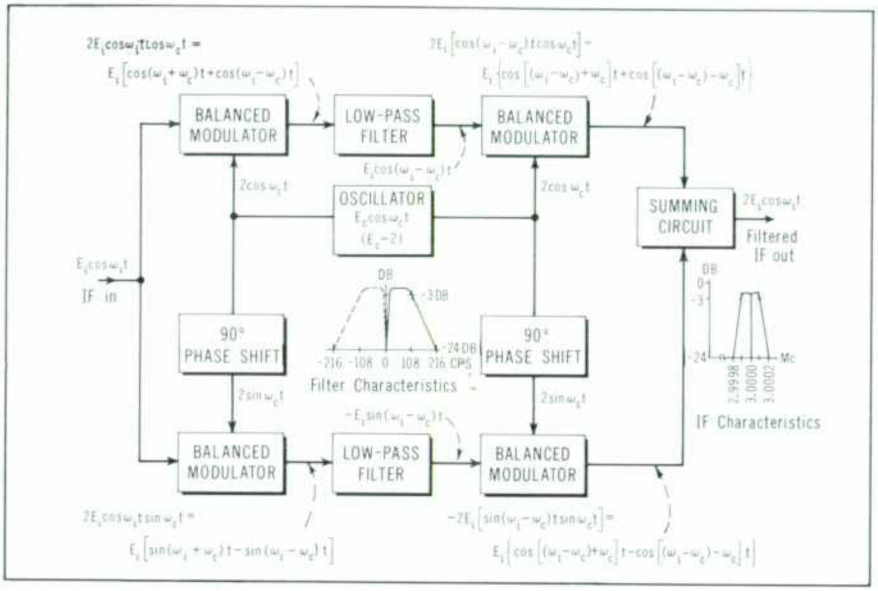

Fig. 7. *Block diagram of IF bandpass filter used in new analyzer. Operations on signal are shown for an input signal lying within filter passband. Carrier amplitude E_c is assumed as 2 v for simplicity.*

one of the resulting two sidebands consisting of "negative" frequencies.) The carrier supplied to the lower modulator, however, is 90 degrees out of phase so that the resulting modulator output is 90 degrees out of phase with the upper channel modulator output.

With the narrow-band filters switched into the circuits, for instance, only those IF signal components which are within 100 cps of the 3 Mc carrier will generate difference frequencies low enough to pass through the filters. The audio range frequencies then are restored to the original IF frequency in a second pair of modulators.

The second modulator in the lower channel is also driven by a carrier which is 90 degrees out of phase, so that the original IF input frequency is recovered in phase with the upper channel, as shown by the trigonometric derivation in Fig. 7. The other "sideband," however, turns out to be 180° out of phase with the same sideband in the upper channel. These signal components therefore cancel each other in the summing circuit, while the desired signal components add. The audio range filter characteristics thus are translated to the IF frequency spectrum.

ACTIVE RC FILTERS

The audio range filters use active RC filters which do not have the difficult hum shielding problems associated with the large inductances required by low frequency LC filters. The filter configuration is shown in Fig. 8. Capacitor C1 is returned to a point of unity positive gain on the amplifier output, and at low frequencies both sides of this capacitor are subject to the same ac voltage so that it exerts no influence on the circuit.

At higher frequencies, C2 and the series resistors attenuate the input signal. The voltage returned to C1 experiences a like amount of attenuation, causing further signal attenuation. At frequencies far above cut-off, the network behaves as a two-section RC filter, dropping off at 12 db per octave. Sharper cut-off characteristics are obtained by using two active filters in cascade falling off at 24 db per octave with respect to the filter cut-off frequency. The cut-off frequency is changed readily by switching network resistors and capacitors.

AC coupling of the filters means,

SPECIFICATIONS

-hp-
MODEL 310A
WAVE ANALYZER

FREQUENCY RANGE: 1 kc to 1.5 Mc (200-cps bandwidth); 5 kc to 1.5 Mc (1000-cps bandwidth); 10 kc to 1.5 Mc (3000-cps bandwidth).
FREQUENCY ACCURACY: ± (1% + 300 cps).
SELECTIVITY: 3 bandwidths, 200 cps, 1000 cps and 3000 cps. Midpoint (f_0) of passband readily distinguished by rejection region (1 cps wide between 3-db points). (See table below)
VOLTAGE RANGE: 10 μvolts to 100 volts full scale. Ranges provided by INPUT attenuator and meter RANGE switch in steps of 1:3 or 10 db.
VOLTAGE ACCURACY: ± 6% of full scale.
INTERNAL CALIBRATOR STABILITY: ± 1% of full scale.
DYNAMIC RANGE: Greater than 75 db.
NOISE AND SPURIOUS RESPONSE: At least 75 db below full-scale reference set on 0-db position of RANGE switch.
INPUT RESISTANCE: 10 k ohms on most sensitive range, 30 k ohms on next range, 100 k ohms on other ranges.
AUTOMATIC FREQUENCY CONTROL: Dynamic hold-in range is ± 3 kc, minimum, at 100 kc. Tracking speed is approximately 100 cps/sec. Locks on signals as low as 70 db below full-scale reference set on 0-db position of RANGE switch.
RESTORED FREQUENCY OUTPUT: Restored signal frequency maximum output is at least 0.25 volt (at full scale deflection) across 135 ohms; approximately 30 db of level control provided. Output impedance, approximately 135 ohms.
BFO OUTPUT: 0.5 volt across 135 ohms; approximately 30 db of level control provided. Output impedance approximately 135 ohms.
RECORDER OUTPUT: 1 madc into 1500 ohms or less for single-ended recorders.
RECEIVER FUNCTION (Aural or Recording Provision): Internal carrier reinsertion oscillator provided for demodulation of either normal or inverted single sideband signals. AM signal also can be detected.
POWER: 115 or 230 volts ± 10%, 50 to 1000 cps; approx. 16 watts.
WEIGHT: Net 44 lbs. Shipping 59 lbs.
DIMENSIONS: 16¾ in. w. x 10½ in. h. x 18⅜ in. d.
ACCESSORIES AVAILABLE: 11001A Cable, BNC male to dual banana plug, 45 in. long; 10503A Cable, BNC male connectors, 4 ft. long; 10111A Adapter, banana plug to BNC female.
PRICE: -hp- Model 310A, $2,000.00.

Prices f.o.b. factory
Data subject to change without notice

BANDWIDTH SETTING	REJECTION		
	≥ 3 db	≥ 50 db	≥ 75 db
200 cps	f_0 ± 108 cps	f_0 ± 500 cps	f_0 ± 1,000 cps
1000 cps	f_0 ± 540 cps	f_0 ± 2400 cps	f_0 ± 5,000 cps
3000 cps	f_0 ± 1550 cps	f_0 ± 5000 cps	f_0 ± 17,000 cps

A QUICK, CONVENIENT METHOD FOR MEASURING LOOP GAIN

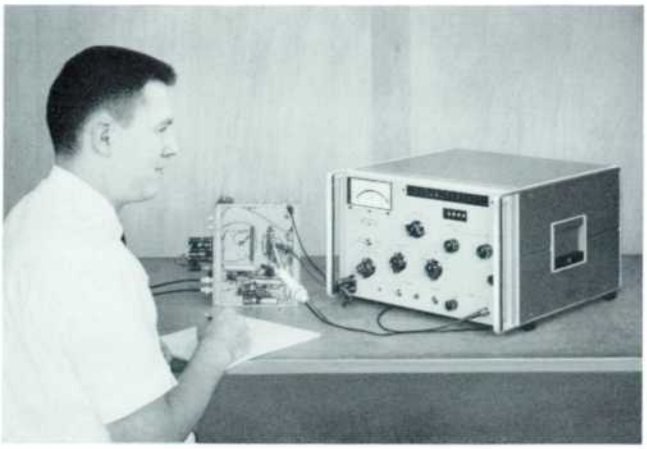

Fig. 1. *By using an -hp- AC-21F clip-on probe to couple signal output from -hp- Models 302A or 310A Wave Analyzer into feedback loop, loop gains can be measured easily and without breaking loop.*

CONVENTIONALLY, measurements of loop gain Aβ are made by opening the feedback loop and then measuring the output obtained in response to a known input. Difficulties arise here, though, because the simulated load impedance must duplicate the impedance presented to the output stage when the loop is closed, and auxiliary bias sources must be added if dc feedback is employed.

New techniques now allow measurement of loop gain with the loop closed, providing rapid, easily-obtained results[1]. These measurements are made with the *-hp-* AC-21F current probe for signal injection, and either the *-hp-* 302A or 310A wave analyzer for signal measurement. The current probe, used inversely to its usual current-sensing function, serves as a coupling transformer for feeding the driving signal into the system, simply by being clipped around a circuit lead. Values of Aβ over a wide range of frequencies and magnitudes, including Aβ less than unity, are readily obtained. In addition, the phase angle of Aβ at frequencies near gain crossover is easily determined.

THEORETICAL CONSIDERATIONS

Insertion of an isolated voltage source in series with the signal path of a feedback system does not alter the characteristics of the feedback loop, an ideal voltage source having zero series impedance and no shunt conductances to ground. Voltages

[1] B. M. Oliver and C. O. Forge, private communication.

WAVE ANALYZER

(Continued from page 4)

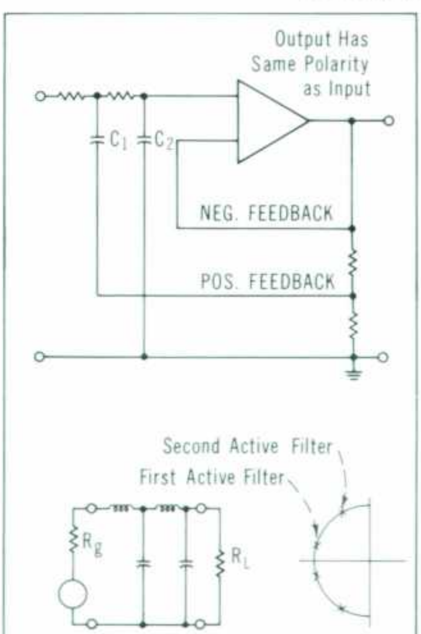

Fig. 8. *Passband of new analyzer is shaped using active filter (top) which minimizes shielding problems. Two filters are cascaded (below) to achieve sharp cut-off. Passband is made maximally flat by placing poles on semicircle in complex frequency plane.*

of course, that frequencies of less than a few cycles per second are attenuated. When translated back to the original IF frequency, the overall passband has a notch at 3 Mc, the notch being less than 1 cps wide at the 3 db points. The notch enables a signal to be tuned precisely to band center for precision frequency measurement. The discriminator, on the other hand, locks on to the edge of the notch so that a harmonic component is not attenuated by the notch during amplitude measurement.

ACKNOWLEDGMENTS

Members of the design team for the *-hp-* 310A were Richard Van Saun, Richard Raven, Richard Osgood, and the undersigned. We are all grateful for the suggestions and ideas of Brunton Bauer, Paul Stoft, Dr. B. M. Oliver, and others.

—Stanley McCarthy

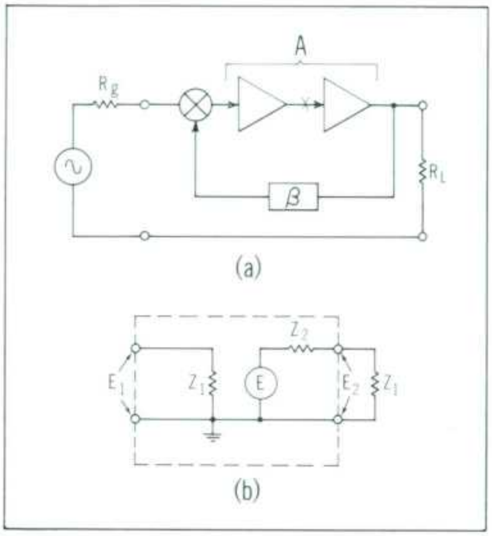

Fig. 2. *(left) (a) Diagram of basic feedback amplifier. (b) Alternate representation of (a).*

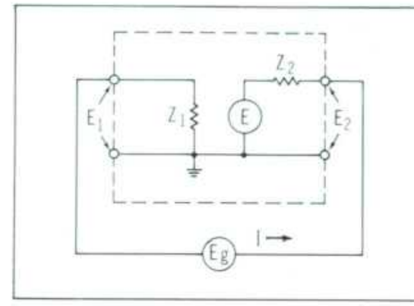

Fig. 4. *Circuit representation when voltage is injected into loop.*

are established, however, which allow $A\beta$ to be determined directly. To understand how $A\beta$ can be measured in this manner, consider the feedback amplifier shown in Fig. 2(a). The amplifier has the normal generator and load impedances connected and the loop is opened at some convenient point (not necessarily in the β circuit). A duplicate of the impedance Z_1, measured when looking into the system at the break point, is connected to the new output, as shown in Fig. 2(b).

Since E_1 is modified by both A and β when traveling around the loop,

$$E_2 = A\beta E_1 \qquad (1)$$

The voltage source E is simply:

$$E = \frac{Z_1 + Z_2}{Z_1} E_2, \qquad (2)$$

$$\text{or, } E \doteq \frac{Z_1 + Z_2}{Z_1} A\beta E_1 \qquad (3)$$

Now consider the situation in Fig. 4. Here, the loop is closed and a voltage source is connected in series with it. This represents the normally closed feedback loop since no additional impedances have been introduced. The disturbance created by the presence of the voltage E_g, however, causes voltages E_1 and E_2 to be established by the reaction of the feedback loop.

The voltage on the output side of the generator is:

$$E_2 = IZ_2 + E \qquad (4)$$

The current may be expressed as:

$$I = E_1/Z_1 \qquad (5)$$

Substituting equations (5) and (3) for I and E respectively in equation (4) yields:

$$E_2 = \frac{E_1}{Z_1} Z_2 + \frac{Z_1 + Z_2}{Z_1} A\beta E_1 \qquad (6)$$

If $Z_2 << Z_1$ then $E_2 = A\beta E_1$, as in equation (1), even though E_g has been added to the circuit. Thus,

$$A\beta = E_2/E_1 \qquad (7)$$

Thus it is seen that simply by introduction of the voltage E_g in series with the loop, two voltages E_1 and E_2 are established which determine $A\beta$ directly.

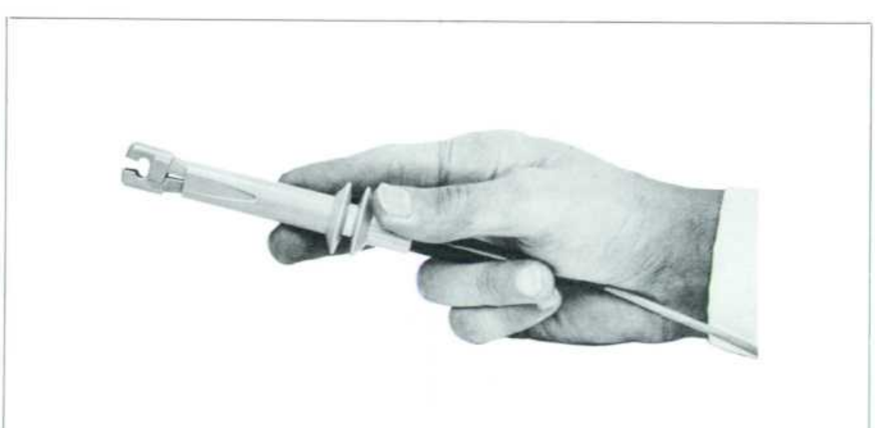

Fig. 3. *-hp- AC-21F clip-on probe, normally used for sampling ac current, is used in loop gain measurements to inject signal from Analyzer into loop.*

The voltage source E_g may be placed at any point in the loop where the signal is confined to a single path and where $Z_2 << Z_1$. The load and generator impedances normally used with the amplifier should be connected to the normal output and input terminals.

The amplitude of E_g must be small enough to avoid saturation in any of the active elements and consequently, either E_1 or E_2 will be quite low. Sensitive wave analyzers, such as the *-hp-* Models 302A or 310A (see article on page 1), are well-suited to making $A\beta$ measurements involving these small signals. Narrow bandwidths insure a high degree of noise and spurious signal rejection. The signal available from the wave analyzer operating in the BFO mode can be used for E_g, so that both source and measurement circuits are tuned simultaneously.

The series impedance introduced into the test circuit by the clip-on ac current probe is approximately 0.01 Ω shunted by 1 microhenry, and shunt impedance is only about 2 pf. When driven by the wave analyzer, the voltage produced in the test circuit is about 10 mv, a convenient level.

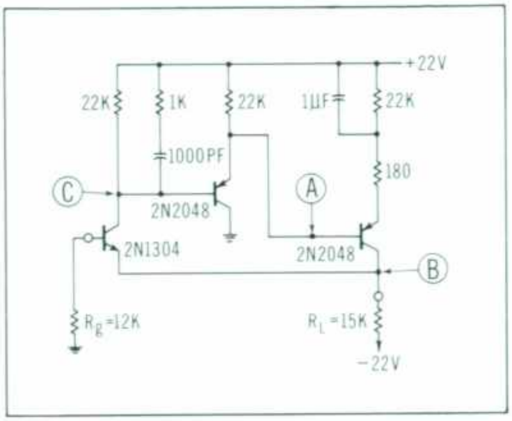

Fig. 5. Circuit of amplifier on which loop gain was measured using technique described in text.

PRACTICAL EXAMPLE

The loop gain of the amplifier shown in Fig. 5 was measured with this technique, E_g being inserted at point A. At this point, Z_2 was calculated to be no more than 400 Ω and Z_1 was about 10,000 Ω. The requirement that $Z_2 << Z_1$ is satisfied here. The plot of measured loop gain versus frequency is shown in Fig. 6.

To read loop gain directly in db units, E_2 is set to the 0 db level on the analyzer by adjusting the amplitude of E_g. E_1 consequently is measured in negative db units and, when the sign is reversed, these readings represent $A\beta$ in db.

Note that loop gains of less than unity (below 0 db) are easily measured. In this case, the 0 db reference is set to E_1 and then E_2 represents the value of $A\beta$ in db units.

Measurement of $A\beta$ values less than unity can be useful. For instance, if the circuit is not stable when the loop is closed, resistive attenuation may be introduced somewhere in the loop to avoid oscillations. The relative values of $A\beta$ then are measured and when plotted, the reasons for instability may be determined.

The phase angle of $A\beta$ is readily determined through construction of a vector diagram, as shown in Fig. 7. This is merely a graphical depiction of the relation: $E_2 = E_1 + E_g$.

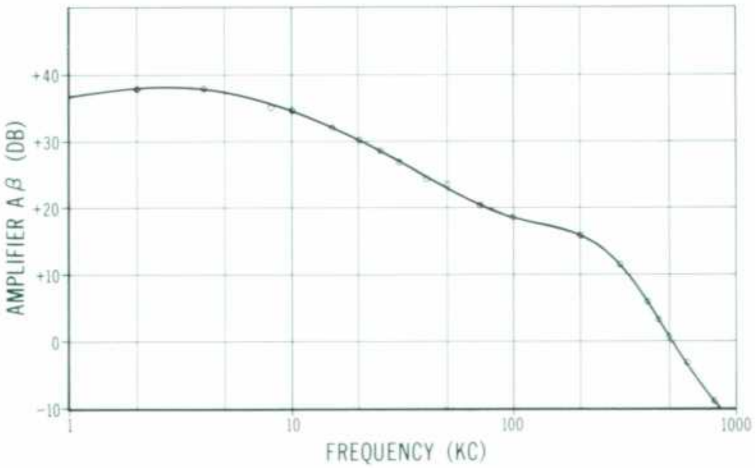

Fig. 6. Loop gain characteristic measured on amplifier of Fig. 5.

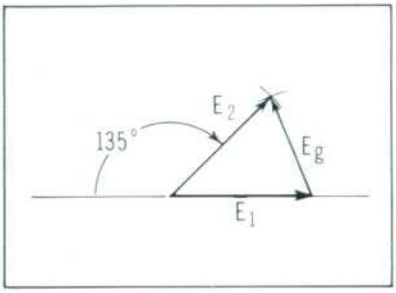

Fig. 7. *Phase angle of loop is easily determined by constructing basic diagram.*

E_2 and E_1 are measured directly and E_g is measured by shorting the voltmeter input leads together and clipping the current probe around them. For negative feedback, the phase angle usually is measured from the -180 degree reference.

ALTERNATIVE METHOD

It may not always be possible to find a point where $Z_2 << Z_1$. A similar measurement technique, the dual of the voltage technique, applies when $Z_2 >> Z_1$. The amplifier of Fig. 2 is shown in Fig. 8 with a current source connected from the signal path to ground. As before, the feedback loop is closed but current source I_g causes I_1 and I_2 to be established.

Referring to Fig. 8:

$$E_1 = -I_1 Z_1 \qquad (8)$$

$$\text{and } E_2 = I_2 Z_2 + E \qquad (9)$$

Substitution of equation (3) gives:

$$E_2 = I_2 Z_2 + \frac{Z_1 + Z_2}{Z_1} A\beta E_1 \qquad (10)$$

Since $E_2 = E_1$, equations (8) and (10) may be combined:

$$-I_1 Z_1 = I_2 Z_2 - \frac{Z_1 + Z_2}{Z_1} A\beta I_1 Z_1 \qquad (11)$$

If $Z_2 >> Z_1$:

$$\text{Then } A\beta = I_2/I_1 \qquad (12)$$

A dual to the first method therefore exists, with currents replacing voltages in the determination of loop gain.

As in the voltage case, the normal input and output load impedances should be connected. The temporary input and output again may be chosen at any point where the signal is confined to one path. A resistor usually is adequate for converting a voltage generator to a current source (a capacitor may be placed in series with the resistor to block dc). In this case, the resistance should be large with respect to Z_1.

This technique was also used to measure the loop gain of the amplifier shown in Fig. 5. Point B was selected as the current node. Here, Z_2 is the output impedance of an amplifier with local emitter feedback, approximately one megohm, and the input impedance of the following emitter is about 270 Ω which meets the requirement that $Z_2 >> Z_1$.

A current source was simulated by connecting a 10K resistor ($>> Z_1$) in series with the wave analyzer's BFO output. The current probe sensed each current I_2 and I_1, supplying a proportional voltage to the input of the wave analyzer (termination of the current probe is not required since only relative measurements are being taken). Using this technique, the maximum deviation from the values of $A\beta$ obtained by the voltage source method was only 0.3 db.

Since $I_2 = I_1 + I_g$, a vector diagram may be constructed to find the phase angle of $A\beta$, as was done in the first method.

DC LOOP GAIN

Another technique, primarily useful for obtaining the dc loop gain, is based upon the equation[2]:

$$Z_{fb} = Z_{nf} \frac{1 - A\beta_{sc}}{1 - A\beta_{oc}} \qquad (13)$$

where:

Z_{nf} = impedance between two nodes with A reduced to zero,

[2] T. S. Gray, "Applied Electronics," 2d Ed. p. 587.

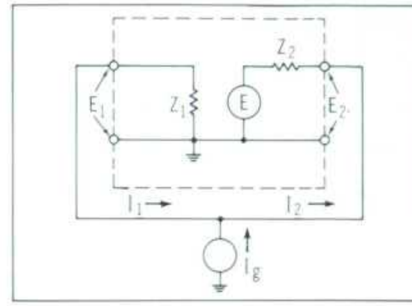

Fig. 8. *Circuit representation when current is injected into loop.*

Z_{fb} = impedance observed when normal feedback is present,

$A\beta_{sc}$ = loop gain with nodes shorted together, and

$A\beta_{oc}$ = loop gain when no external admittance is connected between the nodes.

At dc, two nodes usually can be found where $A\beta_{sc}$ or $A\beta_{oc} = 0$ and where Z_{nf} can be calculated. Then, by connecting a current source between the nodes, and noting the voltage change, Z_{fb} can be calculated from equation (13).

To measure the dc loop gain of the amplifier shown in Fig. 5, a current of 36 μa was injected between point C and ground. A voltage change of 0.4 v at point C was observed. Thus, Z_{fb} is $0.4/36 \times 10^{-6} = 11$ k. Since the input impedance of the stage connected to this point, and also the output impedance of the previous stage, are very high, Z_{nf} is the same as the collector load resistance (22k). If point C were grounded, $A\beta_{sc}$ would be zero and if it were left ungrounded, $A\beta_{oc}$ would equal $A\beta$, the normal loop gain. Substituting these values in equation (13) yields:

$$11{,}000 = 22{,}000 \frac{1}{1 + A\beta}$$

from which,

$$A\beta = \frac{22{,}000}{11{,}000} - 1 = 1$$

—Philip Spohn