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Ethan Aronoff, Engineer, Radio Beam Expansion Project

Measurement of Discontinuities in Waveguides

SUMMARY

Discontinuities, in long waveguide runs carrying a frequency-modulated radio relay signal, cause signal reflections (echoes) which in turn cause intermodulation distortion. A frequency-modulation radar method for determining the location and magnitude of these discontinuities has been prepared.[1,2] This article discusses the application of this method using commercially available equipment.

The two most important characteristics of the measurement method are its sensitivity, i.e., the smallest reflection that can be measured with a given accuracy, and its resolution, the smallest distance between two separately measurable discontinuities. The equipment discussed has a signal-to-noise ratio of 6 db for a 1 percent reflection 12 feet away, with a distance of 1 foot between separately resolvable discontinuities anywhere within a 100-foot run.

The theory of the measurement method, the use of the equipment, and methods for maximizing the sensitivity and resolution are described.

Purpose of Waveguide Runs

The poor front-to-back rejection ratio of the passive reflector antenna systems, currently being installed in conjunction with the Western Union cross-country radio beam expansion program, prevents the use of an r-f channel more than once at a station.[3] The use of long waveguide runs feeding tower mounted antennas permits the use of an r-f channel twice at a station, thus doubling the capacity of the microwave system. As the need for greater system capacity develops, the passive reflector antenna system will be replaced by the long waveguide run and tower mounted antenna system. The measurement system described here may be used to check the quality of the waveguide run at the time of installation and thereafter on a periodic preventive maintenance basis. Some of these waveguide runs will be 300 feet or more in length. The longest Western Union waveguide run tested thus far with this method in the field is 125 feet in length.

This article was delayed in publication until it was presented as a Transactions paper at the Winter General Meeting of the IEEE in New York, January 1963.

Effect of Waveguide Runs

The addition of waveguide runs to the system will cause an increase in system noise. The waveguide attenuation adds to the path loss, thus causing an increase in system white noise, which lowers the fade margin of the system. Signal reflections from the antenna input-port, the equipment output-port, and from discontinuities in the run, cause echo distortion,[4] a form of intermodulation distortion. These discontinuities appear usually at the waveguide joints but they can also be caused by dents in the waveguide between joints. Lewin[1] has shown that the echo distortion is proportional to the magnitude of these reflections, raised to the fourth power.

To ascertain the importance of low reflections in the waveguide run, a calculation was made of the change in system noise due to a change in the magnitude of these reflections, using typical values. The calculation follows: Assume that a given waveguide run with a joint voltage standing wave ratio (VSWR) of 1.02 (approximately 1 percent reflection) is connected between an antenna whose input VSWR is 1.06 and radio equipment whose output VSWR is 1.04. Assume that the echo distortion due to this run is 12 picowatts of psophometric noise.[4]

Then, assume that slightly less care is used in the construction and assembly of the components of this system so that the joint VSWR is 1.03, the antenna input VSWR is 1.09, and the equipment output VSWR is 1.06. (This assumption is made on the basis that the reflections at each discontinuity are now 50 percent higher.) The echo distortion increases from 12 picowatts to 60 picowatts ($1.5^4 \times 12 = 60$). This indicates the importance not only of keeping these reflections as low as possible but also of being able to accurately measure VSWR's in the order of 1.01 to 1.04 and above.

of this energy back towards the mixer. When the returning r-f energy arrives back at the mixer, the forward wave at the mixer has changed in frequency. Figure 2 is a plot of the frequency vs. time, at the mixer, for the forward wave and for the reflected wave, for a single discontinuity. τ is the time required for the r-f energy to travel from the mixer to the discontinuity and back again to the mixer. f is the beat, or difference frequency between the forward and reflected waves. As is shown below, f is directly proportional to the length of waveguide between the discontinuity and the mixer.

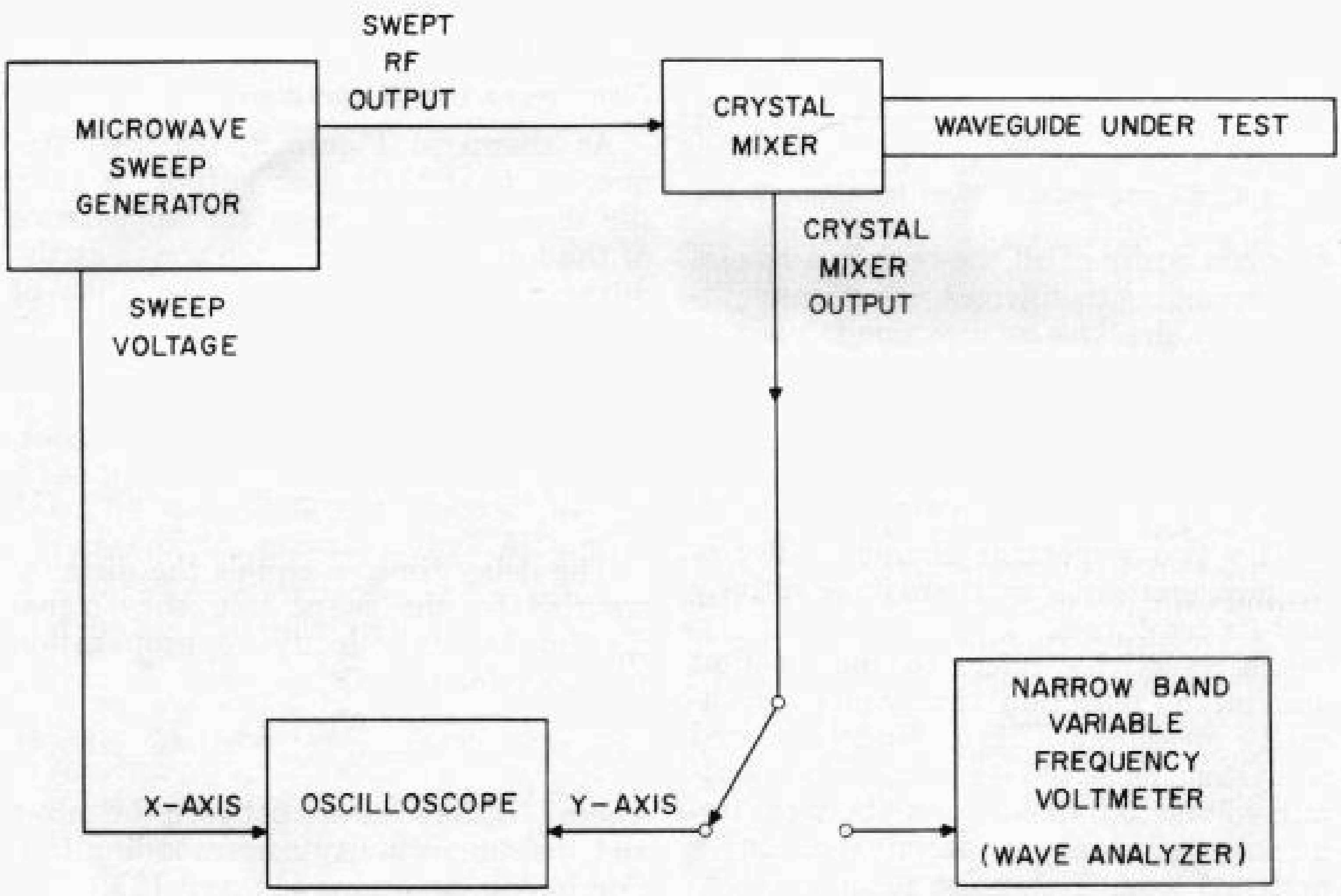


Figure 1. Simplified Block Diagram of Reflection Measurement System

System Block Diagram

A simplified block diagram of the reflection measurement system proposed by Western Union is shown in Figure 1. A microwave sweep oscillator generates r-f energy into a transmission line with a microwave frequency that varies approximately linearly with time. The r-f energy travels past a crystal mixer and down the line. A discontinuity in the waveguide run, such as that at a joint, reflects part

The heterodyning of the forward and reflected waves in the crystal mixer produces the difference frequency at the crystal mixer output. As is shown below, the magnitude of the crystal mixer output is directly proportional to the reflection coefficient of the discontinuity. Thus the equipment measures the two pertinent facts about the discontinuity: its location, and the magnitude of the reflection it causes.

The output of the crystal mixer can be applied either to a wide-band oscilloscope or to a narrow-band variable-frequency voltmeter (wave analyzer). The wave form shown on the oscilloscope is a com-

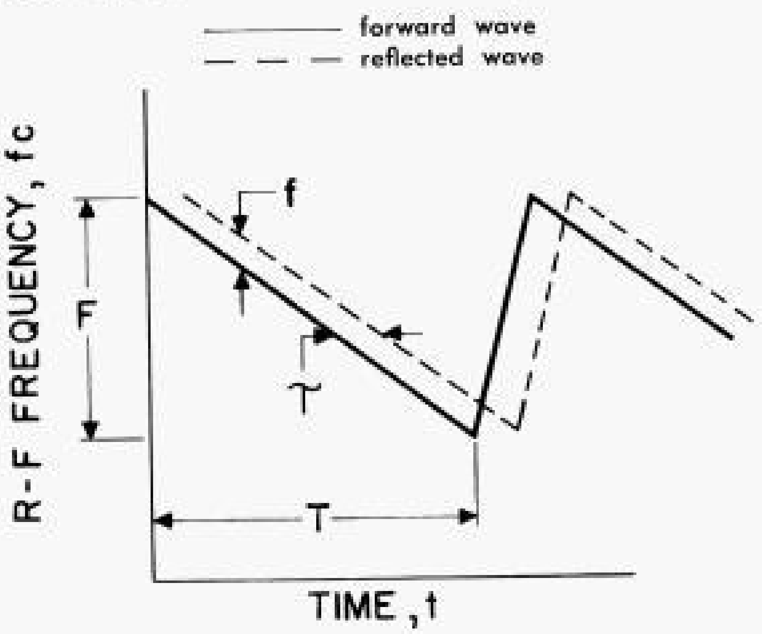


Figure 2. R-F Frequencies at Mixer Input-Linear Sweep

plex composite of all the beat frequencies corresponding to different reflections within the guide. The oscilloscope is used to set up the equipment while the wave analyzer is used to examine specific points within the waveguide run.

Characteristics of the System

The two important characteristics of the measurement system are its sensitivity and its resolution.

The sensitivity refers to the smallest amount of reflection that can be accurately measured. This is directly related to system noise. Noise is the mixer output voltage obtained when the measurement equipment is terminated in a matched load. This noise remains essentially unaltered when the matched load is removed and the waveguide under test is connected to the measuring equipment. The wave analyzer measures only magnitude and not phase. Thus, there is an uncertainty as to the magnitude of a reflection at a given point in the waveguide equal to twice the noise voltage at that frequency corresponding to the given point in the waveguide.

The resolution refers to the minimum distance between discontinuities which can be distinguished one from another. Since the smaller this distance, the better, and since high resolution connotes desirability, the reciprocal of this distance in feet^{-1} is set equal to the resolution. The resolution has been defined as being equal to the reciprocal of the minimum time difference between echoes which can be distinguished one from another[2]. The definitions are almost equivalent but the former is simpler to interpret.

A 1 percent reflection in a 100-ft. WR-229 waveguide run can be measured with a signal-to-noise ratio of 6 db at a discontinuity distance of 12 ft. The signal-to-noise ratio increases rapidly at greater distances. The resolution obtained throughout the 100-ft. run is 1 ft. $^{-1}$.

Beat Frequency Versus the Distance to the Discontinuity

As shown in Figure 2, the beat frequency, f, at the mixer output, is to τ, the delay time between the appearance of the forward and reflected waves at the mixer, as F, the frequency deviation of the sweep generator, is to T, the "on" time of the sweep generator. T is approximately equal to the reciprocal of the sweep generator repetition rate, f_r. Thus,

$$\frac{f}{\tau} = \frac{F}{T} \approx Ff_r \tag{1}$$

The delay time, τ, equals the distance traveled by the swept frequency signal divided by its velocity of propagation (group velocity). Or,

$$\tau = \frac{2L}{V_g} \tag{2}$$

where L is the distance between the mixer and the discontinuity corresponding to f. Combining equations (1) and (2):

$$f = \frac{2L}{V_g}\,\frac{F}{T} \tag{3}$$

$$\text{or } f \approx \frac{2L}{V_g}\,Ff_r$$

When the r-f frequency vs. time plot is linear,

$$f = \frac{2L}{V_g}\,\frac{df_c}{dt} \tag{4}$$

where f_c is the instantaneous r-f frequency, ignoring any negative sign that may appear in the derivative. When a non-linear r-f frequency vs. time characteristic is used (for a reason shown below), neither (3) nor (4) is correct but (4) is a close approximation. The resultant beat frequency when using a sweep generator with a nonlinear r-f frequency vs. time-characteristic is shown in Figure 3. In

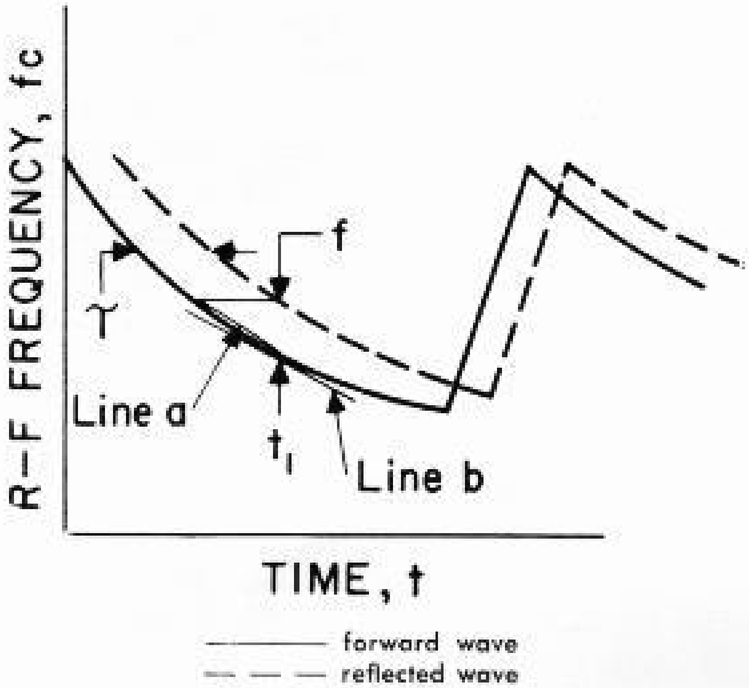
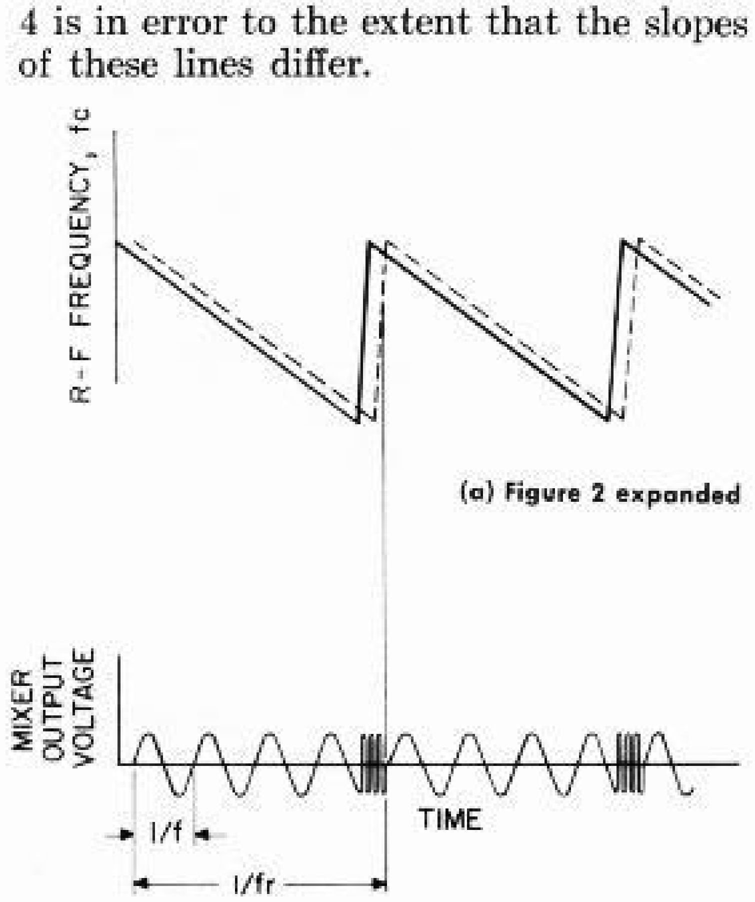


Figure 3. R-F Frequencies at Mixer Input—Non-Linear Sweep

the illustration f at $t = t_1$, is the slope of "Line a" multiplied by the delay time τ. Equation 4 states that f is the slope of "Line b" multiplied by τ. Thus, Equation 4 is in error to the extent that the slopes of these lines differ.



(b) Mixer Output Voltage for Single Echo

Figure 4.

Effect of Sweep Generator Periodicity

If the sweep generator output r-f frequency were to change at a constant rate for all time, a given discontinuity would produce a beat frequency given by Equation 4. The output of the mixer would then be a pure sinusoid for each discontinuity. It is, of course, impossible to maintain this linear frequency change indefinitely. Therefore, a wave form such as that shown in Figure 4a is employed. A typical mixer output, corresponding to this sweep generator periodic wave form, is shown in Figure 4b. This is a periodic wave form with a period equal to the reciprocal of the sweep generator repetition frequency. It is a well-known result from Fourier analysis that any periodic wave form of repetition frequency, f, can be decomposed into a summation of cosines of the argument, $n2\pi ft + \phi_n$, where n takes on only integer values and ϕ_n is a function of n. The repetitive wave form of Figure 4b can be decomposed into an average component, a component $A_1 \cos (2\pi f_r t + \phi_1)$, a component $A_2 \cos(4\pi f_r t + \phi_2)$, and so on. A plot of A_n as a function of n is called the harmonic amplitude spectrum of the signal. (In this system, only the A_n's are measured and not the ϕ_n's.) For the wave form of Figure 4b, the value of n for which A has its maximum amplitude is the integer nearest f/f_r. If f/f_r is an integral value plus or minus ½, the two adjacent A_n's centered about f/f_r are of equal amplitude.

As an example, if $f/f_r = 9.5$, A_9 and A_{10} are the peak values of the harmonic amplitude spectrum. Figure 5 is a plot of the time wave form for this case and Figure 6 shows the corresponding amplitude spectrum. An adjustment of the sweep generator frequency deviation, F, can be made, slightly changing f so that f/f_r takes on an integer, or near integer value. This places most of the mixer output energy due to a single discontinuity into one harmonic. Figures 7 and 8 show the corresponding signal and its amplitude spectrum.

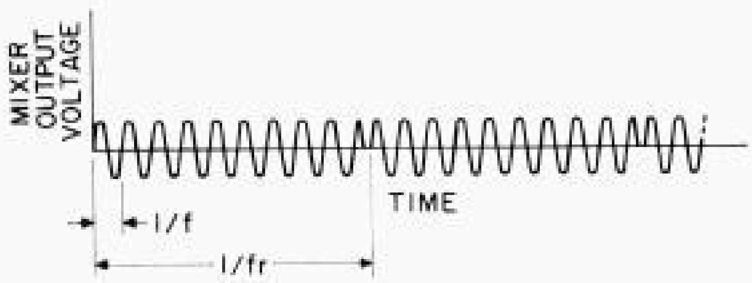


Figure 5. Mixer Output Waveform at $f/f_r = 9.5$

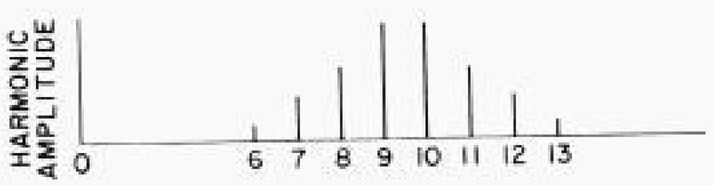


Figure 6. Amplitude Spectrum of Mixer Output Waveform at $f/f_r = 9.5$

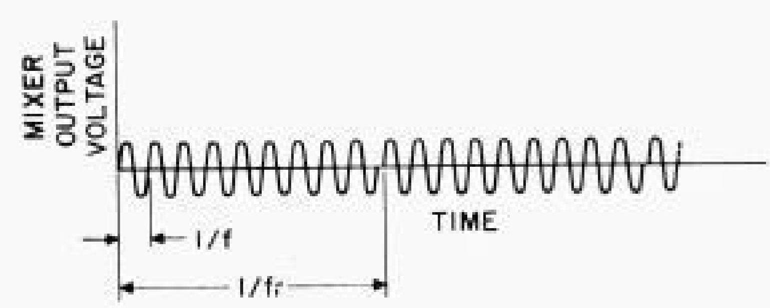


Figure 7. Mixer Output Waveform at $f/f_r = 9.0$

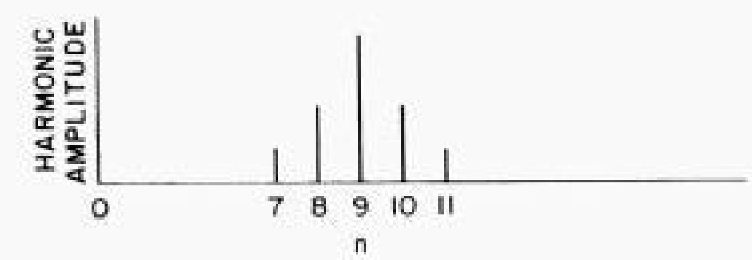


Figure 8. Amplitude Spectrum of Mixer Output Waveform at $f/f_r = 9.0$

Mixer Output Voltage versus Reflection Coefficient of Discontinuity

Essentially the mixer output voltage varies linearly with the magnitude of the reflection coefficient of the discontinuity.

A phasor diagram of the mixer input voltage is shown in Figure 9. E is the phasor representing the mixer input voltage, $E \sin 2\pi f_c t$, due to the forward traveling wave, where f_c is the instantaneous sweep generator r-f frequency. The phasor, rE, represents the mixer input voltage due to the backward traveling wave resulting from a given discontinuity where r is the reflection coefficient of the discontinuity. This phasor represents the following function of time: $rE\sin 2\pi(f_c + f)t$ where f is the beat frequency discussed above. The phasor rE revolves about the arrow-tip of phasor E at the difference frequency f. The instantaneous phase angle between E and rE is $\phi = 2\pi ft$. R, the phasor representing the resultant input voltage to the mixer, has a magnitude

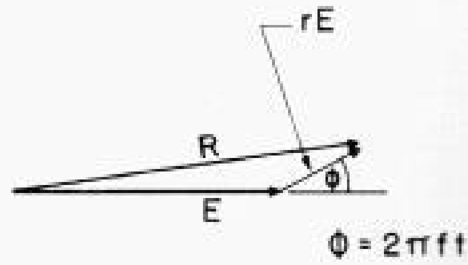


Figure 9. Phasor Diagram of Input Voltages to Mixer

$$E\sqrt{1 + 2r \cos \phi + r^2}.$$

For small values of r, this is approximately equal to

$$E(1 + r \cos \phi)$$

with an error of less than 0.125 percent for an r as large as 5 percent. Thus, phasor R represents the following function of time:

$$E(1 + r \cos 2\pi ft) \sin 2\pi f_c t,$$

the amplitude modulated (by r, at frequency f) input voltage to the mixer, drawn in Figure 10.

Another consequence of a small r is that the envelope of the detected crystal output current corresponds to a linear portion of the crystal characteristic. The consequence of this is that the varying component of the mixer output is a linear function of the varying component of the mixer input (the component due to r). The r-f component of the mixer output is filtered out, and the average value of the r-f wave shown in dotted line in Figure 10 is the component of the mixer output that is observed. The mixer output has a d-c component (Line c in Figure 10) approximately equal to $\frac{1}{\pi}$ E times the slope (A) of the line drawn from the origin to the operating point on the crystal current-voltage characteristic (Line a on Figure 10). The total observed mixer current (after the filtering out of r-f), as a function of the input signal, can then be expressed as:

$$i \simeq \frac{A}{\pi} E + r \frac{BE}{\pi} \cos 2\pi ft \qquad (5)$$

where B is the differential slope of the crystal current-voltage characteristic at the operating point (line b in Figure 10).

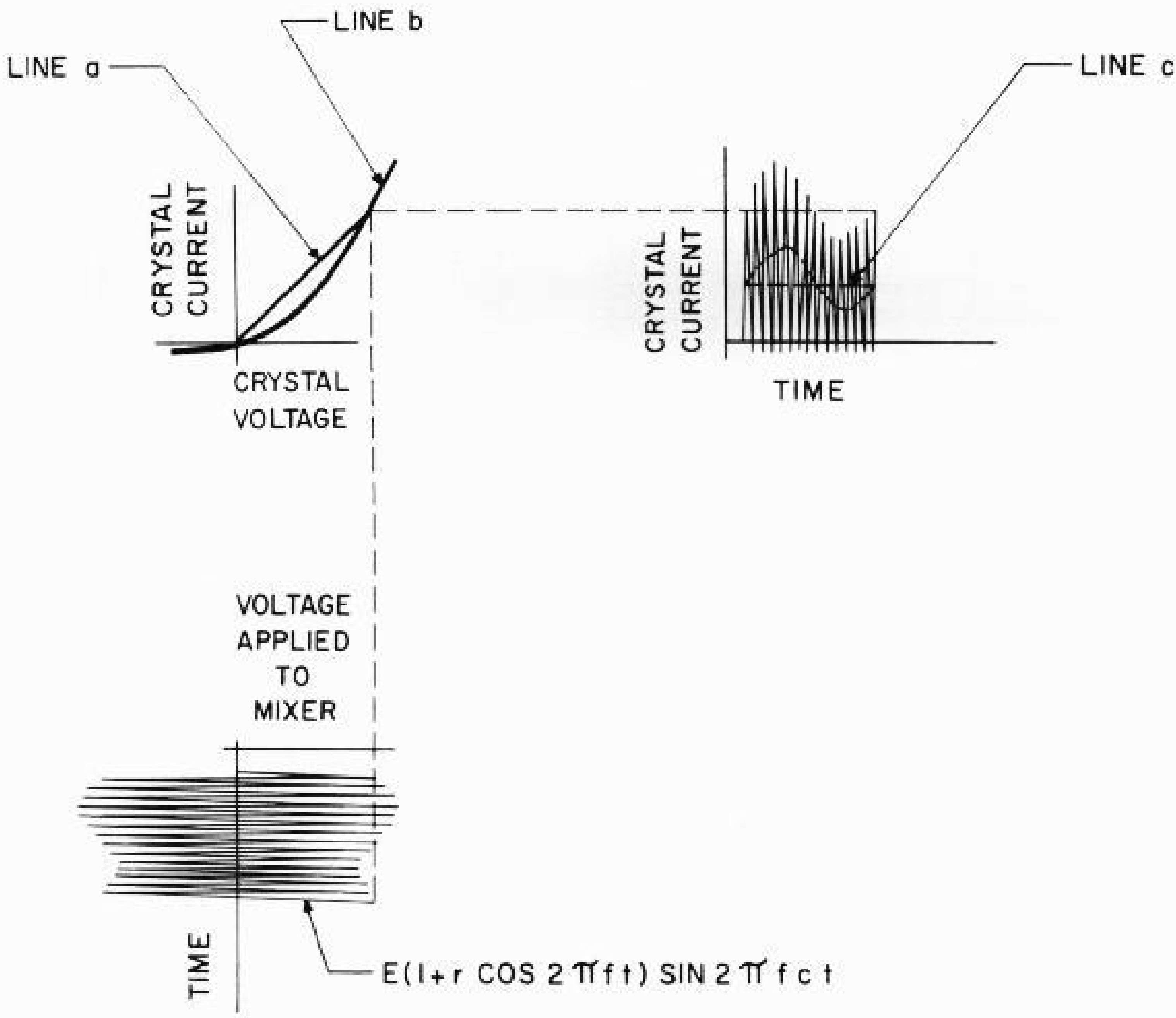


Figure 10. Crystal Mixer Input/Output Waveforms

Effect of Amplitude Modulation of Sweep Generator Output

Ideally, the peak amplitude of the sweep generator r-f output does not vary as a function of frequency. In general, this amplitude varies with frequency—a typical example is shown in Figure 11. An overall variation has superimposed upon it, the fine-grain ripple of the backward wave oscillator (BWO) tube used in the sweep generator. When the mixer is terminated in a good match, so that r is zero, the output current of the mixer (see Equation 5) equals $\frac{A}{\pi}$E. Figure 12, for the ideal case, and Figure 13, for the typical case, show the detected r-f output of the sweep generator with an assumed perfect match after the measurement equipment. The wave form of Figure 12 is a repetitive wave form with an amplitude spectrum consisting essentially of a d-c component alone for very large ratios of T_{on}/T_{off}.

As T_{off}/T_{on} increases, the amplitude of the fundamental and of the harmonics of the repetition frequency increase. In other words, if the sweep generator repetition frequency is 100 cps for a matched load, the mixer output consists of a large d c component, a small component at 100 cps, a smaller component at 200 cps and so on. As T_{off} increases, the voltages at 100 cps, 200 cps, and so on, increase. These voltages are present with a matched load at the mixer output and hence are noise voltages.

The harmonic amplitudes of the amplitude modulated sweep, shown in Figure 13, are even greater than those of the ideal sweep shown in Figure 12. To minimize the effects of this sweep amplitude modulation, a balanced mixer is used. A schematic diagram of the balanced mixer circuit is shown in Figure 14. The balanced mixer consists of two single-ended mixers, one having a normal polarity crystal, the other having reversed polarity crystal. Ideally, the outputs of the two mixers, for a matched load and an amplitude modulated sweeper, add to produce a flat, zero voltage response at the output of the balanced mixer. Signal voltages due to line discontinuities do not cancel but add. Probes in the waveguide sample the electric fields of the forward and backward traveling waves. These probes are spaced n + ¼ wavelengths apart, where n is an integer.

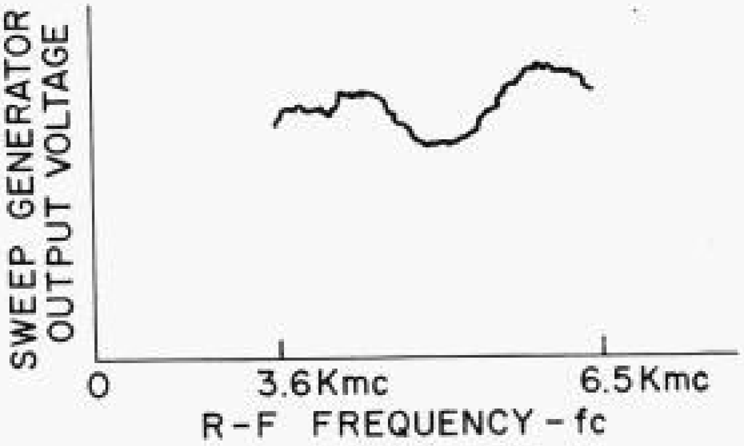


Figure 11. Output Amplitude of Sweep Generator

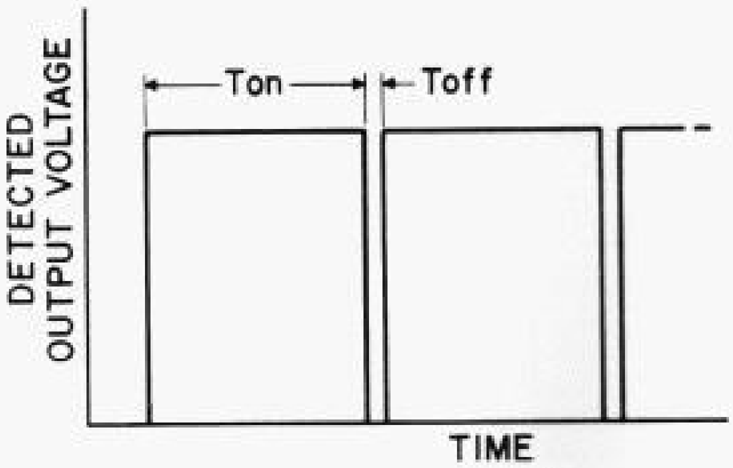


Figure 12. Detected Output Amplitude of Ideal Sweep Generator

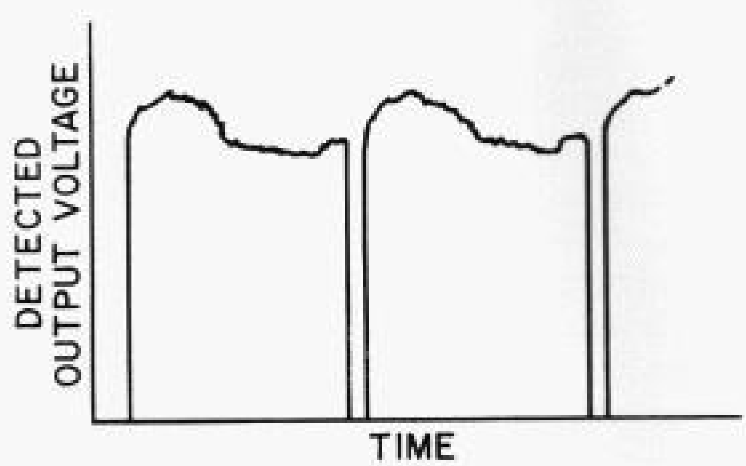


Figure 13. Detected Output Amplitude of Typical Sweep Generator

The phasor diagrams of the forward and backward traveling waves at both probes are shown in Figure 15. The forward traveling wave E at probe #2 leads E at probe #1 by 90°; the backward traveling wave rE at probe #2 lags rE at probe #1 by 90°. The phase angle between E and rE, at probe #1 is

$$2\pi ft$$

at probe #2 is

$$2\pi ft + 180°$$

To a first order approximation, equation 5 applies when E is amplitude modulated by a function of time, m(t). Or,

$$i(t) \approx \frac{A}{\pi} m(t)E + r\frac{B}{\pi} m(t)E \cos 2\pi ft \quad (6)$$

This is only an approximation, since a large variation in m(t) can appreciably change the operating point on the crystal current-voltage characteristic, changing A and B. Since r is usually less than 5 percent and since A and B are of the same order of magnitude, the first term in Equation 6 is much larger than the second. As discussed above, the harmonics of the first term, whose frequency is close to f, may be large in amplitude, causing noise. The output current of the positive polarity crystal connected to probe #1 is:

$$\frac{A}{\pi} m(t)E + r\frac{B}{\pi} m(t) E \cos 2\pi ft$$

The output current of the negative polarity crystal connected to probe #2 is:

$$-\frac{A}{\pi} m(t) E - r\frac{B}{\pi} m(t) E \cos (2\pi ft + 180°)$$

Both crystals feed a common load. The total output current, the sum of the above two expressions is:

$$\frac{2}{\pi} rBm(t) E \cos 2\pi ft$$

Thus, the signal components due to r add while the components due to the amplitude modulation term cancel. Note that the amplitude modulation term m(t) does appear in the expression for the output signals. If the sweep generator amplitude modulation varies widely with r-f frequency, changing the frequency deviation, F, or the start frequency may affect the reflection amplitude calibration of the equipment. (As mentioned above, large

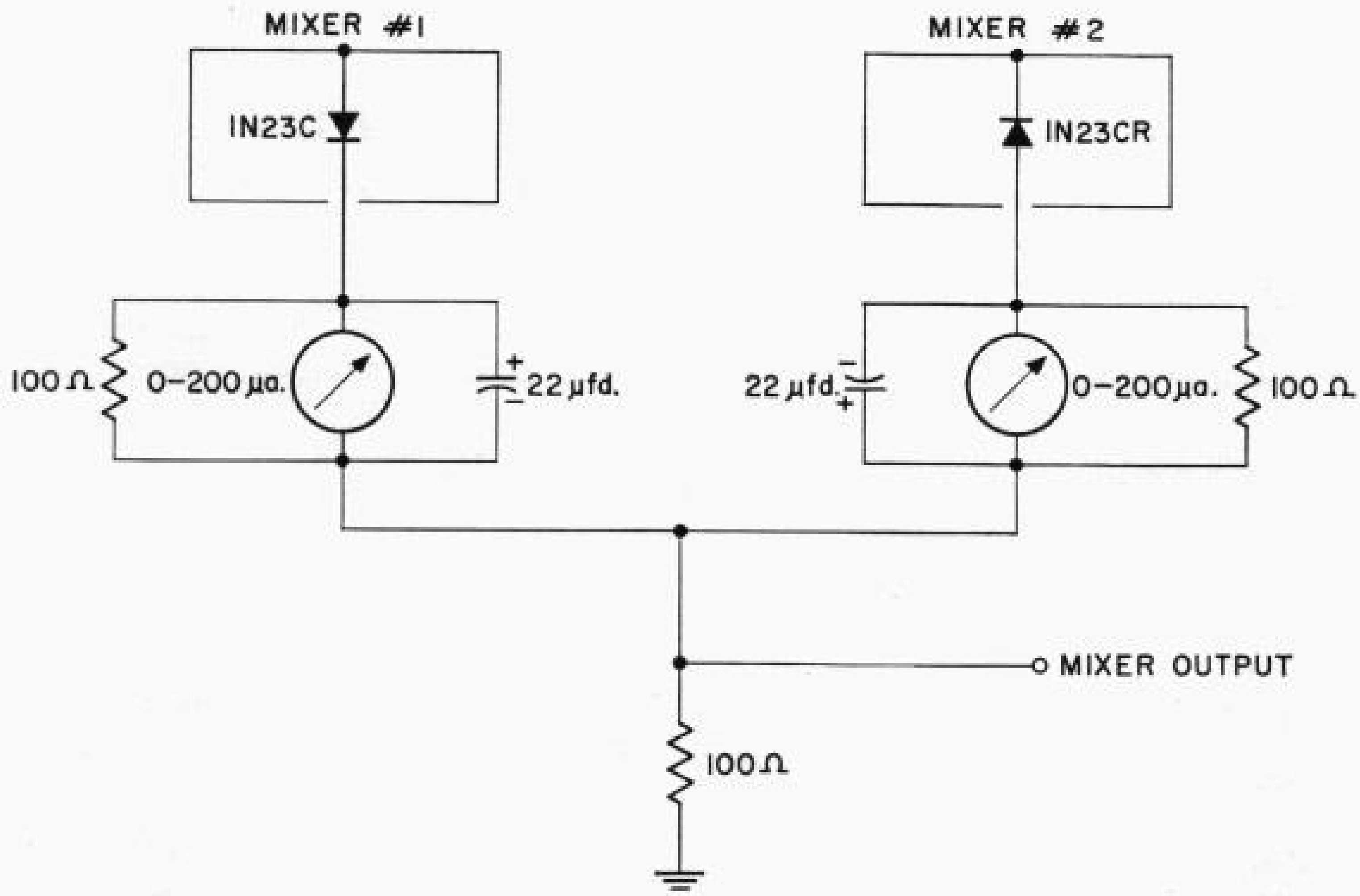


Figure 14. Schematic Diagram—Balanced Mixer Circuit

variations in m(t) affect the linearity of the detection process, thus affecting the amplitude calibration.) It is therefore, most desirable that the sweep generator have as little of this amplitude variation as possible. Also, due care must be taken to check the reflection amplitude calibration when changing the frequency deviation or the start frequency. Commercially available sweep generators use amplitude leveling circuitry which greatly reduce these amplitude variations. However, as discussed below, the use of these amplitude levelers greatly reduces the resolution and so cannot be employed.

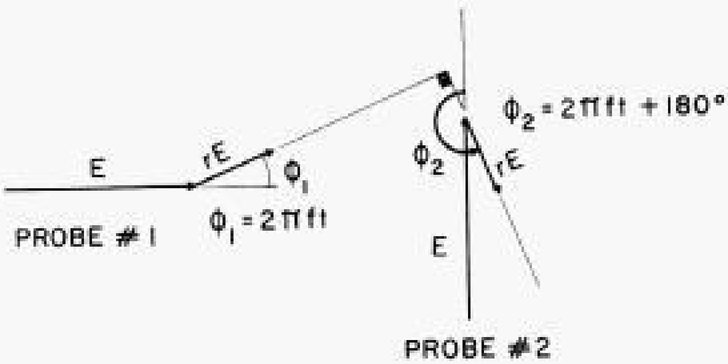


Figure 15. Phasor Diagrams of Input Voltages of Balanced Mixer

The ¼ wavelength spacing is possible at one frequency only. A ¼ wavelength spacing at 3950 Mc (the center of the 3700-4200 Mc common carrier band) is 0.234 wavelengths at 3700 Mc and 0.266 wavelengths at 4200 Mc. An (n + ¼) wavelength spacing can be used, with reduced bandwidth as n increases. A 1¼ wavelength spacing at 3950 Mc is 1.17 wavelengths at 3700 Mc and 1.33 wavelengths at 4200 Mc. Thus, the ¼ wavelength spacing results in a spacing close to ¼ wavelength over the 3700-4200 Mc band; the 1¼ wavelength spacing results in a greater deviation from the ¼ over the same frequency range. For this reason, it is desirable to keep n as low as possible.

In practice, each single-ended mixer is adjusted for maximum response, using a matched load, with the oscilloscope at the mixer output. The input to each mixer is then adjusted to obtain cancellation, or near cancellation of the sweep amplitude modulation. The relative spacing of the two electric field probes leading to the mixers is then adjusted for maximum signal amplitude from a single discontinuity.

It has not been found possible to obtain complete cancellation of the amplitude modulation. The first few harmonics of the sweep generator repetition frequency, with a matched load, are therefore quite large, setting a lower limit on the distance to a measurable discontinuity. At a distance of 12 ft., the noise is about 6 db below the signal corresponding to a 1.02 VSWR discontinuity, the noise decreasing rapidly as the distance increases as shown in Figure 16.

Resolution

The resolution of the measurement system is defined as the reciprocal of the minimum distance between discontinuities which can be distinguished one from another. A single discontinuity produces not a single beat frequency at the mixer output but a spectrum of frequencies with components at the harmonics of the sweep generator repetition frequency. The harmonic with the largest amplitude has a frequency which corresponds most closely to that given by Equation 3. The frequency difference between the maximum amplitude harmonic and the next adjoining harmonic (or between any two adjoining harmonics) corresponds to a distance along the waveguide. There is an uncertainty as to the location of the discontinuity which is equal to the distance between harmonics. Also, it is impossible to resolve, or ascertain the relative magnitudes of, discontinuities spaced closer than the distance between harmonics. Therefore, in order to measure very precisely the distance to a given discontinuity and the reflection from a discontinuity that is near another discontinuity, it is desirable that the distance between harmonics be as small as possible.

It will now be shown that this distance is directly proportional to the group velocity of the traveling waves within the waveguide, and inversely proportional to F, the frequency deviation of the sweep generator output. According to Equation 3:

$$f = \frac{2L}{V_g}\frac{F}{T}$$

Let f_0 be the nth harmonic and f_1 be the $(n + 1)$th harmonic. Then

$$f_1 = (n + 1)\frac{1}{T}$$

$$f_0 = n\frac{1}{T}$$

$$f_1 - f_0 = (n + 1)\frac{1}{T} - n\frac{1}{T} = \frac{1}{T}$$

or

$$f_1 - f_0 = \frac{2L_1}{V_g}\frac{F}{T} - \frac{2L_0}{V_g}\frac{F}{T} \qquad (7)$$

where L_1 is the distance corresponding to the $(n + 1)$th harmonic and L_0 is the distance corresponding to the nth harmonic. From Equation 7,

$$L_1 - L_0 = \frac{V_g}{2F}$$

Thus, the distance between harmonics is proportional to V_g, inversely proportional to F, and independent of the sweep generator repetition frequency. The group velocity of 3950 Mc in WR-229 waveguides is 7.45×10^8 ft/sec. Using a 500 Mc frequency deviation yields a distance between harmonics of 0.745 ft or approximately 9 inches. It is not possible to secure any desired small distance between harmonics by simply increasing the frequency deviation. Because the group velocity of the traveling waves varies with the r-f frequency, the instantaneous beat frequency due to a discontinuity varies over the course of a sweep generator cycle. The equation of this variation is:

$$V_g = c\sqrt{1 - \left(\frac{f_{ct}}{f_c}\right)^2} \qquad (8)$$

where c is the velocity of light, f_c is the instantaneous r-f frequency, f_{ct} is the

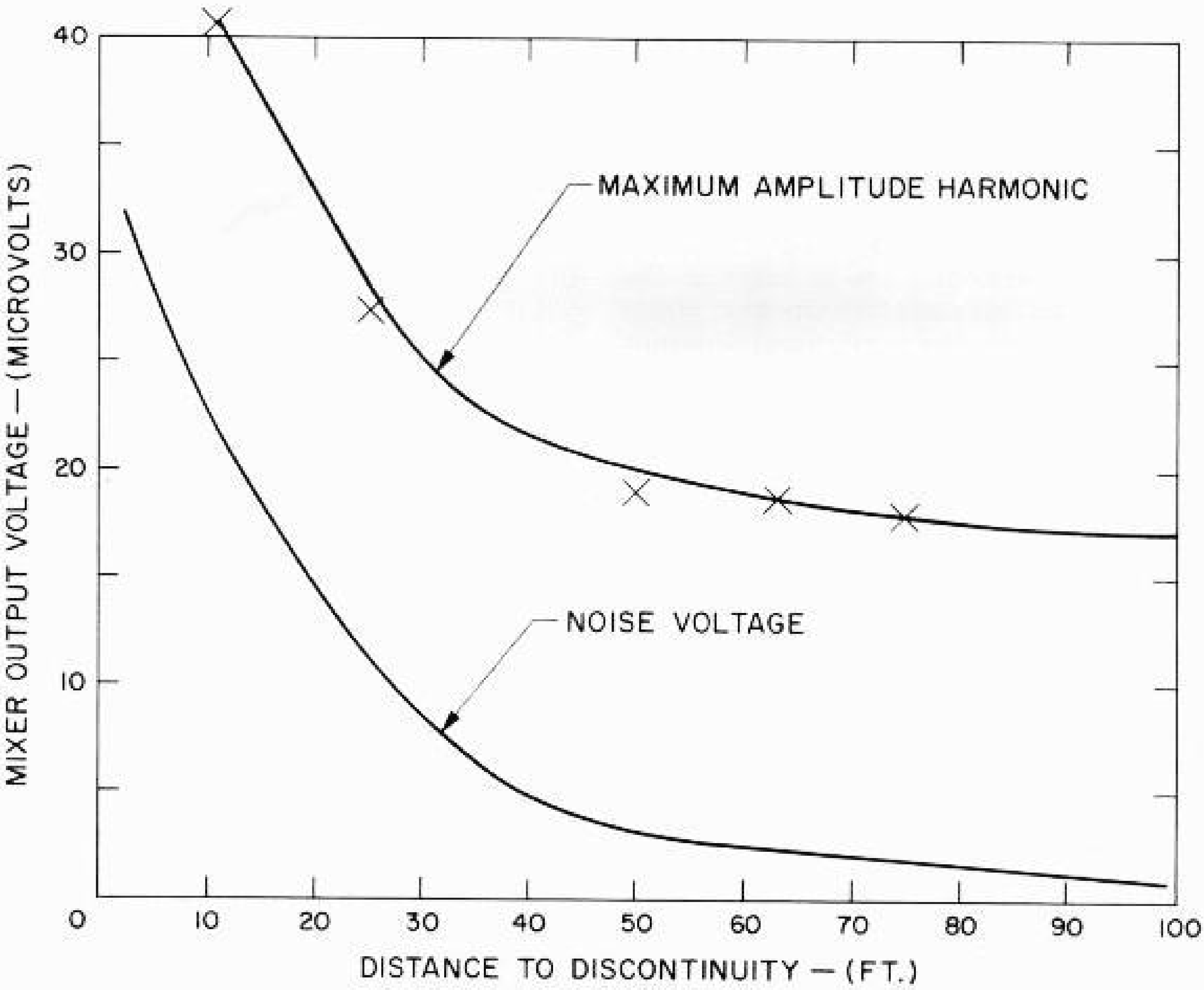


Figure 16. Calibration Curves for (a) Signal vs. Distance to a 1% Discontinuity (b) Noise vs. Distance

cutoff frequency of the waveguide, and V_g is the group velocity of the propagating waves. As the instantaneous r-f frequency decreases, so does the group velocity. From Equation 3, the beat frequency increases over the course of this cycle. The variation in beat frequency is shown in Figure 17. The change in beat frequency over the course of a sweep cycle increases with r-f frequency deviation and with distance to the discontinuity. The effect of this frequency modulation of the beat frequency is a decrease in amplitude of the maximum amplitude harmonic, and an increase in amplitude of the harmonics near the maximum amplitude harmonic. Using wide r-f frequency deviations, while observing discontinuities at large distances, can increase the frequency modulation of the beat frequency to the point where the adjoining harmonics are as large as the "maximum" amplitude harmonic. This greatly reduces the resolution. One possible solution to this problem is to control the r-f frequency deviation so as to prevent the change in beat frequency over the course of a sweep cycle from exceeding a fixed percentage of the sweep generator repetition frequency. The amount of change in beat frequency permissible was arbitrarily fixed as equal to the sweep generator repetition frequency.[2] This results in a frequency modulation index of 0.5 since the maximum change in the beat frequency from its mean value is then one-half the sweep generator repetition frequency. It has been pointed out that in a sinusoidally frequency-modulated wave of modulation index 0.5, 10 percent of its energy is in the sidebands.[2] Actually, the modulation in this case is far from sinusoidal—it is closer to linear, and permitting the total change in beat frequency to be equal to the sweep generator repetition frequency will result in two harmonics having equal or near equal amplitudes with the harmonics on either side then diminished in amplitude.

Since the change in beat frequency increases with distance to the discontinuity and with r-f frequency deviation, it is necessary when using this solution to lower the r-f frequency deviation with increasing distance to the discontinuity. But as shown above, lowering the r-f frequency deviation automatically lowers the resolution due to the resultant increased distance between harmonics. This solution is not satisfactory when measuring the individual magnitudes of reflections from joints one foot apart where these joints are 100 ft. or more from the measuring equipment.

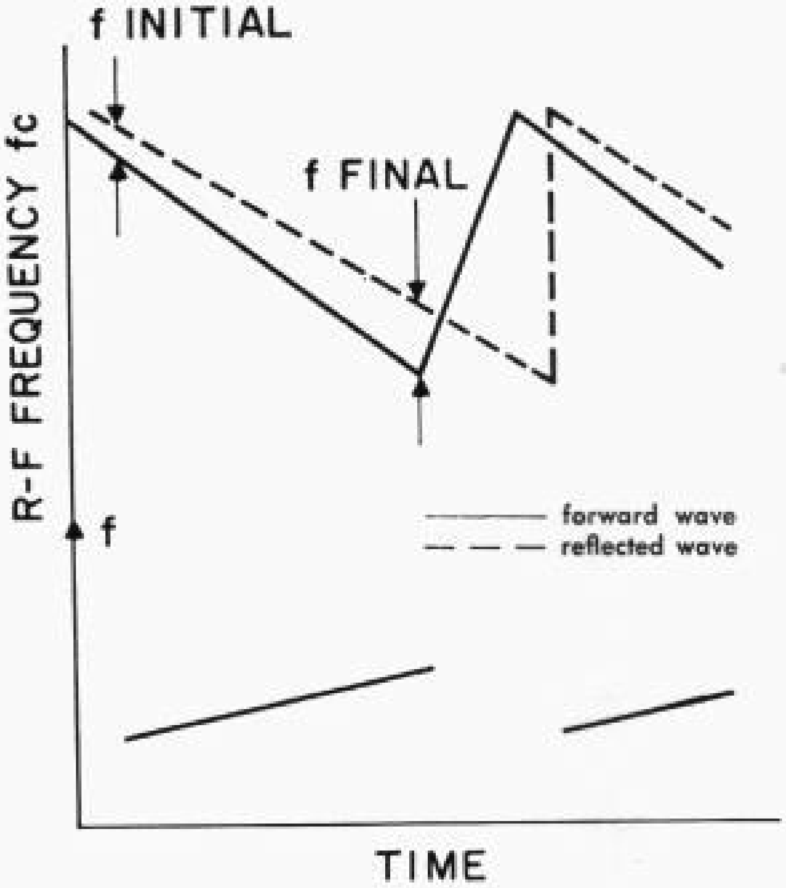


Figure 17. Variation of Beat Frequency due to Change of Vg with R-F Frequency

Another method for reducing the frequency modulation of the beat frequency is to use a non-linear frequency sweep. Figure 18 shows a heuristic motivation for such a sweep. τ increases with decreasing r-f frequency as shown in Figure 17, but because the tail of the frequency vs. time characteristic of the forward traveling wave is bent, the beat frequency remains more nearly constant over the modulation cycle. An approximate derivation of this curve is given in the Appendix. If a swept frequency which increased with time were used, the necessary modifications that would have to be made in the sweep characteristic to maintain a constant beat frequency are shown in Figure 19. The curve in Figure 18 is a decaying exponential while that in Figure 19 is an increasing exponential. Since the former is much easier to generate than the latter, a sweep generator whose output frequency decreases with time is used.

A linearly decreasing voltage applied to the helix of the BWO produces an exponentially decreasing output frequency. The commercial sweep generator uses a linear sawtooth waveform shaped by a series of diodes, each biased so as to produce the resultant voltage waveshape at the helix. This yields a linear r-f frequency vs. time characteristic at the BWO output. A frequency modulation input jack on the sweep generator enables the user to apply a voltage in series with the sweep's linear sawtooth, or in place of the sweep's linear sawtooth. Because of the diode shaping network, the output frequency vs. time characteristic should closely follow the voltage vs. time characteristic of the voltage applied at the frequency modulation input jack. Little work has been done on this latter approach of applying a voltage in place of the sweep's linear sawtooth; the former approach of applying a voltage in series with the sweep's linear sawtooth is particularly simple to effect. When the sweep's linear sawtooth is functioning, a sawtooth voltage, intended for application to the CRO's x-deflecting plates, and in exact time synchronism with the sawtooth applied to the diode shaping network, is available at a front panel connector. This sawtooth voltage is shaped by an R-C circuit, and applied to the frequency modulation input jack. In this way the non-linear frequency sweep of Figure 18 is approximated.

A specially shaped r-f frequency vs. time characteristic is required to generate a single beat frequency for a single discontinuity. If this characteristic varies in a random manner, such as the sweep output amplitude variations discussed above, the frequency modulation of the beat frequency would become very great, thus lowering the resolution. Some commercial sweep generators use external levelers which are amplifiers in a negative feedback loop, amplifying detected r-f energy and sending a correction voltage back to the anode to maintain constant output power over the frequency band being swept. This method works because the output power is a function of the

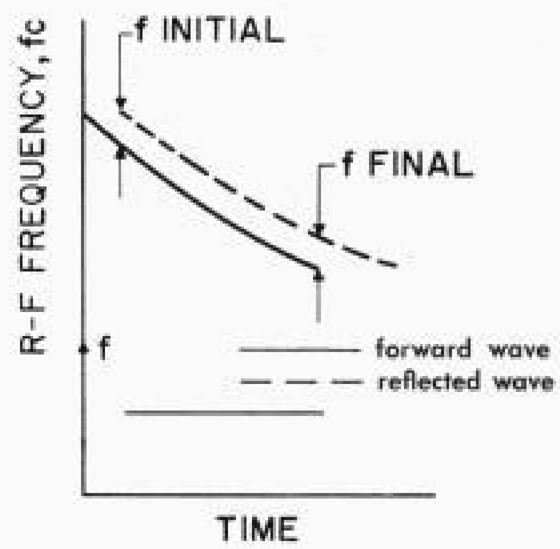


Figure 18. Constant Beat Frequency due to Non-Linear Sweep (R-F decreases with time)

anode voltage. But the output frequency is also a function of the anode voltage. Thus, the correction voltage maintains the output power constant but causes the instantaneous r-f frequency to vary from a nominally linear characteristic.

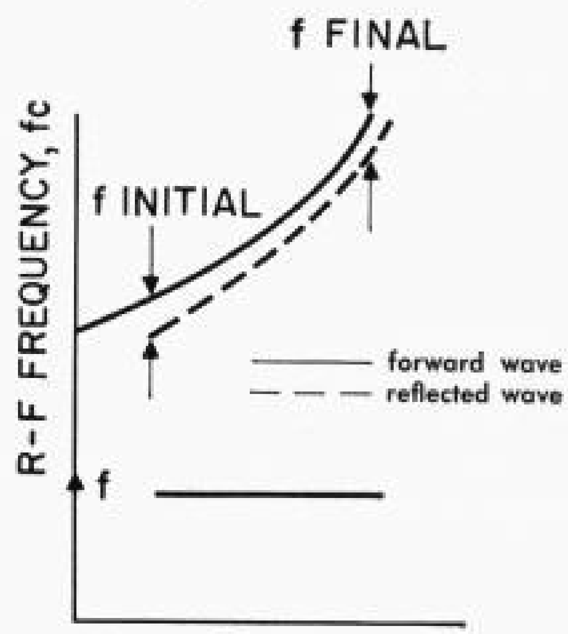


Figure 19. Constant Beat Frequency due to Non-Linear Sweep (R-F increases with time)

A simpler type of leveling arrangement is one where a programmed voltage, factory adjusted for the particular BWO tube installed, is applied to the anode in series with the normal anode voltage.

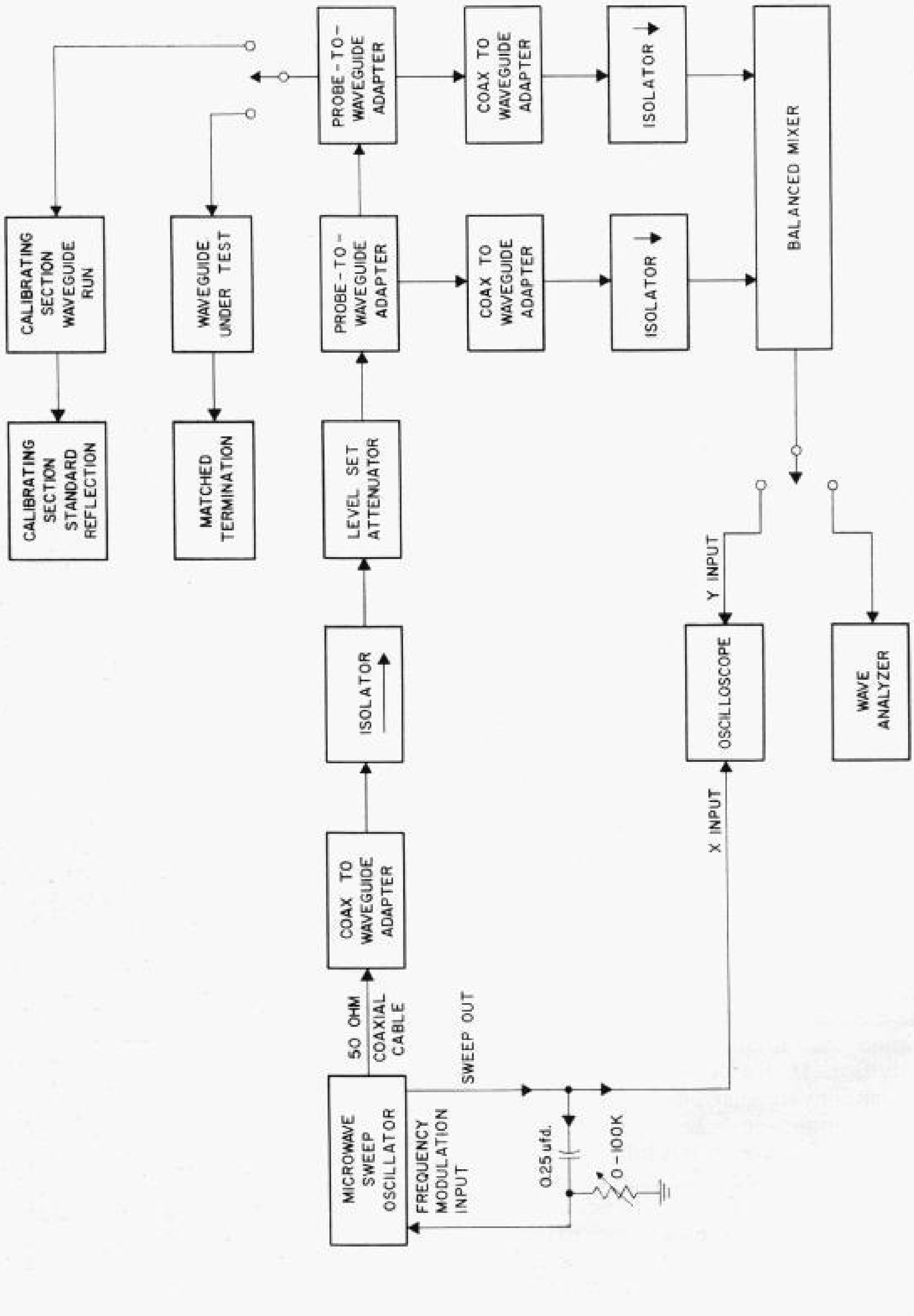


Figure 20. Overall Block Diagram of Reflection Measurement System.

This produces a flatter amplitude characteristic but changes the frequency characteristic to something other than linear—how much from linear depending upon the amplitude response of the particular BWO. This second type of leveling can usually be bypassed by a front panel control. Neither type of leveling can be used in this measurement method, nor can any type where the anode voltage is varied to correct for output amplitude variations.

Description of Equipment

A block diagram of the equipment used is shown in Figure 20. The insertion depth of the electric field probes, as well as the distance between probes, is adjustable. For low noise, this depth is kept as small as possible. The probes feed into coaxial cables, coax-to-waveguide adapters, isolators and the mixers. The length of the coaxial cables is kept to a minimum and their electrical lengths are matched. The isolators are indispensable for low noise.

The wave analyzer has a bandwidth of $\pm 3\frac{1}{2}$ cps and a frequency range of 50,000 cps. It can measure voltages in the five microvolt range. Because of the extremely narrow bandwidth of the wave analyzer, thermal noise is negligible and low noise factor crystals do not have to be used. The microwave sweep oscillator has the stable repetition frequency required to prevent the harmonic under observation from falling outside of the wave analyzer's narrow pass band.

The level-set attenuator can be adjusted for a convenient reference reading when using the calibration section. Its attenuation should, however, be kept as small as possible to permit a minimum probe insertion depth. The system as shown in Figure 20 is relatively low cost. The use of a higher cost precision waveguide attenuator (directly after the second probe-to-waveguide adapter) and shorting plate (at the end of the calibration section) to simulate a standard reflection permits a more convenient and flexible operation. This arrangement at 6-Kmc is shown in the photographs, Figures 21, 22 and 23.

The oscilloscope presentation in Figure 21 corresponds to a 26 db return loss with a coaxial cable calibration section and a coaxial short circuit. The end of the coaxial cable with its short circuit can be seen on top of the waveguide at-

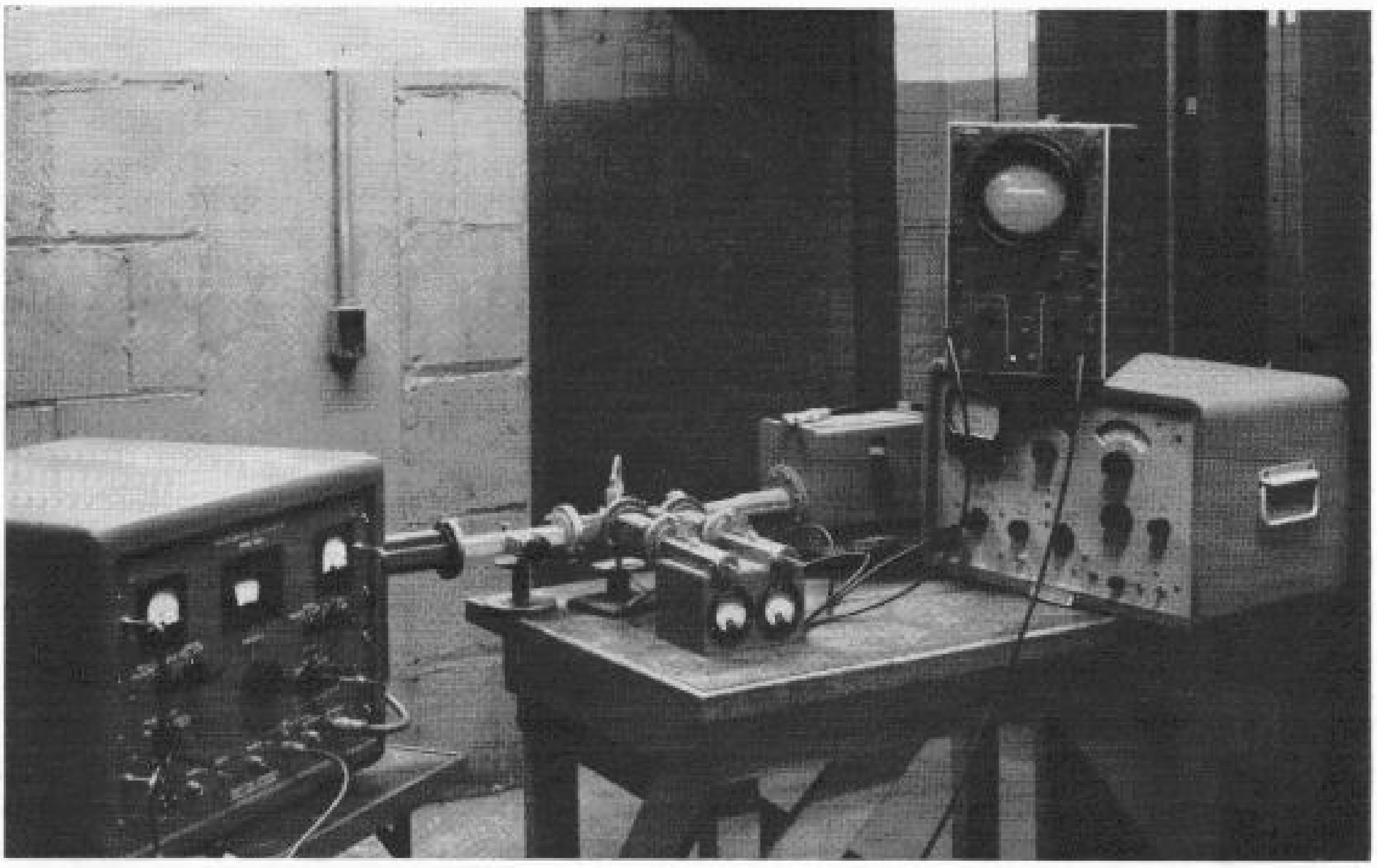


Figure 21. Arrangement of Equipment for Measurement of Discontinuities at 6 Kmc.

tenuator. The attenuation of the coaxial cable should be added to that of the precision waveguide attenuator when calibrating the equipment.

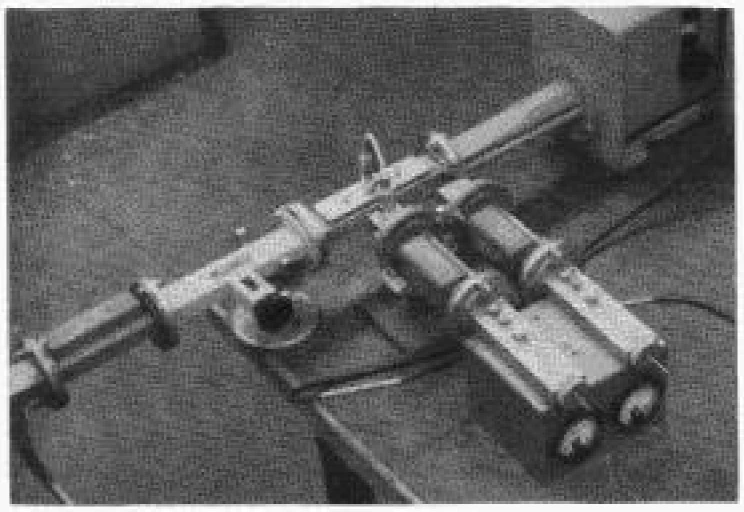


Figure 22. Detail View of Figure 21 (taken from above) showing Isolator, Level-set Attenuator, Balanced Mixer and Precision Attenuator.

The two probe-to-waveguide adapters are clearly shown in Figure 23. The close relative spacing of the two probes is obvious here.

Calibration

A 100-ft length of WR-229 waveguide with approximately 12 ft. spacing between readily accessible joints was constructed. A shorting plate was placed at different points within this run and a precision waveguide attenuator at the measurement equipment end of the run set to 20 db attenuation, thus simulating a 1.02 VSWR mismatch at the point where the shorting plate was located. (The attenuator reading must be doubled to obtain the inserted return loss since the traveling wave is attenuated twice, once in the forward direction and once in the backward direction.) With this arrangement, the shorting plate completely dominates other reflections in the waveguide run. The signal voltages due to the 1.02 VSWR at different distances is shown in Figure 16. The noise voltages are also shown in Figure 16. The fact that the signal voltage decreases with increasing distance, rapidly at first, and then more gradually, is due to (1) the waveguide attenuation, (2) the high audio-frequency losses from the mixer output to the wave analyzer input, and—(3) the frequency modulation of the beat frequency discussed above. Some of this frequency modulation does occur, despite the R-C compensating network because the exact wave shape required for zero frequency modulation is not generated. However, the compensating network does strongly peak the maximum amplitude harmonic. The latest measurements indicate that the maximum amplitude harmonic is at least 35 percent greater than the adjoining harmonics after peaking the maximum amplitude harmonic with the compensating network.

The calibration curve must be used if accurate results are to be obtained. A different calibation curve is required if the r-f frequency deviation is changed over a wide range, since the frequency modulation of the beat frequency is a function of frequency deviation. A different calibration curve is not required for each sweep generator in the field; the same curve suffices for all, provided that they use approximately the same r-f frequency deviation. Also, the same curve suffices when mixers, or mixer crystals are changed. An ordinate normalization at a single point compensates for these changes. It does not seem necessary to construct the calibration waveguide run to the length of the longest run to be measured, as it seems reasonable to assume that the curve can be extrapolated. No special care need be taken in the construction of the calibration waveguide run. As many bends can be used as are desired, and the joints do not have to be high quality since the reflection from the shortening plate overpowers all other reflections. Failure to use the calibration curve will result in an overly optimistic evaluation of the waveguide run under test. Using a calibration section approximately 15 ft. long results in a 6-db error on the measurement of reflections at the end of 100 feet of WR-229, when no calibration curve is used.

Operation

The output of the probe-to-waveguide adapter is terminated in a low VSWR load. The sweep generator is set to the correct frequency deviation and start frequency for the waveguide under test and the resolution desired. (A 375 Mc deviation with WR-229 yields a 1 ft. distance between harmonics.) The repetition frequency is set at 100 cps. The output

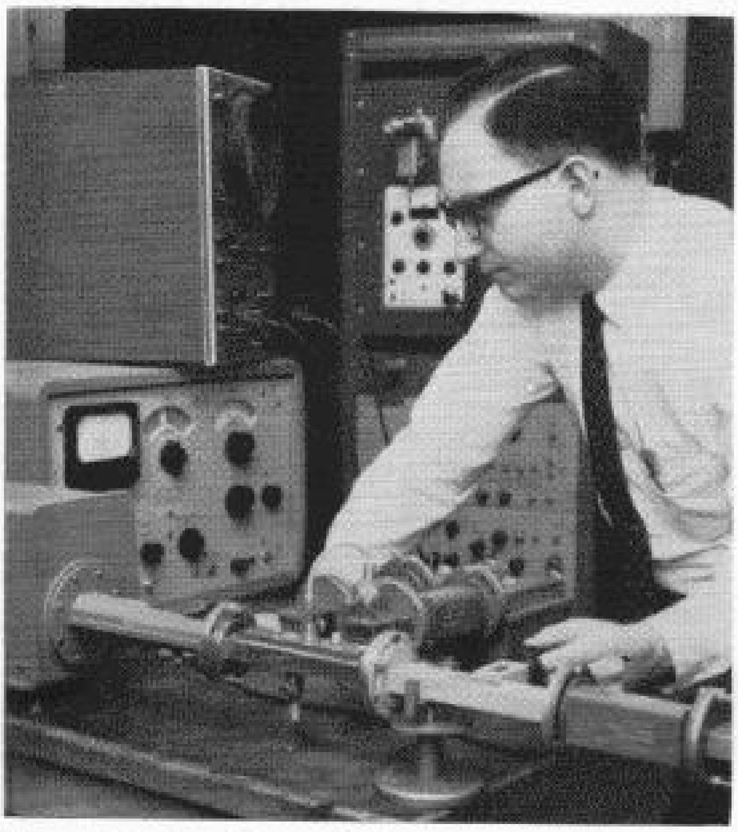


Figure 23. Rear View showing Location of Probes in Probe-to-Waveguide Adapter.

of the balanced mixer is applied to the wide-band amplifier and oscilloscope. The input from its probe to each mixer is in turn disconnected while the other mixer, with its input connected, has its tuning screws and end plate adjusted for maximum output. The inputs are then applied to both mixers together and the insertion depth of the two probes adjusted so that the output of the balanced mixer is zero, or near zero, during the entire sweep. Direct coupling should be used throughout the mixer, amplifier, and oscilloscope Y-axis amplifier when making this adjustment. Of course, complete removal of both probes yields a zero voltage output. Some experience with the equipment enables the user to know when too much depth of both probes is being removed. As a check that the probes are not too deeply inserted, the input to a mixer from one probe is disconnected and that probe slightly withdrawn. The output wave shape of the balanced mixer should show little or no change. The input to the mixer should be reconnected, the probe reinserted to the optimum depth and the check repeated for the other probe.

A calibration section of waveguide is attached to the measurement equipment and terminated in a standard waveguide mismatch, e.g. having a 1.1 VSWR. The calibration section length plus the length of waveguide in the measurement equipment between the middle of the two probes and the output flange is an integral number of feet, e.g. twelve feet. The output of the balanced mixer is then applied to the wave analyzer. The wave analyzer (whose zero set should be checked periodically) is set to 100 cps times the overall length of the calibration section or 1200 cps.

Operating Procedure

The procedure for operating the equipment is as follows:

- Slightly adjust the sweep generator repetition frequency control to peak the wave analyzer reading.
- Adjust the relative spacing of the two probes for maximum wave analyzer response.
- Slightly adjust the sweep generator frequency deviation control for maximum wave analyzer response.
- Peak the wave analyzer response using the R-C compensating circuit with the rheostat at or near its maximum resistance point.
- Check that the wave analyzer response at 1100 and 1300 cps is below the response at 1200 cps.

The ordinate scale of the curve of Figure 16 is changed so that the response at 12 ft. (or whatever the calibration section length is) is one-fifth the wave analyzer reading at 1200 cps. The calibration section should be removed and a good load placed at the measurement equipment output flange. The voltage reading at 1200 cps (a noise voltage) should then be less than one-tenth the wave analyzer reading at 1200 cps.

Connect the waveguide under test to the measurement equipment output flange. Vary the wave analyzer frequency, noting those maximum amplitude harmonics whose values approach the curve of Figure 16. Peak the readings of these maximum amplitude harmonics using small changes in the r-f frequency deviation control and the rheostat of the compensating network. Note the frequencies of those maximum amplitude harmonics whose magnitude exceeds that of Figure 16. If none are found, the run is satisfactory (assuming that a 1 percent reflection is deemed the maximum tolerable). The harmonics that do exceed the values of Figure 16 will generally be found to correspond to joints. These joints should be visually checked for loose flanges, improper bolt tightening (too little or too much torque), missing alignment pins, and the like. After making the necessary corrections, the harmonics corresponding to these joints are rechecked.

A variation in the above procedure is one using the entire waveguide run under test as the calibration section. This requires that the length of the run be known beforehand. The R-C network peaks the response due to the mismatch at the far end of the waveguide run under test and is not touched thereafter. No effort is made to obtain the precise 100 cps/ft calibration as is done above and a slide rule is used to calculate the distance to the maximum amplitude harmonics.

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3. *Engineering of Microwave Systems*, W. H. Francis, L. A. Byam, R. E. Greenquist, J. J. Lenehan, E. H. Mueller, W. B. Sullinger, and C. B. Young, *Western Union Technical Review*, Vol. 14, No. 3, July 1960.
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Mr. Ethan Aronoff has been associated with the Radio Systems Division in the Microwave group for almost three years. This group has recently become part of the Radio Beam Expansion Division. He has been working actively on intermediate frequency preamplifiers, intermediate frequency amplifiers, automatic gain control circuitry and the echo measurement system described in this article.

He received his B.E.E. degree from City College of New York in 1960 and is currently attending the Polytechnic Institute of Brooklyn studying for his M.S.E.E. degree. He is a member of Eta Kappa Nu and IEEE.

APPENDIX

Derivation of optimum r-f frequency vs. time characteristic for minimum frequency modulation of beat frequency.

$$f = \frac{2L}{V_g}\frac{df_c}{dt} \qquad (4)$$

$$V_g = c\sqrt{1-\left(\frac{f_{ct}}{f_c}\right)^2} \qquad (8)$$

For a given L, a constant f is desired,

$$\frac{df_c}{dt}\frac{1}{c\sqrt{1-\left(\frac{f_{ct}}{f_c}\right)^2}} = K, \textit{ a constant} \qquad (9)$$

This differential equation has the solution:

$$f_c = \sqrt{f_{ct}^2 + (Kct + C)^2} \qquad (10)$$

where C is an arbitrary constant

An approximation to (10) is

$$f_c = f_0\, e^{-at} \qquad (11)$$

then,

$$f_c^2 = f_0^2 e^{-2at} \qquad (12)$$

Expanding (12):

$$f_c^2 = f_0^2\,[1 - 2\,at + 2(at)^2 - \ldots] \qquad (13)$$

From (10):

$$f_c^2 = f_{ct}^2 + C^2 + 2KcCt + K^2c^2t^2 \qquad (14)$$

Equating coefficients of like powers to t in (13) and (14):

$$f_0^2 = f_{ct}^2 + C^2 \qquad (15)$$

$$-\,2\,a\,f_0^2 = 2KcC \qquad (16)$$

a and f_0 are determined from (11) and the desired swept frequency range: To sweep from 4200 Mc to 3700 Mc at a 100 cps repetition rate

$$f_0 = 4200\ Mc$$

From (11), with $f_c = 3700$ Mc and $t = 0.01$ seconds

$$4200e^{-a(0.01)} = 3700$$

$$a = 12.8\,(\text{seconds})^{-1}$$

f_{ct} determined from waveguide size used; the f_{ct} for WR-229 is 2.577 kMc, therefore (15) and (16) can be solved for K, C:

$$C = 3330\ mc$$

$$K = -\ 226.3\ cycles/meter\text{-}second$$

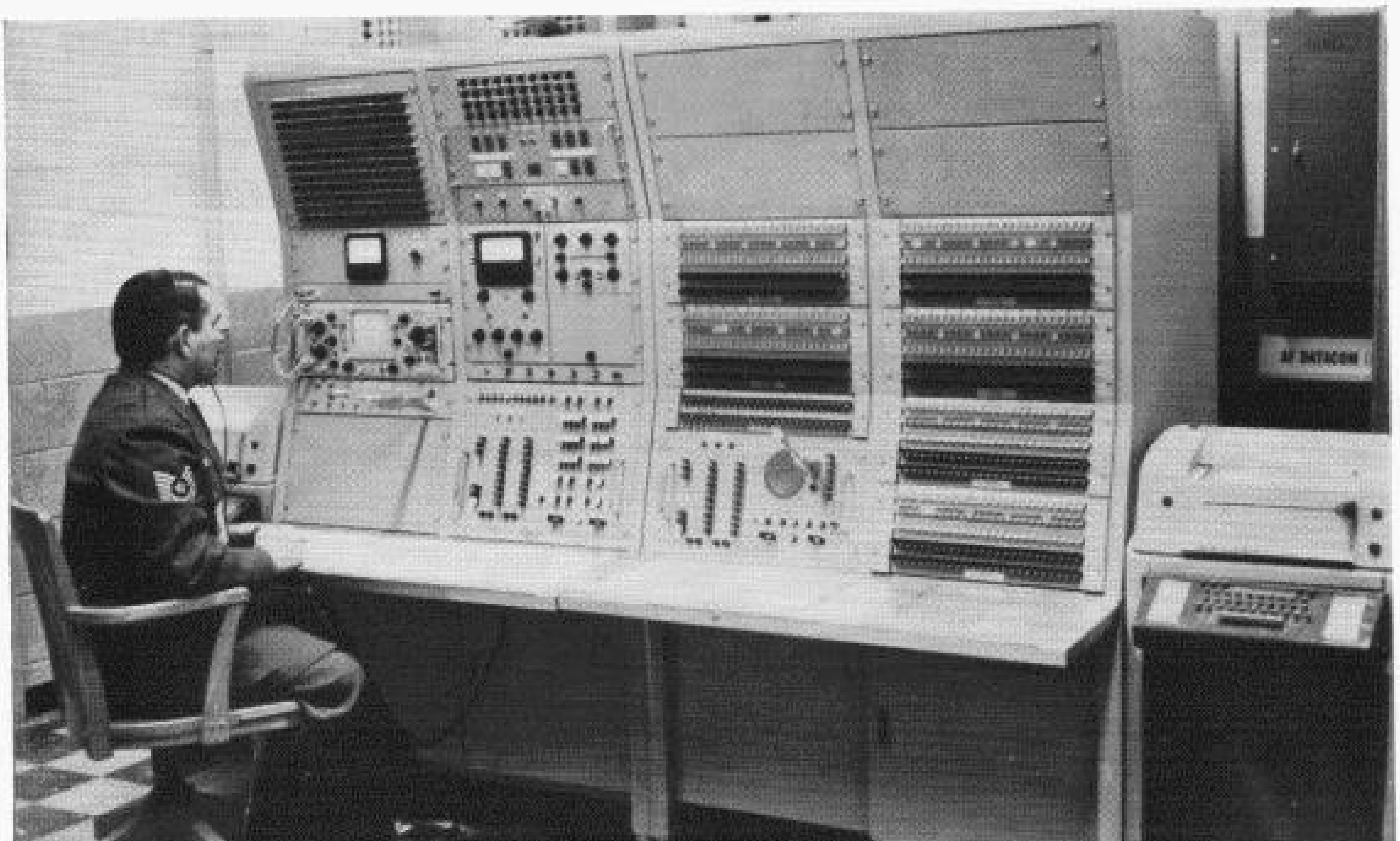
At $t = 0$, both (10) and (11) yield the value $f_c = 4200$ Mc At $t = 0.01$ (end of sweep), (10) yields the value $f_c = 3689$ Mc while (11) yields the value $f_c = 3700$ Mc.

Thus, (11) is a reasonable approximation of (10) and the decaying exponential r-f frequency vs. time characteristic gives a close to minimum frequency modulation of the beat frequency.

AF DATACOM System Completed

Andrews AFB, the last of five special automatic electronic switching centers for the AF DATACOM network, was turned over to the Air Force on February 27, 1963. This is the fifth AF DATACOM operation of the world's largest and most advanced data communications network. The first center at Norton AFB was completed in November 20, 1962. Then McClellan AFB, Tinker AFB, Gentile AS followed in rapid succession. In its present configuration, the system can transmit 100 million words daily from punched cards and perforated tape with an accuracy such that only one error will remain undetected in 10 million characters.

AF DATACOM, originally designed for the Air Force has been renamed AUTODIN, *auto*matic *dig*ital *net*work for use by the Department of Defense.



Monitor Console and Crypto Console in the Technical Control Area at Tinker AFB, Oklahoma. This console was completely designed and built by Western Union

R. L. Ernst and J. K. Fitzpatrick, Engineers, Radio Beam Expansion Project

VARACTOR DIODE

Part II–Applications to Microwave Systems

In Part I of this article, which appeared in the January 1963 issue of the Western Union TECHNICAL REVIEW, the voltage variable capacitance of the varactor diode was described. Equivalent circuits were presented, and the Manley-Rowe power relations were introduced. The basic varactor parameters used in circuit synthesis were also defined.

In Part II various applications which illustrate the potential of the varactor in modern circuitry are outlined.

Frequency Multipliers

One of the most common applications of the varactor is the frequency multiplier, a device which converts power at one frequency into power at an integral multiple of that frequency.

From the Manley-Rowe relations for nonlinear reactances:

$$\frac{P_{in}}{f} + \frac{n\,P_n}{n\,f} = 0 \qquad (14a)$$

where P_{in} = power input at f,
P_n = power output at nf.

Thus, efficiency

$$\eta = -\frac{P_n}{P_{in}} = \frac{n\,f}{n\,f} = 100\% \quad . \qquad (14b)$$

Actually efficiency in varactor multipliers is less than 100 percent, because of:

a) losses in circuit elements,
b) series resistance of the varactor,
c) small but finite amounts of power dissipated at harmonics other than the desired output frequency.

The Doubler

The simplest frequency multiplier is the multiply-by-two circuit or doubler. It has two possible configurations of circuit elements; the most common, shown in Figure 14 is referred to as a shunt or charge-controlled type doubler. The term "charge-controlled" is derived from the fact that the varactor is supplied with sinusoidal current to develop non-sinusoidal voltage. The filters indicated are ideal pass devices, i.e. they present short circuits to the frequency of interest and open circuits to all other frequencies. The

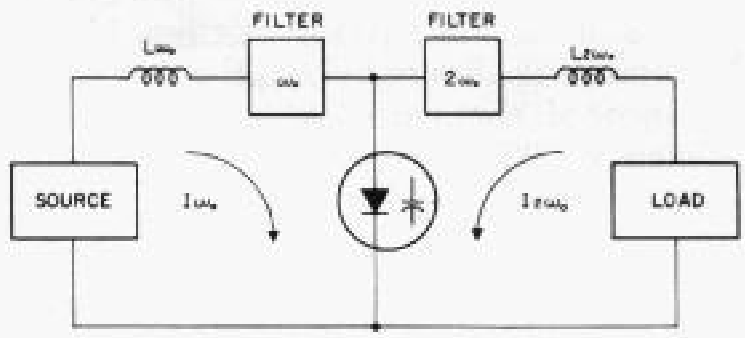


Figure 14. Ideal Charge-Controlled Doubler

inductances L_{ω_0} and $L_{2\omega_0}$ are resonated with the effective capacitance of the varactor so that maximum currents can flow in both the ω_0 and $2\omega_0$ loops.

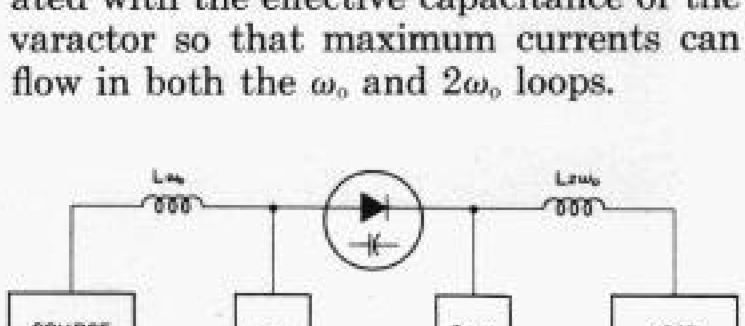


Figure 15. Ideal Voltage-Controlled Doubler

Figure 15 illustrates the series or "voltage-controlled" doubler in which a sinusoidal voltage is impressed across the varactor to develop non-sinusoidal currents through it. Ideal reject filters are utilized in this configuration to short circuit all frequencies except the frequency of interest, which is open-circuited. The inductances serve the same function as the analogous devices in the shunt doubler.

Series vs. Shunt Configurations

The charge-controlled doubler has several significant advantages over the series type, both theoretical and practical. With an abrupt junction varactor the shunt configuration can develop only the second harmonic of voltage when a sinusoidal current flows through it. For $\gamma < \frac{1}{2}$, where γ is the exponent in Equation 6, amplitudes of voltages at harmonics of the current frequency in excess of two are small. However, the series doubler can generate currents at all harmonics of the input sinusoidal voltage with any available junction characteristic. These currents can be supported by virtue of the short circuit at all frequencies other than at ω_o and $2\omega_o$. The power at these higher harmonics is dissipated in the series resistance of the varactor, thus reducing efficiency. This problem does not arise with the shunt multiplier since the varactor is open circuited at all frequencies except ω_o and $2\omega_o$. Of practical importance is the ease of circuit synthesis. The complex design of "ideal" reject filters, necessary in a series configuration, can be overwhelming. As will be demonstrated, the transformation of ideal to practical is readily accomplished when operating with the charge-controlled device.

Lumped Parameter Doubler Design

To illustrate the principles of varactor multipliers, the design of a typical charge-controlled multiplier is described. Desired output frequency is of primary interest in the design of such multipliers. It determines the class of circuit components, either lumped, coaxial, or waveguide. Units with output frequencies in excess of 800-1000 Mc must be developed by means of microwave techniques.

For this illustration a 100 Mc doubler will be synthesized, selecting a varactor with a high P_{norm} and attempting to attain the maximum power output at maximum efficiency. The voltage across the device will be restricted to the limits of ϕ and V_B.

The following approximations apply to the design of maximum-efficiency, low-frequency ($\omega_o << \omega_c$) multipliers utilizing abrupt junction varactors:

Efficiency	$\eta = 1 - 20 \dfrac{\omega_o}{\omega_c}$	(15a)
Source Impedance (optimum)	$R_{in} = \dfrac{0.08}{\omega_o C_{min}}$	(15b)
Load Impedance (optimum)	$R_{out} = \dfrac{0.14}{\omega_o C_{min}}$	(15c)
Maximum Input Power	$P_{in} = 0.03\, P_{norm} \dfrac{\omega_o}{\omega_c}$	(15d)
Average voltage (across varactor)	$V_o = 0.35\, V_B$	(15e)

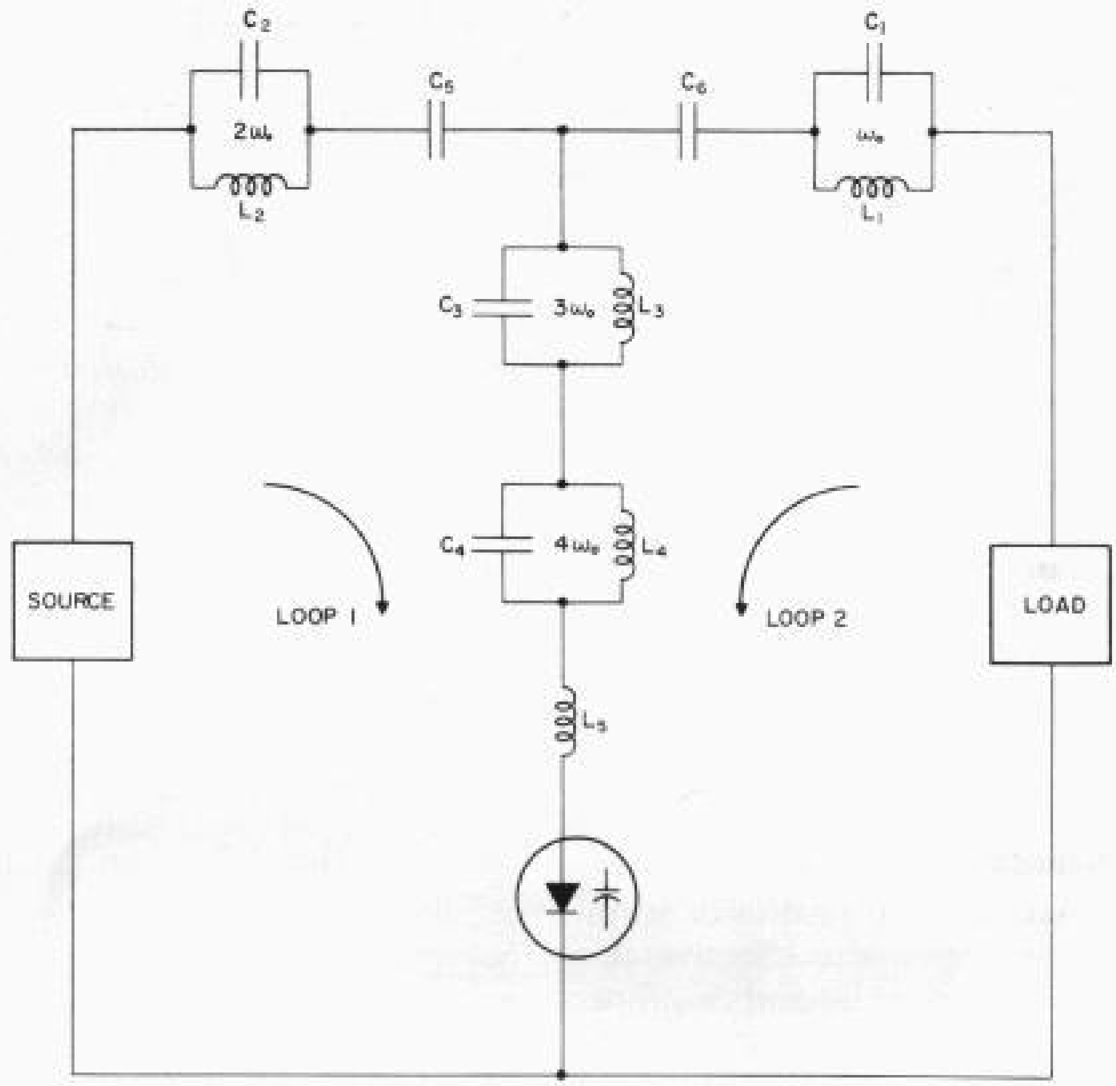


Figure 16. A Practical Charge-Controlled Doubler

In all formulas, $\omega_0 = 2\pi f_0$ (operating input frequency). The actual circuit configuration to be utilized is shown in Figure 16; the actual unit is shown in Figure 17. This design can be manipulated to achieve the following conditions:

1) loop 1 can be tuned to series resonate at ω_0 (maximum input power transferred to varactor).
2) loop 2 can be tuned to series resonate at $2\omega_0$ (maximum output power transferred to load).
3) no current may flow at $2\omega_0$ in loop 1 (L_2C_2 forms a parallel resonant trap at $2\omega_0$).
4) no current may flow at ω_0 in loop 2 (L_1C_1 forms a parallel resonant trap at ω_0).
5) the varactor can be open-circuited at harmonics in excess of two.

With an abrupt junction varactor, no current can be supported at harmonics in excess of four. A graded junction varactor will convert power to higher harmonics but power levels at harmonics greater than the fourth are of such small magnitude that only $3\omega_0$ and $4\omega_0$ resonant traps are necessary.

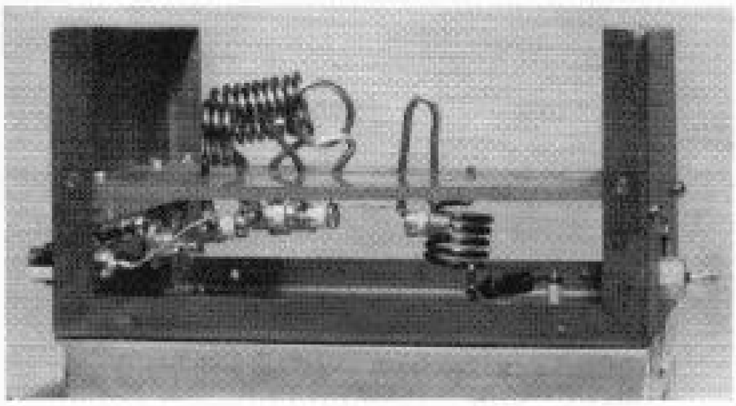


Figure 17. Actual VHF Doubler

These conditions can be developed into the following equations:

$$X_{2\omega_o TRAP} + X_{3\omega_o TRAP} + X_{4\omega_o TRAP} + X_{c5} + X_{L5} + X_{VAR} = 0 \text{ at } \omega_o \qquad (16a)$$

or

$$\frac{4}{3}\,\omega_o L_2 + \frac{9}{8}\,\omega_o L_3 + \frac{16}{15}\,\omega_o L_4 - \frac{1}{\omega_o C_5} + \omega_o L_5 + X_{VAR(\omega_o)} = 0 \qquad (16b)$$

$$X_{\omega_o TRAP} + X_{3\omega_o TRAP} + X_{4\omega_o TRAP} + X_{c6} + X_{L5} + X_{VAR} = 0 \text{ at } 2\omega_o \qquad (17a)$$

or

$$\frac{-2}{3}\,\omega_o L_1 + \frac{18}{5}\,\omega_o L_3 + \frac{8}{3}\,\omega_o L_4 - \frac{1}{2\omega_o C_6} + 2\omega_o L_5 + X_{VAR(2\omega_o)} = 0. \qquad (17b)$$

Effective Reactance

At this point in the synthesis it is important to determine the effective reactance of the varactor at ω_0 and $2\omega_0$. For an abrupt junction varactor with sinusoidal current drive, there will be a sinusoidal elastance variation.

Thus,

$$S_{avg} = \frac{1}{2}(S_{max} - S_{min}) \qquad (18)$$

If S_{min} is assumed to be zero, then

$$S_{avg} = \frac{1}{2}\, S_{max} \qquad (19)$$

Thus, the effective reactance is equal to:

$$X_{VAR(\omega_o)} = \frac{-1}{2\omega_o C_{min}} \text{ at } \omega_o \qquad (20a)$$

$$X_{VAR(2\omega_o)} = \frac{-1}{4\omega_o C_{min}} \text{ for } 2\omega_o \qquad (20b)$$

where $C_{min} = C$ at breakdown.

Considering the shunt capacitance associated with the case, these equations become:

$$X_{VAR(\omega_o)} = \frac{-1}{\omega_o(2C_{min} + C_{case})} \qquad (21a)$$

$$X_{VAR(2\omega_o)} = \frac{-1}{2\omega_o(2C_{min} + C_{case})}. \qquad (21b)$$

Assuming the following parameters for a varactor:

P_{norm}	$= 15{,}000$ watts
V_B	$= 120$ volts
C_{min}	$= 8$ pf
ω_c	$= 130 \times 10^9$ rps (20.8 Gc)
C_{case}	$= 1.5$ pf
If ω_o	$= 628$ Mrps. (100 Mc) then
$\eta = 90\%$	from 15a
$R_{in} = 16\Omega$	from 15b
$R_{out} = 30\Omega$	from 15c
$P_{in} = 2.2$ watts	from 15d
$V_0 = 42$ volts	from 15e
$X_{VAR\omega_o} = -90\Omega$	from 21a
$X_{VAR2\omega_o} = -45\Omega$	from 21b.

To approach the theoretical efficiency, circuit inductances and capacitances should have high Q's. Equations 16b and 17b may be satisfied by using the following typical set of parameters:

L	nh	C	pf
L_1	500	C_1	5
L_2	125	C_2	5
L_3	60	C_3	5
L_4	30	C_4	5
L_5	200	C_5	8
		C_6	8

This design assumes that both load and source impedances, with resistances equal to R_{out} and R_{in} respectively, contribute zero reactance. In the practical case, this may be accomplished by transformers and matching elements.

Bias Considerations

An average voltage or self bias, V_0, is developed across the varactor when fully driven. (The input power may be computed from equation 15d.) This is attributed to the clamping action of the varactor. As voltage is developed from the sinusoidal current, it will be non-sinusoidal in form and result in a non-zero dc component. However, to develop maximum sinusoidal current the input and output loop reactances must equal zero. If power other than P_{in} is applied, the effective reactance due to the varactor will not correspond to Equations 21a and 21b and the loops will be mistuned. This can create a hysterisis effect.

To avoid this critical dependence on input power a fixed bias may be applied via external means. However, if V_0 is applied before input power, then C at V_0 does not equal the effective C under drive and the loop will again be mistuned. Therefore, it would be most effective to apply a voltage V_1 through a large resistance where C at V_1 corresponds to the effective C under drive. If the varactor is an abrupt junction device, $V_1 = \frac{V_B - \phi}{4}$. Then there is no critical dependence on either P_{in} or the time of application of P_{in}. Yet V_0 can be developed across the varactor by virtue of the large resistor (>10K). This method of biasing is termed "soft biasing."

Overdriving

Higher power outputs and slightly higher efficiencies may be attained by overdriving. (This occurs when input power exceeds the value of P_{in} calculated from Equation 15d.) However, avalanche and shot noise output increases significantly. Unfortunately no simple theoretical circuit model has been accepted for the varactor in the overdriven state. Thus, the development of an overdriven multiplier is largely an empirical modification of a conventionally driven device. Results of studies in this field have indicated that input power can exceed the theoretical maximum by a factor of 5 to 7 and still not destroy the varactor. This is generally attributed to a failure of the diode conduction mechanism at high frequencies. If power inputs exceeding 5 P_{in} are necessary and varactors with higher $\frac{P_{norm}}{\omega_c}$ are unavailable, multiple varactors can be arranged in push-pull and bridge configurations within a single multiplier stage.

High-Frequency Doublers

Theoretical efficiencies approach 40% as the input frequency approaches one-tenth of the cutoff frequency. In this frequency range:

$$P_{in} = 0.5\, P_{norm} \left(\frac{\omega_0}{\omega_c} \right)^2 \qquad (22)$$

As before, if the power handling ability of a single varactor circuit is not adequate, push-pull and bridge circuits can be used. However, at higher frequencies it becomes difficult and expensive to obtain the matched varactors required for this circuitry.

High-Order Multipliers

A tripler or other high-order multiplier can utilize the same shunt configuration as the doubler with one significant modification. Since the shunt configuration minimizes the production of voltages at harmonics of current frequency in excess of two the output loop cannot merely series resonate at nf. It must provide for maximum currents through the varactor at (n−x)f. For example, in a tripler the varactor must be provided with a shunt tank series resonant at 2f. This shunt circuit is described as an idler at the second harmonic. In this application "idler" refers to the fact that the current is temporarily stored at a particular frequency. However, no real power is allowed to be dissipated in these shunt resonant idlers, as is the case with the idler of the four-frequency upconverter mentioned in Part I. In common high order multipliers, the output frequency (f_n) relates to the input and idler frequencies in one of the following ways:

1) f_n is a sum of two frequencies (input plus idler or one idler plus another).
2) f_n is twice any idler frequency.

Thus an octupler may have the following idler configurations:

a) 1–2–4–8,
b) 1–2–3–5–8,
c) 1–2–3–6–8.

Single Stage High Order Multipliers vs. Cascaded Doublers

It is a common fallacy that a cascade of doublers is always more efficient than a single-stage, high order multiplier. It is true that efficiency of a single stage decreases with the order of multiplication. But at low frequencies, the efficiency of a single high order stage is only slightly less than the overall efficiency of a series of low order devices. However, when f approaches 1/10 f_c, the multiple doubler is better than the single stage. There are disadvantages to the high order multiplier scheme even at lower frequencies, such as

1) the circuit configuration becomes more difficult to synthesize as the order of multiplication increases.
2) the entire power loss of the multiplying process is required to be dissipated in a single varactor.

The advantages of high order circuits are the elimination of varactors which cost $35 - 150 each and the elimination of circuit losses due to interstage coupling.

Upper-Sideband Upconverters

An upper-sideband upconverter, shown in Figure 18 and commonly abbreviated as usbuc, mixes power at a given signal frequency, f_s, with power at a higher frequency known as the pump frequency, f_p, to give useful output at the sum of these frequencies, known as the upper-sideband frequency, f_u. This relationship is shown

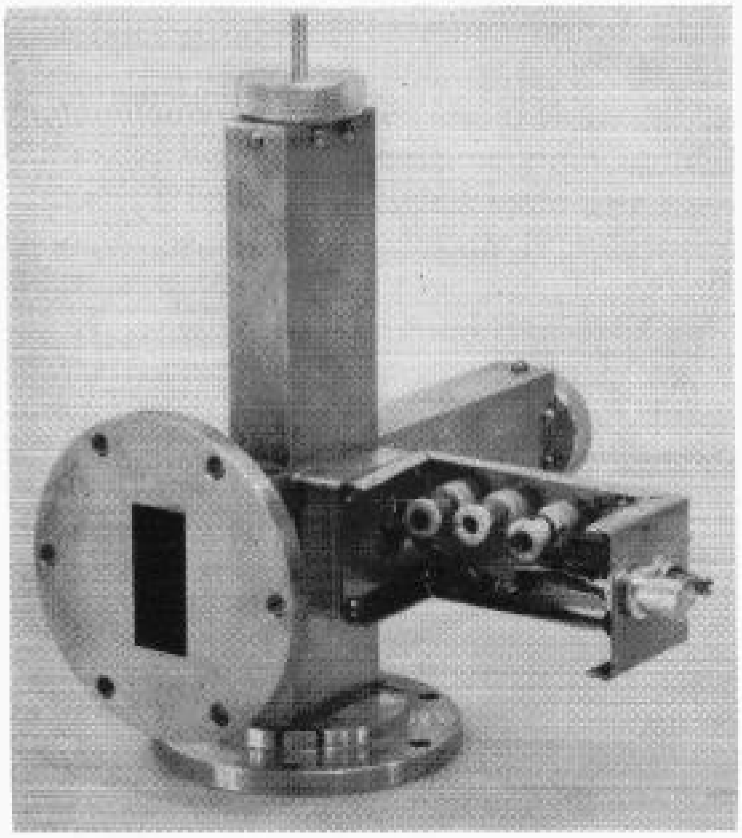


Figure 18. Actual Upconverter (70—6000 Mc)

in Figure 19. The pump power is not shown as an input in the same manner as the signal power because the signal power is much smaller than the pump power.

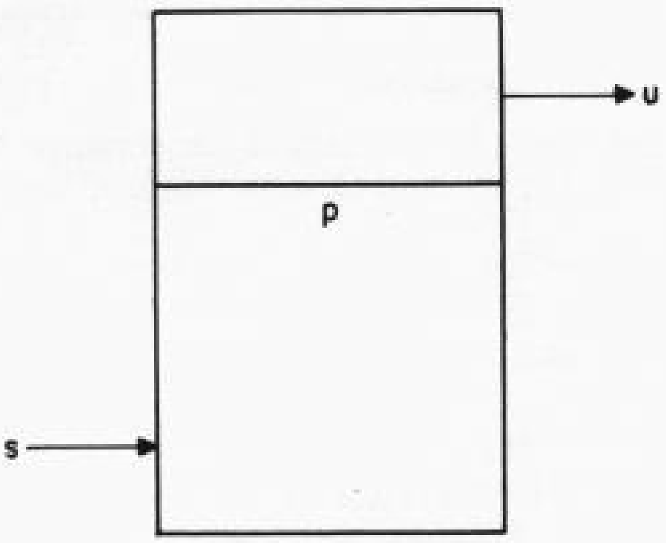


Figure 19. Representation of an Upper-Sideband Upconverter

To the signal the varactor appears as a capacitance varying periodically at the pump frequency rate.

To apply the Manley-Rowe power relations to the usbuc, the values of m and n required are shown in Table II:

which $\omega_c = \infty$, the above expression reduces to the Manley-Rowe gain as given by Equation 25.

TABLE II

Circuit Frequency	m	n	Manley-Rowe Frequency
f_s	1	0	f_1
f_p	0	1	f_2
f_u	1	1	f_1+f_2
$-f_s$	-1	0	$-f_1$
$-f_p$	0	-1	$-f_2$
$-f_u$	-1	-1	$-f_1-f_2$

The two Manley-Rowe equations become:

$$\frac{P_{1,0}}{f_1} + \frac{P_{1,1}}{f_1 + f_2} = 0 \qquad \text{(23a)}$$

$$\frac{P_{0,1}}{f_2} + \frac{P_{1,1}}{f_1 + f_2} = 0. \qquad \text{(23b)}$$

In terms of s, p, and u these become:

$$\frac{P_s}{f_s} + \frac{P_u}{f_u} = 0 \qquad \text{(24a)}$$

$$\frac{P_p}{f_p} + \frac{P_u}{f_u} = 0. \qquad \text{(24b)}$$

The first equation can be used to solve for gain:

$$g = -\frac{P_u}{P_s} = \frac{f_u}{f_s}. \qquad \text{(25)}$$

That is, the upconverter, using a perfect nonlinear capacitance, provides a gain equal to the ratio of output to input frequency as well as shifting the information from one frequency to another.

When a lossy varactor is used in a usbuc configuration which is tuned for maximum gain, the gain is expressed as:

$$g = \frac{\left(\dfrac{m_1\,\omega_c}{\omega_s}\right)^2}{\left(1 + \sqrt{1 + \dfrac{m_1^2\,\omega_c^2}{\omega_s\,\omega_u}}\right)^2}. \qquad \text{(26)}$$

It should be pointed out that, if a perfect nonlinear capacitance is used, i.e. one in which $\omega_c = \infty$, the above expression reduces to the Manley-Rowe gain as given by Equation 25.

Certain requirements have to be satisfied by the pump circuit. As shown in Figure 9 of Part I of this article, the varactor acts as a nonlinear capacitance only between the voltages ϕ and V_B. Therefore, the pump voltage must not at any time extend beyond these limits. It can be seen that a negative bias voltage will be needed when the pump voltage swings between the limits set by ϕ and V_B. For example, a sinusoidal pumping voltage will require a bias voltage of $-\frac{\phi - V_B}{2}$, or half way between the voltages ϕ and V_B. The pump power required to fully cover the entire capacitance swing of the varactor for properly-biased sinusoidal voltage pumping is expressed as:

$$P = 0.267\,P_{norm}\left(\frac{\omega_p}{\omega_c}\right)^2 \qquad \text{(27)}$$

for graded-junction varactors and as:

$$P = 0.500\,P_{norm}\left(\frac{\omega_p}{\omega_c}\right)^2. \qquad \text{(28)}$$

for abrupt junction varactors.

In each case, the pump power required for full pumping is directly proportional to P_{norm}. Therefore, when two varactors are available and have comparable performance characteristics, it is usually preferable to select the varactor having the lower value of normalization power. This means that less pump power is required.

A typical 70 Mc to 6000 Mc usbuc is shown in Figure 18.

Lower-Sideband Upconverters

A lower-sideband upconverter (abbreviated as lsbuc) is shown in the schematic representation in Figure 20.

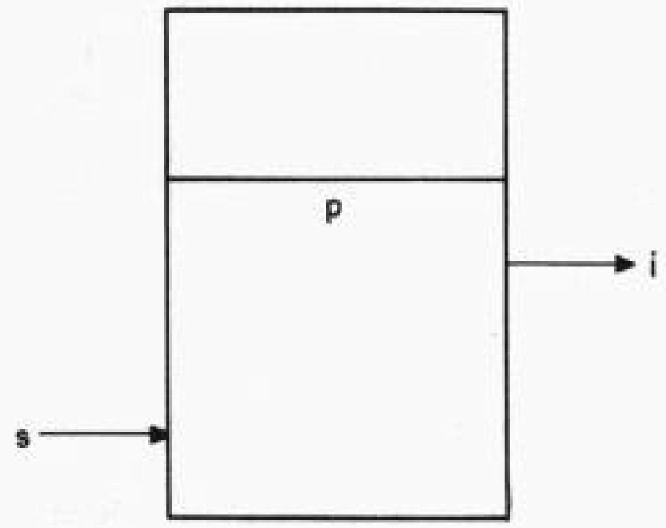


Figure 20. Representation of a Lower-Sideband Upconverter

The Manley-Rowe parameters of m and n for the upconverter are shown in Table III:

TABLE III

Circuit Frequency	m	n	Manley-Rowe Frequency
f_s	1	0	f_1
f_p	0	1	f_2
f_i	-1	1	$f_2 - f_1$
$-f_s$	-1	0	$-f_1$
$-f_p$	0	-1	$-f_2$
$-f_i$	1	-1	$f_1 - f_2$

The power relations become:

$$\frac{P_{1,0}}{f_1} + \frac{P_{1,-1}}{f_1 - f_2} = 0 \qquad (29a)$$

$$\frac{P_{0,1}}{f_2} + \frac{P_{-1,1}}{f_2 - f_1} = 0. \qquad (29b)$$

In terms of the circuit parameters these expressions are:

$$\frac{P_s}{f_s} - \frac{P_i}{f_i} = 0 \qquad (30a)$$

$$\frac{P_p}{f_p} + \frac{P_i}{f_i} = 0. \qquad (30b)$$

If it is assumed that P_s is positive and P_i is negative, the first equation can never be fulfilled. Looking at the second equation, it can be seen that for P_i negative, P_p must be positive, which is as would be expected. Therefore, the first equation reveals that for P_i negative, P_s must also be negative. Therefore, there is a net power flow out of the lsbuc at the signal frequency. The pump power divides between P_i and P_s in the ratio of $\frac{f_i}{f_p} P_p$ and $\frac{f_s}{f_p} P_p$ respectively. This regeneration of the signal means that the lsbuc has greater gain than the corresponding usbuc.

Parametric Amplifiers

The Manley-Rowe relations for the lsbuc reveal that more power is reflected than transmitted at the signal frequency, indicating that a negative resistance exists at the input port. In other words the reflection coefficient at the signal port is greater than one. It is therefore possible to build a varactor amplifier as shown in Figure 21. To do so, it is necessary that the varactor be pumped at a frequency higher than the signal and power be dissipated at the difference frequency, also known as the idler frequency.

This device is usually called a "parametric amplifier" since amplification is caused by the varying capacitance parameter as explained above. There are two types of simple parametric amplifiers: nondegenerate and degenerate. The nondegenerate paramp is one in which the idler frequency is different from the signal frequency, and the degenerate paramp is one in which the idler frequency is almost equal to it.

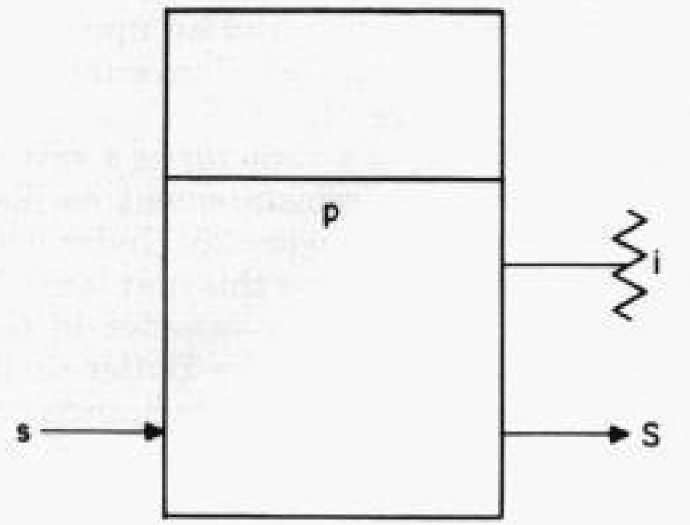


Figure 21. Representation of a Parametric Amplifier

To obtain optimum noise performance in the simple properly tuned nondegenerate paramp, the pump frequency, f_p should be chosen so that:

$$f_p = \sqrt{m_1^2 f_c^2 + f_s^2} \qquad (31)$$

This optimum noise performance is obtained when the idler is terminated in only the series resistance R_s. The usual technique for loading the idler is to place the varactor in a circuit which resonates at the idler frequency with the effective elastance of the diode. The gain of the parametric amplifier is expressed as:

$$g_c = \frac{4 R_o^2}{\left[R_o + R_s - \frac{(m_1 \omega_c)^2 R_s^2}{\omega_s \omega_i (R_s + R_i)} \right]^2} \cdot \qquad (32)$$

Because this expression is not found in any of the cited references, the circuit and the complete derivation is presented in Appendix II. Also shown in the Appendix II, the negative resistance is expressed as:

$$R = \frac{(m_1 \omega_c)^2 R_s^2}{\omega_s \omega_i (R_s + R_i)} \quad . \qquad (33)$$

Thus, it can be seen that to have gain, the negative resistance must be greater in magnitude than the loss within the varactor, represented by R_s. To avoid oscillation the denominator in the gain expression must not equal zero. Therefore

$$R_s < R < R_s + R_o \qquad (34)$$

The expression for the input resistance is

$$R_{in} = R_s \left[1 - \frac{(m_1 \omega_c)^2}{\omega_s \omega_i} \frac{R_s}{R_s + R_i} \right]. \qquad (35)$$

The negative resistance portion can be maximized by making $R_i = 0$. This is generally accomplished by resonating the varactor at the idler frequency. The input resistance for this case becomes

$$R_{in} = R_s \left[1 - \frac{(m_1 \omega_c)^2}{\omega_s \omega_i} \right]. \qquad (36)$$

It can be seen that to have gain

$$\omega_s \omega_i < (m_1 \omega_c)^2. \qquad (37)$$

For positive ω_i, this can be rewritten as:

$$\omega_s < \omega_p < \frac{(m_1 \omega_c)^2}{\omega_s} + \omega_s. \qquad (38)$$

Dividing Circuits

As indicated above, if the negative resistance of the paramp is made large enough, it will begin to oscillate. Then there would be only one input at the pump frequency and two outputs.

If a degenerate paramp is allowed to oscillate, a divide - by - two circuit is formed where only one output exists, as shown in Figure 22. It can be shown that this configuration has a phase-locking mechanism in which the phase at the output is at every instant related to the phase of the input. This circuit will therefore be stable in frequency output, a property not possessed by any other varactor dividing circuit.

An oscillating nondegenerate paramp can be used as a divide by-n-circuit, where $n > 2$. Consider the divide-by-three circuit shown in Figure 23. This circuit can also be used as a multiply by two-thirds circuit. The only requirement is that the two output frequencies add up to the pump frequency. One disadvantage of this circuit is that it has no mechanism which keeps the two outputs from drifting in frequency. One output may increase a certain amount while the other may decrease the same amount, and vice versa. The two outputs together will still add up to the pump frequency.

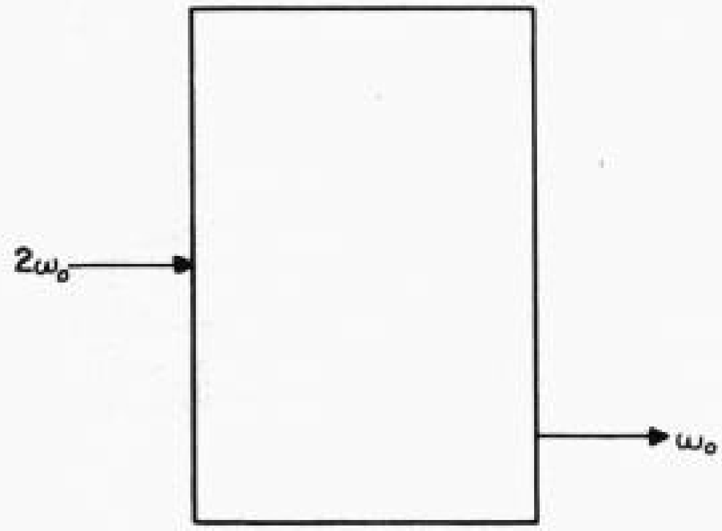


Figure 22. Representation of a Divide-by-Two Circuit

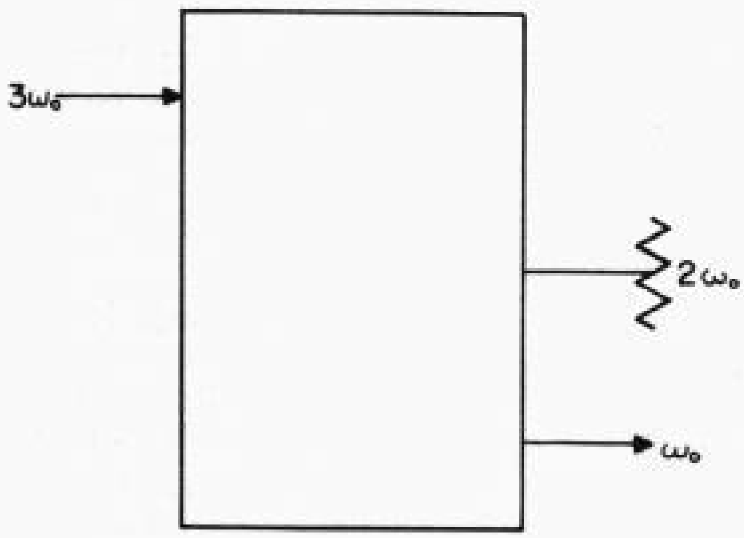


Figure 23. Representation of a Divide-by-Three Circuit

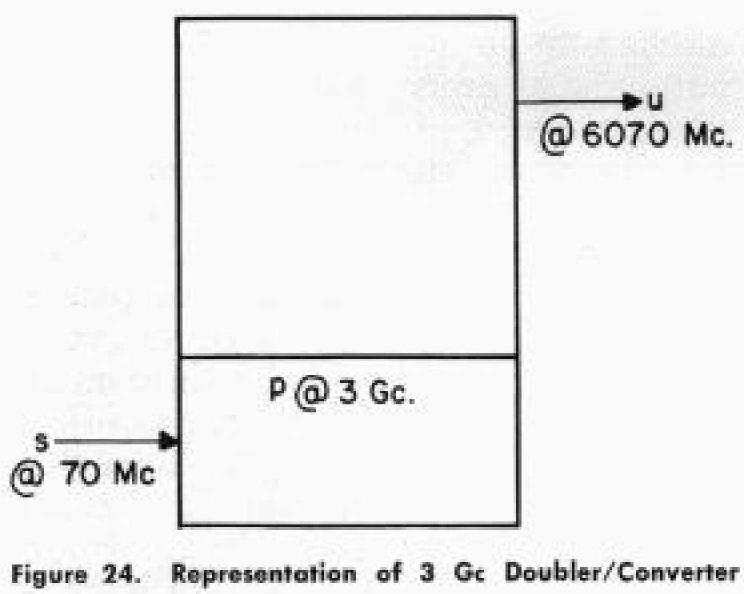


Figure 24. Representation of 3 Gc Doubler/Converter

Applications of Varactors

At present, voltage-variable capacitors are utilized at Western Union in various roles in the WLD-6 cross-country radio beam equipment. Receiver and transmitter local oscillators use these devices in frequency multiplier circuits.

Another application where one varactor serves as a doubler and an upconverter is in a 3Gc Amplifier/Converter unit as shown in Figure 24.

The properties of a varactor as a switch are employed in a reinstatement oscillator circuit shown in Figure 25. Under normal operating conditions this unit is made dormant by grounding a varactor in the output tuned circuit of the Butler oscillator. The resulting high capacitance detunes the tank. When the 70 Mc input to the Converter-Driver fails, a large bias is applied to the varactor producing a capacitance which tunes the output circuit to 70 Mc and initiates oscillation. The advantage of this scheme over B+ switching is the higher reliability of vacuum tubes when plates continuously draw current.

Future microwave systems will see increased use of the varactor. Varactor multipliers packaged with transistorized oscillators are achieving the capability of replacing conventional sources of microwave power in communication relay systems. The advantages of such packages are solid-state reliability, elimination of cooling problems, and simplified power supply requirements.

While upconverters provide a simple and reliable method of applying information to a microwave frequency, their most attractive feature is that this frequency shifting is accomplished with an extremely low noise figure as compared with conventional modulation techniques.

Low-noise parametric receivers are being planned for routes where long distances over isolated terrain have to be traversed via radio. In a low interference environment paramps could extend hop lengths considerably.

The outstanding feature of the parametric amplifier is its ability to amplify a low level microwave signal with a very low noise figure. It is more attractive than devices which can give better noise performance, such as the maser, because it is relatively inexpensive and simple to operate, and it does not usually require elaborate cooling mechanisms.

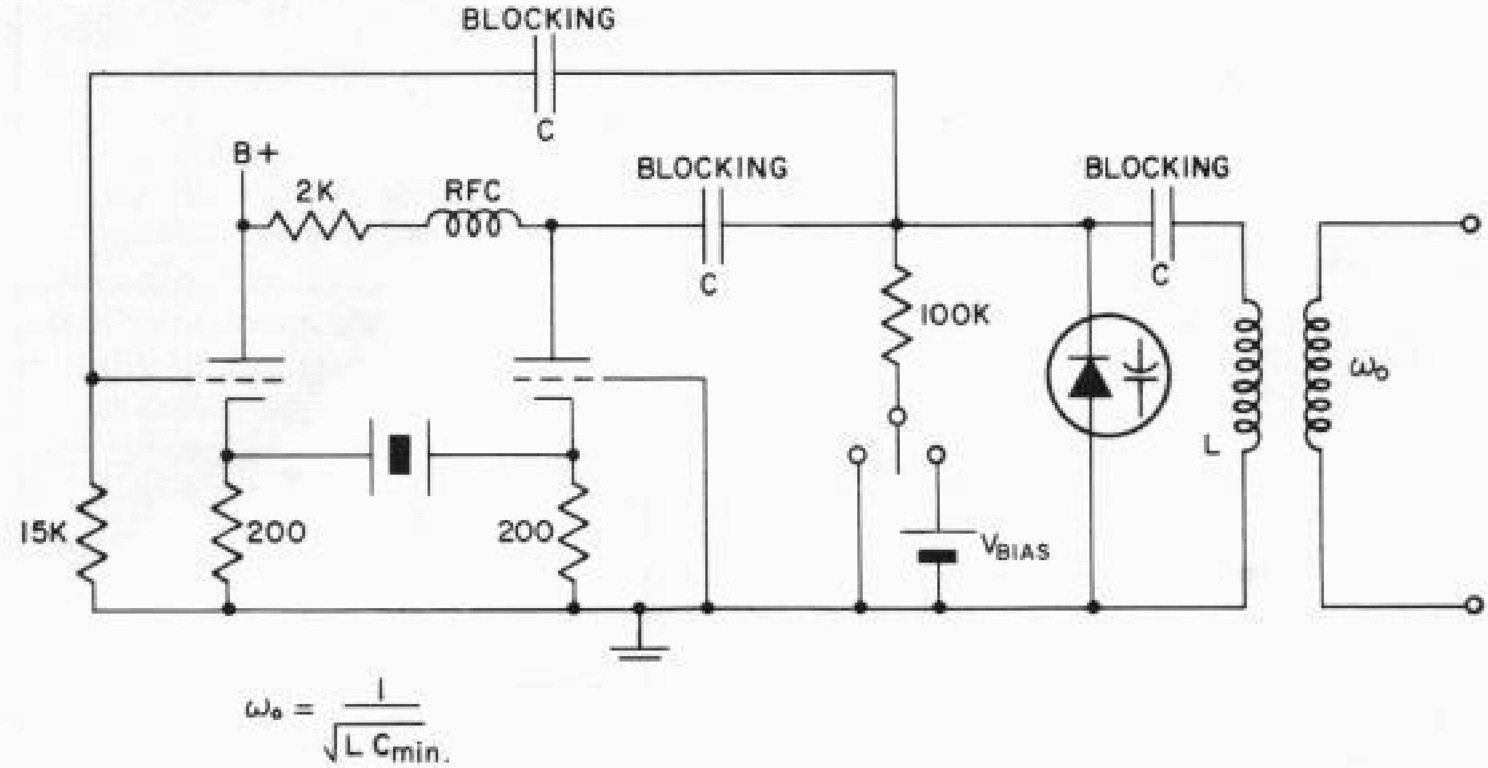


Figure 25. Varactor-Tuned Reinstatement Oscillator

Many varactor applications are not discussed in this article because of their limited use in telegraphy at present. Among these applications are limiters, phase shifters, attenuators, duplexers, electrically tunable filters, and frequency modulators. The total number of possible varactor applications are limited only by the creativity of design engineers.

* * *

References

5. Diamond, B. L., "Idler Circuits in Varactor Ferquency Multipliers" Thesis EE Dept. M.I.T. February 1961.

6. Luettegenau, G.G., "Analysis and Design of the Varicap Frequency Doubler" Application Note 25A Pacific Semiconductors Inc., Lawndale, California.

7. Damon, R. W., "Solid State Control of Microwaves" Application Note Microwave Associates, Burlington, Mass.

• • • •

Errata—In Part 1, Equation 2b should read:

$$\sum_{m=-\infty}^{\infty} \sum_{n=o}^{\infty} \frac{n\,P_{m,n}}{mf_1 + nf_2} = o \qquad (2b)$$

R. L. Ernst

R. L. Ernst and J. K. Fitzpatrick complete their article in this issue. Part I including their biographical sketches appeared in the January 1963 issue of the Western Union Technical Review.

They have been engaged in the development of varactor diode circuitry and have been concentrating on the applications of this circuitry to microwave systems. Previously Messrs. Ernst and Fitzpatrick were assigned to the Radio Systems Division. They are now responsible to the Radio Beam Expansion Project Team.

J. K. Fitzpatrick

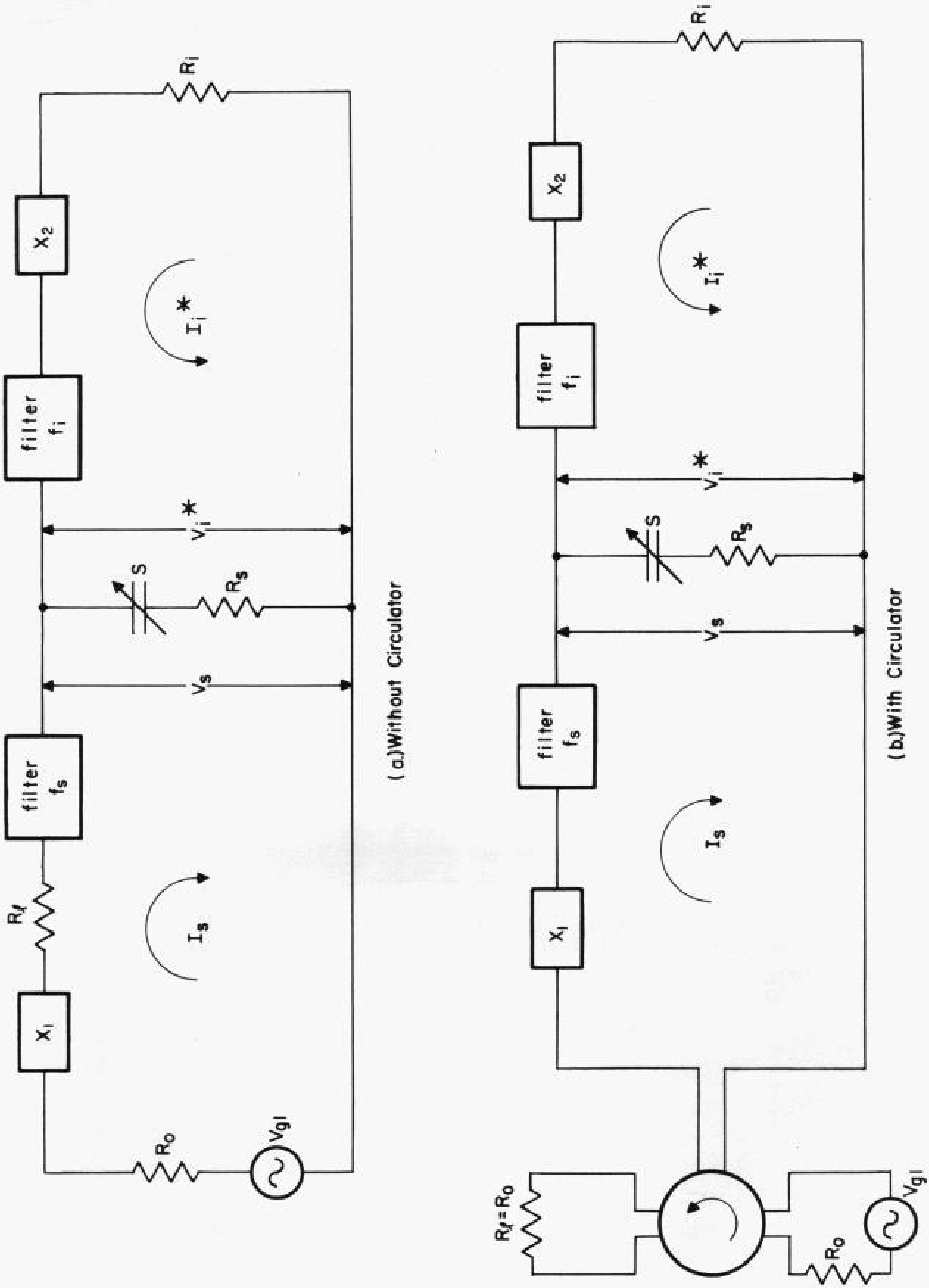


Figure II-1. Parametric Amplifier Circuit Models

APPENDIX II

Gain in a Parametric Amplifier

Consider the circuit shown in Figure II-1a. The loop on the left is at the signal frequency; the loop on the right at the idler frequency; the elastance varies periodically at the pump frequency. The matrix equation for this circuit is:

$$\begin{bmatrix} V_{g1} \\ 0 \end{bmatrix} = \begin{bmatrix} Z_{11} + Z_{T1} & Z_{12} \\ Z_{21} & Z_{22} + Z_{T2}^* \end{bmatrix} \begin{bmatrix} I_s \\ I_i^* \end{bmatrix}, \quad \text{(I-1)}$$

where $Z_{T1} = R_o + j X_1 + R_l$,

and $Z_{T2} = R_i + j X_2$.

The terms Z_{11}, Z_{12}, Z_{21}, Z_{22} are the corresponding terms of the matrix for the varactor:

$$\begin{bmatrix} V_s \\ \\ V_i^* \end{bmatrix} = \begin{bmatrix} R_s + \dfrac{S_o}{j\omega_o} & \dfrac{S_1}{j\omega_i} \\ \\ -\dfrac{S_1^*}{j\omega_s} & R_s - \dfrac{S_o}{j\omega_i} \end{bmatrix} \begin{bmatrix} I_s \\ \\ I_i^* \end{bmatrix}. \quad \text{(II-2)}$$

Manipulating the equations of the first matrix yields:

$$I_i^* = -\frac{Z_{21}}{Z_{22} + Z_{T2}^*} I_s, \quad \text{(II-3)}$$

$$V_{g1} = (Z_{11} + Z_{T1})\, I_s - \frac{Z_{12} Z_{21}}{Z_{22} + Z_{T2}^*} I_s, \quad \text{(II-4)}$$

$$I_s = \frac{V_{g1}\,(Z_{22} + Z_{T2}^*)}{(Z_{11} + Z_{T1})\,(Z_{22} + Z_{T2}^*) - (Z_{12} Z_{21})}. \quad \text{(II-5)}$$

The transducer gain at the signal frequency is given by the ratio of the power dissipated in R_l to the power available from the generator. Therefore:

$$g_t = \frac{|I_s|^2 R_l}{|V_{g1}|^2 / 4R_o} \quad \text{(II-6)}$$

$$g_t = \frac{4 R_o R_l |(Z_{22} + Z_{T2}^*)|^2}{|(Z_{11} + Z_{T1})\,(Z_{22} + Z_{T2}^*) - Z_{12} Z_{21}|^2} \quad \text{(II-7)}$$

When the two loops are tuned:

$$Z_{11} + Z_{T1} = R_o + R_l + R_s$$

$$Z_{22} + Z_{T2}^* = R_i + R_s\,.$$

The gain then becomes:

$$g_t = \frac{4 R_o R_l (R_s + R_i)^2}{\left[(R_s + R_o + R_l)\,(R_s + R_i) - \dfrac{|S_1|^2}{\omega_s \omega_i}\right]^2}, \quad \text{(II-8)}$$

$$g_t = \frac{4 R_o R_l}{\left[(R_s + R_o + R_l) - \dfrac{m_1^2 (S_{max} - S_{min})^2}{\omega_s \omega_i (R_s + R_i)}\right]^2} \quad \text{(II-9)}$$

Realizing that:

$$m_1 (S_{max} - S_{min}) = \frac{m_1 (S_{max} - S_{min})\, R_s}{R_s}$$

$$= m_1 \omega_c R_s \quad \text{(II-10)}$$

the gain expression becomes:

$$g_t = \frac{4 R_o R_l}{\left[(R_s + R_o + R_l) - \dfrac{(m_1 \omega_c)^2 R_s^2}{\omega_s \omega_i (R_s + R_i)}\right]^2}. \quad \text{(II-11)}$$

It can be seen that the parametric amplifier has a negative resistance component at the signal frequency given by:

$$R = \frac{(m_1 \omega_c)^2 R_s^2}{\omega_s \omega_i (R_s + R_i)}. \quad \text{(II-12)}$$

This circuit has the disadvantages that it may easily break into oscillations and that since it would be impossible to match the generator, the gain given above could never be achieved. To solve these problems at microwave frequencies, a circulator is employed as shown in Figure II-1b. The gain of this circuit is the square of the reflection coefficient or:

$$g_c = \left|\frac{Z_l - Z_o}{Z_l + Z_o}\right|^2. \quad \text{(II-13)}$$

When tuned this becomes:

$$g_c = \frac{(R_s - R - R_o)^2}{(R_s - R + R_o)^2}. \quad \text{(II-14)}$$

At high gain $R \approx R_s + R_o$ and

$$g_c = \frac{4 R_o^2}{(R_o + R_s - R)^2} \quad \text{(II-15)}$$

or

$$g_c = \frac{4 R_o^2}{\left[R_o + R_s - \dfrac{(m_1 \omega_c)^2 R_s^2}{\omega_s \omega_i (R_s + R_i)}\right]^2}. \quad \text{(II-16)}$$

G. B. Woodman, General Supervisor of Operations, Plant Department

New Ideas in Microwave System Maintenance

A description of our Maintenance practices on the original microwave triangle was published in the October 1951 issue of the TECHNICAL REVIEW. That system was expanded to include Pittsburgh, Cincinnati and Chicago. We are now completing building a transcontinental microwave transmission system. Since the new system is considerably more sophisticated than the original triangle our maintenance practices have required a review and up-dating to provide and maintain the necessary high quality service in Western Union facilities.

Before describing the concepts of a proper maintenance organization for our program, it seems desirable to briefly describe the main features of the new microwave system.

Expansion Program

The system now under construction will include Avoidance Routing. From the East Coast to Los Angeles, the main line will be located some distance from the major cities, served by spurs. Other major routes will interconnect with the main line via a number of major and minor junction stations. At present the main routes of the system are being constructed to carry 600 voice bands; the spurs will carry 240 voice bands.

Routes handling 600 voice bands will operate in the 6kMc range, while routes handling 240 voice bands will operate in the 4kMc range. Equipment for the former has been designed, built and is being installed by RCA. The 4kMc equipment, designed by Western Union and being built and installed by Raytheon, is a modified version of the Pittsburgh to Chicago leg of the existing system.

General Electric has the contract for supplying the Multiplexing (Carrier) Equipment at all terminal cities and at junction stations. To obtain the necessary interconnections at junction stations, the radio beam will be demodulated into supergroups or groups. No voice bands will be available at those locations. However, at terminals, voice bands will be available.

This system will use combiners at all terminal and junction stations. In addition, four other combining stations are being installed where the distance between junction stations is excessive.

Automatic I.F. switching at each relay station will add greater reliability to the system. While this particular feature is not accomplished without a delay of a few milliseconds, in contrast to the instantaneous action of the combiners, it eliminates the failure of the beam due to simultaneous break-downs in either path, except at the same station.

Fault-Locating System

A very sophisticated fault-reporting system will permit constant surveillance of equipment conditions at all stations. This system will center on six major supervisory control stations which are the nerve centers of the entire complex. Basically each station automatically and electronically interrogates itself six times a minute. Up to sixty items may be questioned at some stations and if any are found at fault a signal will be transmitted to a fault-locating console at the manned junction.

This signal is unique to one station and each station is assigned its own signaling frequency. This arrangement permits identification of each relay station in the selected complement of stations associated with that manned junction. Assuming that a trouble is signalled from a certain station, a lighted panel marked "FAULT" appears on the fault panel at the control junction opposite the name of that station. At the same time a chime sounds to call the attendant's attention to the irregularity. By suitable manipulation of controls the maintainer can determine which of the troubles have actually occurred at the station indicated.

It is then the attendant's duty to call the maintainer for that station and advise him of the trouble and its nature.

The above briefly describes the handling of the RCA, WLD-6 troubles. Separate boards for the GE multiplex and MLD-4B equipment are located adjacent to the RCA board at the manned junction station. While the detailed operation of the latter two varies from that described, the main function of providing fault-locating ability is achieved. In certain instances where the trouble is due to the failure of multiplex equipment, it can be corrected by swapping in operating spares by remote control. These major junction stations will be manned 24 hours a day and seven days a week by radio maintainers.

Another feature of the fault-locating system is the "Hit Detectors." Early in our experience with the working microwave systems, we found that momentary breaks, or hits, appeared as a wrong character or "garble" in the message.

It was observed that the frequency of these hits, as reported from the operating people, tended to increase over a period of time until some tube failure or other break down occurred. This observation led to devising a means for recording these hits without having to depend on reports from Traffic Department. The result was a biased relay device, operating over a spare telegraph channel, which would activate an alarm whenever the banked tongue of the relay broke away from the contact. Each section of our present system is equipped with such a hit detector, and they will also be provided on the new transcontinental network.

Experience has shown that observation of increased hit activity has made possible preventive maintenance which has forestalled beam failures in many cases.

Power Plants

Emergency power plants, as shown in Figure 1, are being installed at all relay stations to carry the power loads during periods of commercial power failure. At five stations, where commercial power will not be available, large diesel-driven generators will supply prime power. It is imperative that failure of the normal power is not permitted to cause the slightest interruption to the radio signal. This can be accomplished by driving an ac motor from the commercial source and coupling it directly to an ac generator which carries the radio load. This same shaft carries an extremely heavy flywheel which stores energy during normal operation. Upon failure of the outside power, a clutch engages the flywheel with the drive shaft of a diesel engine; thus the energy in the flywheel cranks the engine. This energy is also sufficient to keep the generator output practically constant until the engine takes over.

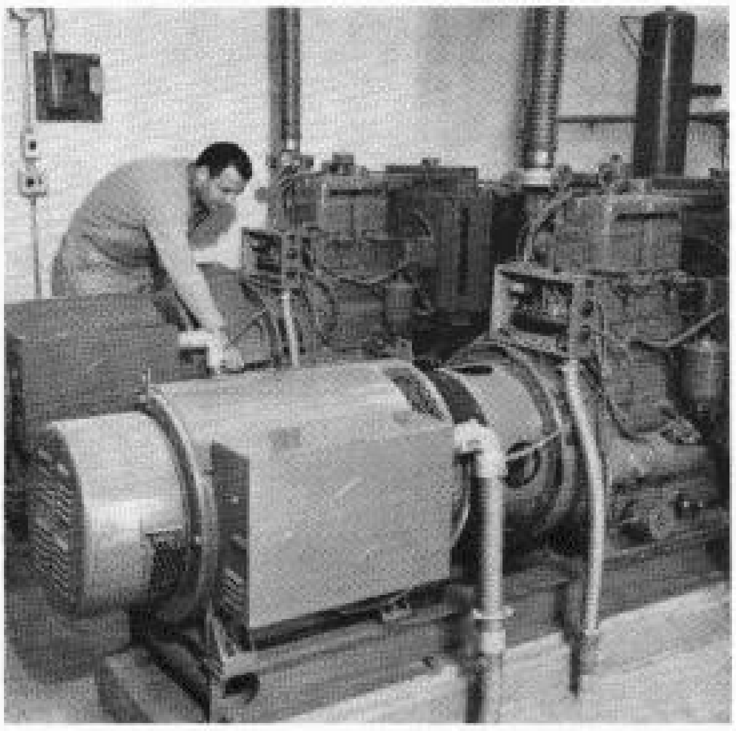


Figure 1. Emergency Power Plant.

A lapse of about ten seconds or six cycles occurs between the time outside power fails and the time the engine takes over. Each of the two beams at any station has its own motor generator-engine for individual protection. A cabinet of suitable relays and transfer switches provides for automatic or manual load switching from one machine to the other for maintenance or safety purposes.

Fuel is stored in large tanks, these vary in size depending upon the power load at the station, the distance from a supply source and the accessibility of the station under all weather conditions.

Prime power is supplied at five stations. At each of these, two machines, each capable of carrying the entire station load, are being installed. They will be operated on alternate weeks, so that one machine is always available for overhaul and maintenance. These machines feed into emergency plants exactly like those provided where commercial power is supplied. This provides continuous power until a maintainer can reach the station, should the main source fail.

Maintenance Organization

The key man in the new maintenance concept is the trained maintainer, he will be assigned three or four stations which he will visit periodically. He will be backed up by a traveling team of two field engineers who will be assigned 25 to 35 stations. It will be the responsibility of this team to visit each station in turn and completely overhaul all equipment. It is expected that this will take about a week at a station as an average; so that each station will be visited every six to eight months. Complete test equipment will be available to the traveling team so that all checks and tests can be made. A partial listing of 60 or more types of test equipment, which will be available to the maintainers and mobile teams, is as follows:

- Amplifiers
- Attenuators
- Counters
- Demodulators and Modulators
- Generators (Signal & Sweep)
- RF Loads
- Meters
  - RF power
  - Selective Volts
  - AC
  - DC
  - Wavemeters
- Oscillators
- Oscilloscopes
- White Noise Test Set
- Group Delay and Linearity Test Set
- Tube Testers

To provide top level supervision of the entire Radio Beam system, three supervisors are assigned to three of the six major junction stations, namely:

- McGraw, near Syracuse, New York
- Berwick, near Kansas City, Mo.
- Mt. Aukum, near Sacramento, California

These men will keep in close touch with all activity within their jurisdiction and exercise supervisory control. Each man reports to the Area Maintenance and Operations Supervisor of the Plant Area in which his headquarters station is located.

Vehicles

The maintainer's vehicle, as shown in Figure 2, will be a Chevrolet "Carryall," either two or four-wheel drive with positraction depending upon the assignment. Twelve snow vehicles are planned for use in snow country. Two types of tracked units have been tried out in Utah and Nevada the past two winters. As might be expected, there are advantages and disadvantages in each.

Figure 2. Maintainer's "Carryall".

Training of Maintenance Personnel

To prepare our people to maintain and operate this complex, a training school was established at Chattanooga, Tenn. A complete installation of all types of equipment used was set up.

Each man was given ten weeks of intensive instruction combining lectures with laboratory work. In addition, he was given 4 weeks assistance in preparing for his 2nd class Radio Telephone license, which is an FCC requirement. Wire and Repeater Technicians and Supervisors from the terminal cities were given two weeks training on the multiplexing equipment only. About 250 men have been trained at Chattanooga for this program. While this was a rather accelerated course, it was intended to provide a sound basis for the on-the-job training which followed.

Intra-System Communication

Communication between relay stations, junction stations, terminal cities and the maintainers will be available in five ways:

- A teleprinter circuit connecting certain stations with the control junction stations.
- A local Order-Wire connecting all stations.
- A supervisory Order-Wire connecting junction stations and terminal cities.
- An express Order-Wire connecting Major Points.
- A VHF Mobile radio connecting the manned junction stations with the Maintainers' Vehicles. This permits contact with the Maintainer when he is on his way to or from a station.

The Mobile Teams' vehicles will also be equipped with two-way radio. This part of the program involves 108 base stations at selected relay stations and 101 mobile two-way radios installed in cars. The 108 base transmitters are activated either by selection coding from manned junction stations or from mobile units.

Scope of the Maintenance Program

This brief article has necessarily avoided many of the technical details which require much description. It is intended only to indicate the broad aspects of the maintenance program planned for the Western Union Microwave Transcontinental Network. A summary of a few statistics may serve to point up the magnitude and scope of the project.

- 246 repeater stations and 27 terminal cities are involved.
- 84 carryalls, 12 utility sedans and 12 snow vehicles will be used by maintenance personnel.
- 520 power plants will be installed.
- 250 men have been trained at the school.

The program, including training of maintainers, the selection of test equipment, and the providing of tools and spare parts necessary to maintain the microwave system represents a considerable planning effort on the part of all concerned.

MR. G. B. WOODMAN is currently responsible for the maintenance and operation of the Microwave Systems, and also for carrier, PCH, weather-fax, broadband switching fax/phone/data circuits, and voice circuits.

After joining Western Union in 1925, he was associated with the Eastern Plant Division until 1947, when he was promoted to the General Plant Office.

Mr. Woodman received his B.S. degree in Electrical Engineering from the University of New Hampshire.

The previous articles entitled, "Maintenance of a Radio Relay System," and "Improved Vacuum Tube Reliability through Maintenance" appeared in the October 1951 and October 1957 issues of the Western Union TECHNICAL REVIEW.

E. C. Ottenberg, Senior Project Engineer, Radio Beam Expansion Project

Delay, Linearity and White Noise Testing of F. M. Radio Relay Systems

WESTERN UNION is now installing a high capacity coast-to-coast radio relay system which replaces and/or extends the low capacity, short-haul systems with which we have had experience since 1945. Test equipment requirements for maintenance of the new system are far in excess of that required for the short-haul systems, in cost, quantity and sophistication. Maintainers now require some $29,000 worth of equipment to perform their jobs in a satisfactory manner.

This equipment is used to make measurements necessary for the proper performance of high-capacity, long-haul microwave radio relay systems. Some measurements — Delay, Linearity, and White Noise Loading—are used to determine the performance capability of the system.

For the purposes of this discussion the following definitions apply:

- A frequency modulation (F.M.) generator does not include limiters and the baseband amplifier which may be part of an F.M. modulator.
- An F.M. detector does not include limiters and the baseband amplifier which may be part of an F.M. demodulator.
- The I.F. or intermediate frequency is a frequency in the tens of Mc range which is common to all radio repeaters and terminals.

Delay and Linearity

In the case of a F.M. system which is modulated with a 1-Mc tone at a peak deviation of 2-Mc, the theoretical sideband power distribution is shown in Figure 1a; all sidebands with less than 1% of the power are not considered. In order that the 1-Mc tone be reproduced faithfully, the radio relay system must generate in a modulator, amplify in repeaters, and detect in a demodulator this theoretical sideband distribution. This assumes perfect propagation. In a radio relay system 186 miles long, it would take a signal, with or without a particular sideband distribution, one millisecond to travel the length of this system, provided that there were no repeaters. When the signal arrives at the F.M. detector (assuming a perfect F.M. generator and detector), all sidebands have the same relative time and amplitude relationship as they did when they were produced. The 1-Mc tone would then be reproduced perfectly. If each part of the system (modulator, demodulator, radio repeater) had an amplitude and delay response as shown in Figure 2, the relative time and amplitude relationship of the sidebands would change and the F.M. detector would not reproduce the 1-Mc tone faithfully.

The Linearity Test

The linearity test (sometimes called derivative test) is primarily a test of the F.M. modulator and demodulator. It is a function of the modulator linearity and the detector linearity or its derivative response. This assumes a limiter located in the line before the detector which would then remove all amplitude fluctuations from the I.F. signal over the designated frequency range. In practice, there is sufficient limiting in the demodulator to make its amplitude response flat enough so that any amplitude fluctuations in other parts of the system are not considered. Then the derivative response of the demodulator is considered to be that of the F.M. detector.

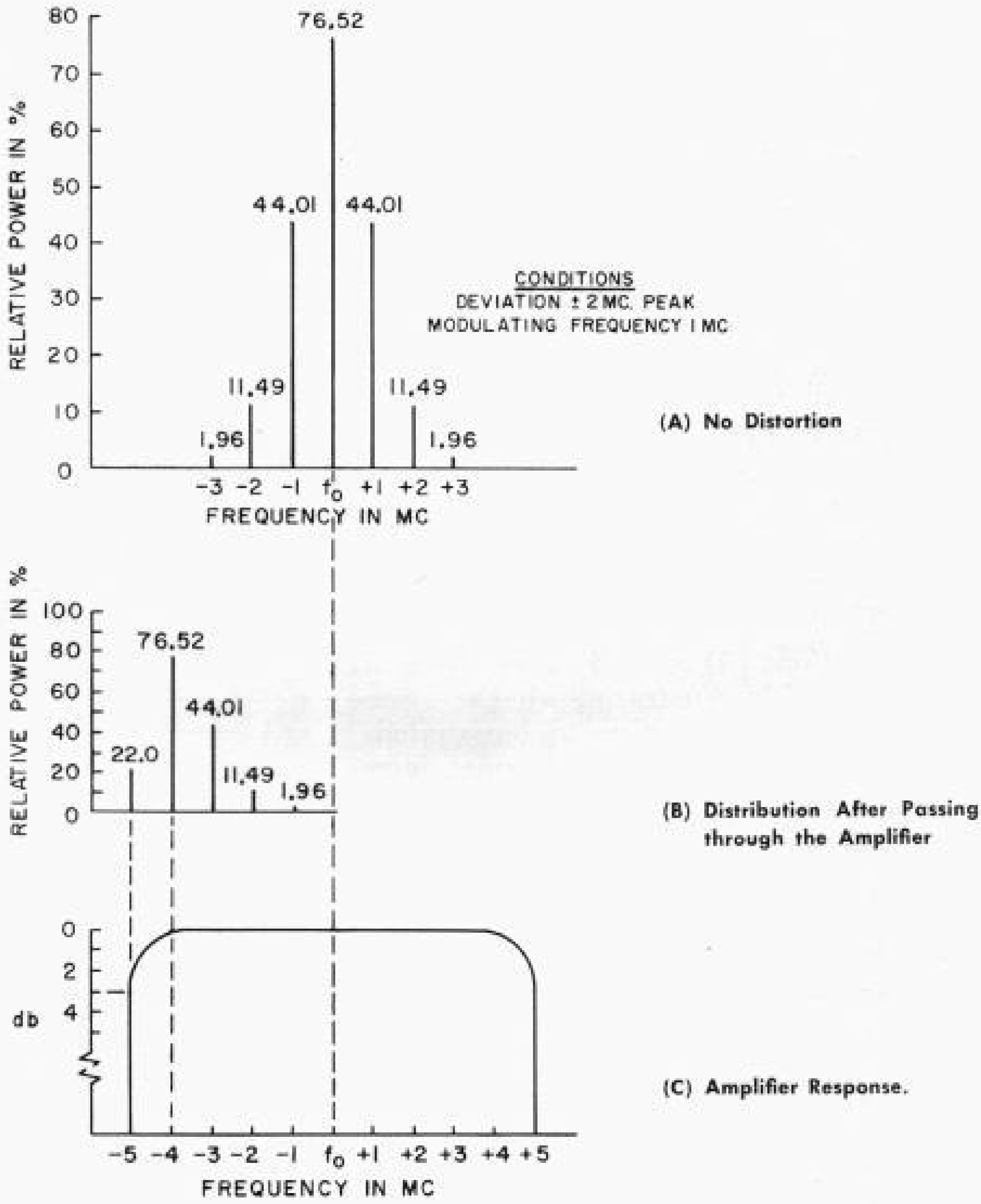


Figure 1. Sideband Power Distribution

Consider a perfect F.M. detector driven by an F.M. modulator; the two units being required to reproduce a baseband tone of 1-Mc. This tone must be reproduced at the detector output without distortion. It requires that the deviation of the modulator be the same for all I.F. frequencies in the designated range. As a simple example assume that the F.M. modulator had in its output an isolation amplifier which has the amplitude response shown in Figure 1c. As the I.F. frequency is changed from f_0 to (f_0-4) Mc, the two lower sidebands are not passed and the first lower sideband is reduced 50 percent. The sideband distribution will now be that of Figure 1b and the detected 1-Mc tone will not only be distorted, but will produce a different output from that if the I.F. frequency were centered on the amplifier amplitude response. Other parts of the modulator may exhibit a characteristic which would produce sideband distortion of the modulating tones. The linearity test is a measure of the distortion in the modulator.

Now let us consider a perfect modulator driving an F.M. detector. Figure 3a shows an F.M. detector characteristic. The equation of the section A-A_1 may be written:

$$X = By \text{ (where B is a constant)} \qquad (1)$$

The equation for section B-B_1 may be written:

$$X' = C(y')^n \text{ (where C is a constant and } n > 1.) \qquad (2)$$

If we differentiate equations (1) and (2), we obtain:

$$\frac{dx}{dy} = B \qquad (3)$$

$$\frac{dx'}{dy'} = Cny'^{\,(n-1)} \qquad (4)$$

Equations (3) and (4) may be better understood by illustrating what was done when the equations were "differentiated." Starting from f_o, as small equal increments are taken along the y axis, the

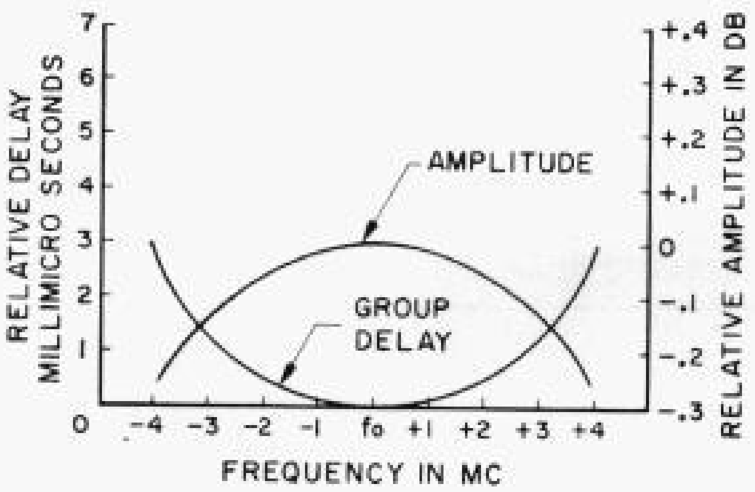


Figure 2. Group Delay and Amplitude Response vs. Frequency

increments on the x axis are recorded. If, for small equal increments in y, the x increments are equal, the equation is linear and its derivative is a constant. If, for small equal increments in y, the x increments are not equal, the equation is non-linear and the derivative is some function of y. To "differentiate" an equation then is to express its linearity mathematically.

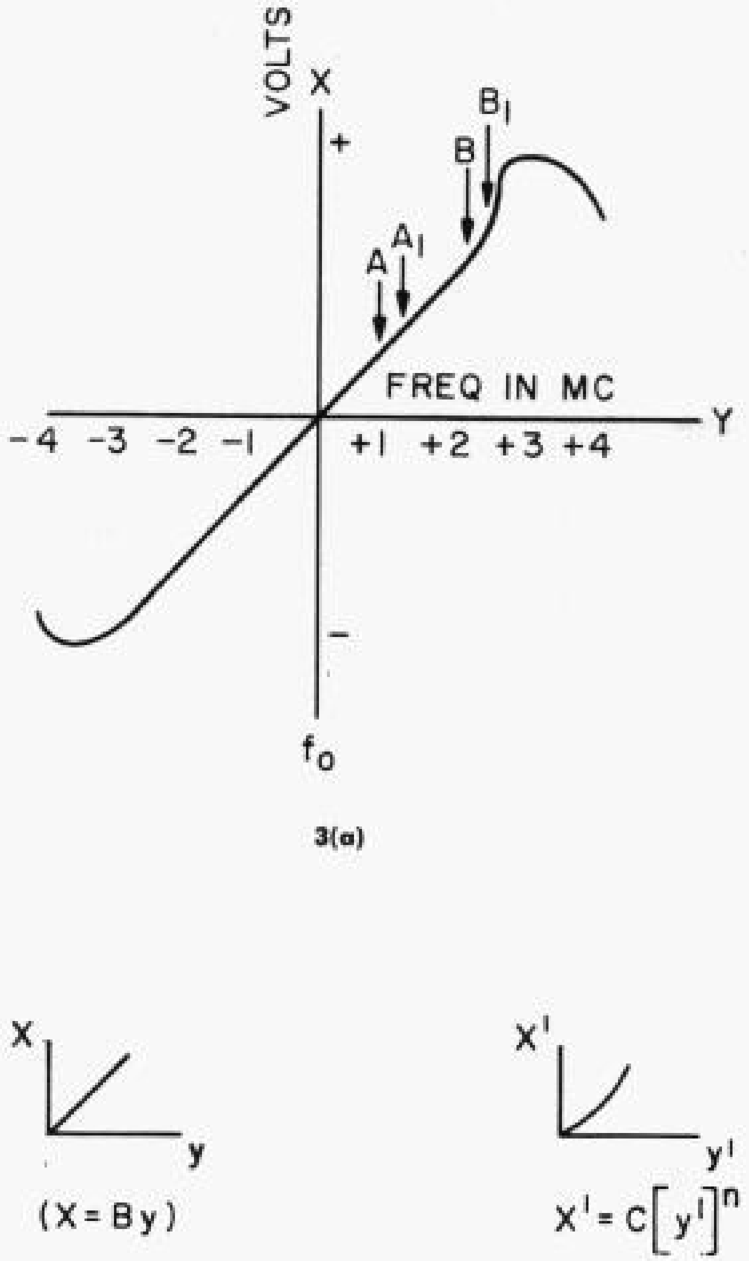


Figure 3. F.M. Detector Characteristic

If instead of performing the actual measurements we look at the differentiated Equations (3) and (4) and substitute for y (frequency) and solve for $\frac{dx}{dy}$ (volts output), it is easily seen that Equation (3) gives a constant $\frac{dx}{dy}$ for any y; while in Equation (4) $\frac{dx'}{dy'}$ varies with y'. It is apparent then, that if a perfect modulator drives a detector described by (3), the relative levels of the demodulated tones will be the same, but will be different if the detector characteristic is described by (4). As in the case of the modulator, this is a test of the ability of the detector to reproduce signals without distortion. Linearity may then be designated a system test since the modulator and the demodulator are located hundreds of miles apart—and are not being bench tested on a "bust-back" basis.

Group Delay Test

As mentioned previously, the relative time delay of the sidebands produced by any one tone must be unchanged to reproduce perfectly the tone at the output of the F.M. detector. The absolute time delay is not important for this test. However, it is important that any part of the system (modulator, demodulator, or repeater) produce a relative delay vs. I.F. frequency which is a constant. Therefore, it is necessary to measure the delay not only of the modulator and demodulator, but of all circuits which carry the I.F. signal. The delay vs. frequency response is then another parameter which measures the distortion of a system. Again the Group Delay Test is a system test for the same reason given for the linearity test.

Delay and Linearity Test Set

In practice it must be realized that modulators, demodulators and test equipment are not perfect. Certain assumptions must be made, and procedures based on these assumptions are then prepared. A little thought will lead to the conclusion that any test set should be made in two parts since the equipments to be tested may be hundreds of miles apart. In addition, the test set should have the facility of testing the demodulator with an I.F. signal and the modulator with a baseband signal. One point should be recognized. The linearity test is a function of the modulator and the demodulator. If the linearity of a modulator or demodulator is bad, either unit can be tested separately. The delay is a function of all units between and including the modulator and demodulator. In this case a modulation section may appear to meet the delay specification, however, if one unit is changed, it may not. This may be caused by two units located hundreds of miles apart, having delay curves which are complementary but do not meet the unit specifications. Therefore, there should be some means of checking and locating such units. The delay of a system may still meet the specification but will increase as more hops are added, but the "I.F.-to-I.F." delay over one hop should be the same. This is one method of localizing a defective unit or, of checking its conformance to specification when the unit is replaced.

Figure 4 is a simplified block diagram of a Delay and Linearity Test Set. All practical test sets may or may not incorporate all of the blocks specified. If, for example, an oscilloscope is to be used for other test purposes, it would not be designed into the test set but vertical and horizontal outputs would be provided. The equipment and system tests performed by various sections of the equipment are listed in Table I.

The transmitter in Figure 4a supplies either two baseband tones or an I.F. signal frequency modulated with these tones. One tone (low frequency oscillator) is just a means of changing the R.F. frequency electronically instead of by hand. It changes the center frequency of an F.M. generator which may be located at an R.F. frequency of a few hundred to a few thousand megacycles over the desired frequency range and makes possible a

TABLE I

Transmitter		Receiver	
Baseband Section	I.F. Section	Baseband Section	I.F. and Baseband Section
Modulator delay and linearity	Demodulator delay and linearity	Demodulator delay and linearity	Modulator delay and linearity
System delay and linearity	Hop-to-Hop delay	System delay and linearity	Hop-to-Hop delay

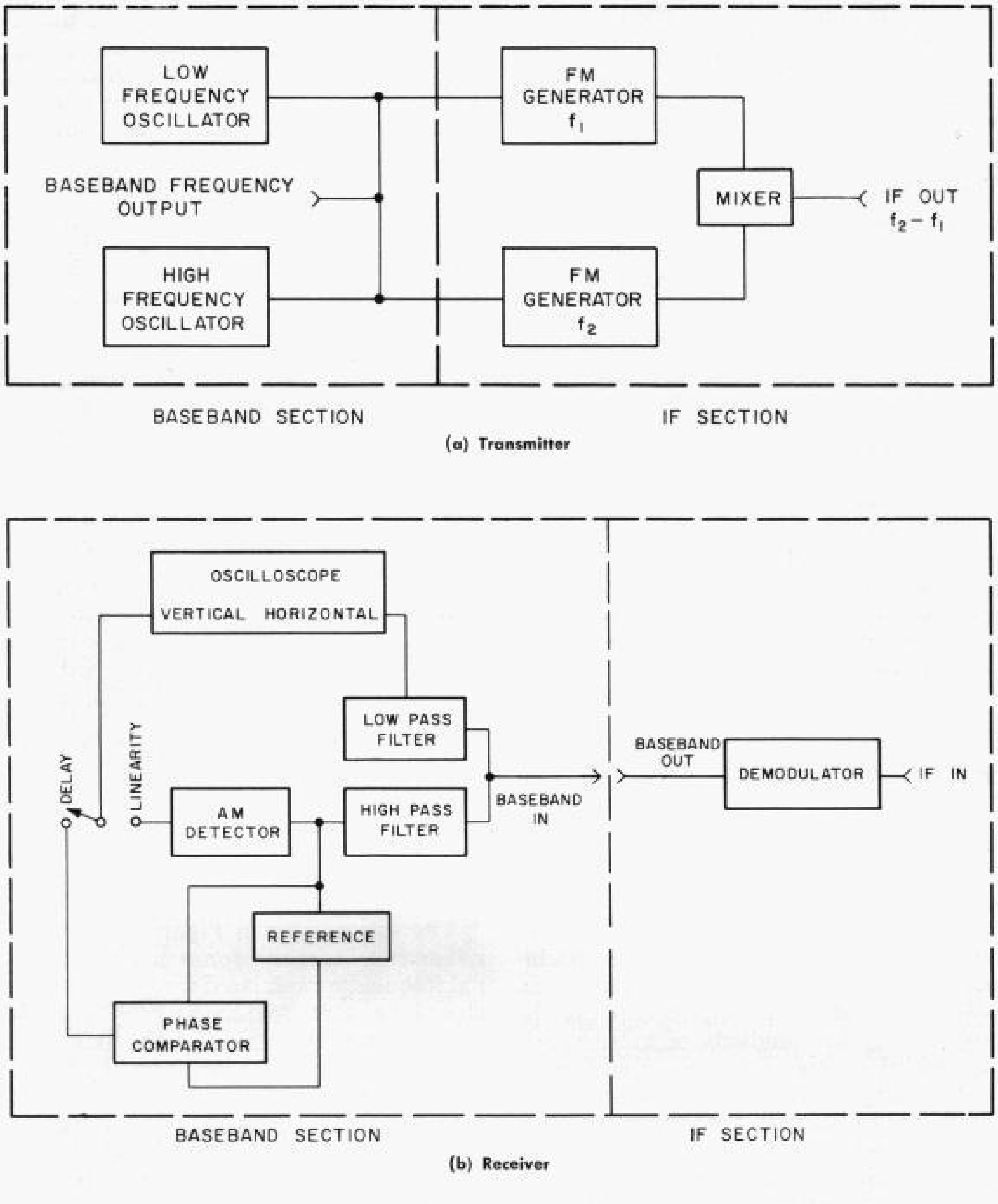


Figure 4. Delay and Linearity Test Set

visual display on the oscilloscope. This is the "sweep generator" part of the transmitter. The low frequency oscillator operates at a few hundred cycles. Sixty cycles or harmonics of sixty cycles are not used since the receiver may be hundreds of miles away. Because this frequency is used for display synchronization, there would be a problem with pickup of locally generated 60 cycles. The other tone (high frequency oscillator) is of the order of tens of thousands of cycles. It deviates an F.M. generator, whose frequency differs from the other

F.M. generator frequency by the I.F. frequency, at a very low peak deviation. The two R.F. oscillator frequencies are chosen so that, when they beat together in a mixer, the proper I.F. frequency is generated. No matter how it is introduced into the radio system it is the high frequency oscillator tone which is used to detect delay and/or linearity changes.

The receiver shown in Figure 4b is a little more complex. After the two tones are demodulated, the low frequency tone is passed through a filter and is used to synchronize the oscilloscope. The high frequency oscillator tone is passed through a filter and then is used in one of two ways. For linearity measurements it is passed through an A.M. detector which drives the vertical section of an oscilloscope.

Let us refer back to Figure 3a. As the low frequency tone sweeps the I.F. signal through points $A-A_1$ and $B-B_1$, the I.F. signal is also being frequency modulated with the high frequency tone at a low peak deviation. Instead of sweeping the I.F. signal consider it centered first between $A-A_1$ and then $B-B_1$. Since the deviation is constant, a higher level, high frequency tone will be detected when the I.F. frequency is centered between $B-B_1$ than if it were centered between $A-A_1$. When the F.M. detected tone is then passed through an A.M. detector, more DC will be produced when centered between $B-B_1$. If the I.F. frequency is then swept through these points and displayed on an oscilloscope, which is synchronized with the sweep, a visual display will result. Any deviation from a flat horizontal line is an indication of nonlinearity in the unit or system under test.

For delay measurements, the high frequency tone is passed into a reference block and a phase comparator. The reference block averages the phases of the high frequency tone as it is swept through the I.F. band to provide a tone of relative standard phase. The high frequency tone is then compared to the tone of relative standard phase as it is swept through the I.F. range in a phase detector whose output is a dc signal having an amplitude proportional to the deviations in phase away from the relative standard. Again, any variations away from a flat horizontal line is an indication of an undesirable delay characteristic. Phases are compared directly since they are related to the delay.

Any practical delay and linearity test set requires the incorporation of limiters, delay equalizers, and other circuits to insure that the test set itself, when tested by connecting the transmitter I.F. output to the receiver I.F. input, has a flat horizontal delay and linearity characteristic. In addition a method of calibrating the test set for the delay and linearity range(s) to be measured should be incorporated.

White Noise Test

Many articles have been written on the suitability of using white noise to test microwave radio systems and free use is made of the material in one of these.[1]

Let us examine two 1 kc sinusoidal tones. Each tone has a peak value 3 db above its rms value. If these two tones are phased so that at the same instant of time they reach a positive maximum, the peak of the two tones will be 6 db higher than the peak for a single tone. Now consider the same two tones with a random phase relationship. In this case the peak of the two tones will still be 6 db higher than one tone but it will not occur as often. If one of the two equal tones was a different frequency and the two tones again had a random phase relationship, the peak value of the two tones would again be 6 db higher than the peak of one tone but the percentage of the time this occurred would again be different. As equal level tones of different frequencies and random phase are added a certain statistical relationship takes place. As additional tones are added, the rms value of the multitone signal increases. For a large number of tones the instantaneous peak-to-average power, which will be exceeded only 0.003% of the time, is 13 db.

The multitone situation just described is the output of a high capacity carrier multiplex system. Statistical theory also shows that white noise, restricted to the bandwidth of a multitone carrier multiplex system, has about the same characteristics. In fact, the white noise signal is different only in that all frequencies are present in the bandwidth under consideration, while a carrier system has only discreet frequencies, the sum of which approaches the white noise signal as more channels are added. The white noise load is then a more severe load on a system than if the system was loaded with the carrier equipment. It should also be pointed out that a high capacity carrier system fully loaded with voice traffic may also be approximated by a white noise signal. The level at which the multitone carrier or white noise is placed on the system is chosen to provide the best compromise between performance and costs.

This level is set in relation to a Test Tone (T.T.) level. Both the test tone and multi-channel load levels have been defined by the International Radio Consultive Committee (C.C.I.R.) and are as follows, for 240 and 600 channel systems:

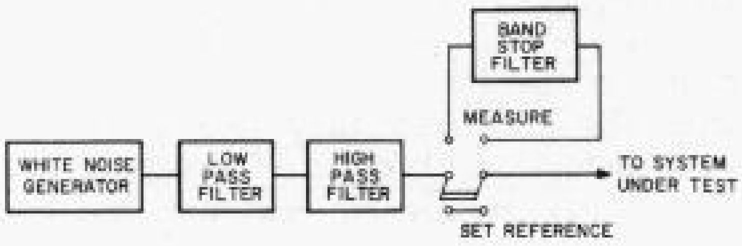
1.) The T.T. level should deviate the radio system 200 kc rms per channel.
2.) The carrier multiplex or white noise load should be at a level of $(-15 + 10 \log n)$ db above T.T., where n is the number of channels.

Other capacity systems may use the same or other levels which are given by the C.C.I.R.

White Noise Test Set

Figure 5 is a simplified block diagram of a White Noise Test Set. The noise generator produces a flat, wideband noise output. The low and high pass filters restrict this noise band to the carrier multiplex band under test. Idle bands are required in order that distortion or Noise Power Ratio (NPR) and idle noise or Baseband Intrinsic Noise Ratio (BINR) may be measured. For this purpose band stop filters which can be switched in and out of the line are used. These filters must have excellent rejection to insure that noise frequencies, in the channel to be measured, are not transmitted. In addition these filters should not reject a very wide band lest too much noise power be taken from the test signal. A bandwidth of about three to four kc is usually chosen. The noise receiver has band pass filters which are centered on the band reject filters of the channel to be measured. These filters are narrower than the band stop filters in order to insure that the white noise at the edges of the band stop filters does not cause false readings. In addition the band pass filters must have good out of band rejection without transparencies over the frequency range of the white noise test signal so that only intermodulation products are passed through the filter.

The theory of operation is straight forward. The system under test is loaded with the proper send level with the band stop filter removed. A reference level is then set up on the receiver through the bandpass filter. The band stop filter is then switched into the circuit and the difference in db in the receiver level (NPR) is read. The white noise load is then removed from the system, and again the difference in level from the reference level (BINR) is read. If a modulator and demodulator are incorporated as part of this test equipment the range of uses



(a) Transmitter

(b) Receiver

Figure 5. White Noise Test Set

may be extended. The hop-to-hop NPR and BINR may be measured if heterodyne repeaters are used by sending from the modulator into the I.F. point of one repeater and driving the demodulator with the I.F. output of the following repeater.

While the noise receiver looks simple, it is more sophisticated than a cursory examination may show. The band pass and stop filters required are very expensive. They are the heart of the test set; one set of filters being required for every frequency measured. Why cannot the whole receiver be eliminated and replaced with a frequency selective voltmeter which may be required for other purposes and/or which may be cheaper? This assumes that the filter in the selective voltmeter has the same characteristics as the one in the white noise receiver. Selective voltmeters are continuously tunable devices which incorporate a local oscillator and mixer to heterodyne any incoming frequency down to a basic I.F. frequency. The filter is at the I.F. frequency. Now, if an NPR of 40 db is to be measured, the level out of the filter is 40 db below the reference level in the filter. It is also 40 db below all other signals in this filter bandwidth if the selective voltmeter were tuned to scan the white noise test signal. In this case the mixer is not only seeing the wanted signal which is 40 db down from the reference but a wide range of signals at a much higher level. The mixer cannot handle this high level wide band signal and produces crosstalk which falls in the bandwidth being measured. This cross talk causes a false NPR reading. If the input level is dropped enough so that the mixer is not overloaded, NPR readings cannot be made since the dynamic range of the selective voltmeter is not large enough. A selective voltmeter may be used for BINR measurements since after the reference level is set all white noise is then removed.

Reference

1. Myron E. Ferguson, "Intermodulation Testing of Multichannel Radio Systems," AIEE Conference Paper #CP 59-1194.

MR. E. C. OTTENBERG, Senior Project Engineer, is a Group Head in the Radio Beam Expansion Project.

Since joining the Radio Beam Expansion Project in 1959 his main areas of responsibility have been Traveling Wave Tubes and Amplifier Klystrons used in the radio systems, radio beam test equipment, and participating in the design of the MLD-4B radio equipment. He has participated in the system testing of various radio links both in manufacturer's facilities and in the field.

Upon joining Western Union in 1951 he participated in the design, system and field testing of the MLD-4A system, in which he was also responsible for the design of the Transmitter and Baseband Combiner.

Mr. Ottenberg received his B.E.E. degree from the College of Engineering at New York University in 1951. He is a member of the IEEE.

Random Notes at the IEEE Winter Meeting—January 1963

In 1963 Winter General Meeting of the IEEE was opened by the Committee Chairman, Mr. W. G. Vieth, Research and Engineering Department of the Western Union Telegraph Company on January 28, 1963. This was the first meeting of the merged societies of the AIEEE and the IRE.

A—General Session

At the General Session Mr. A. C. Monteith of Westinghouse Electric Corporation was awarded the 1962 Edison Medal. In his acceptance speech he pointed out the significance of the merger of the AIEE and the IRE in forming the new professional society of 150,000 engineers—now called the IEEE, the Institute of Electrical and Electronics Engineers. He stressed the new trend in educating engineers — concentration in the fundamentals of electrical engineering and in advanced concepts of science and mathematics. He pointed out industry needs engineers thoroughly grounded in the classic fundamentals, with a saturating comprehension of the basic principles of electrical science and engineering.

Changes are happening so fast in our industry that "the best educated specialist soon finds himself lost with obsolete ideas and knowledge when faced with new equipment whose design and function he must master. The graduate engineer today is reckoned to have a half-life of about 10 years—that is, half of what he knows will be obsolete in a decade. The man with the basic grounding is best able to adapt himself to these fast changes."

B—Transmission Session

The symposium was supposedly held to narrow the language barrier between the theorist and practitioner of communications. While interesting and entertaining at times just what was accomplished is questionable.

The obvious was again emphasized; that the theorist bases his studies on mathematical models which assume certain ideal conditions, while the practitioner must work with real-life models many characteristics of which defy mathematical description. In the latter case the "science" has many aspects of an "art" and intentive reasoning still has its place.

The impression left with this observer was that the gap between theorist and practitioner is real and much effort on the part of both plus a great deal of tolerance will be required to close it.

C—Data Communications Session

One session on data communication switching was devoted almost entirely to the Bell System's recently inaugurated automatic (dial) TWX system. Of particular interest was a paper by E. J. Tyberghein of Bell Telephone Labs. on "The Modernization of TWX." (Paper No. CP 63-581.) The author gave a good general description of the dial TWX system and, to the careful observer, at least, revealed some of the shortcomings of this system as compared to Telex. Another author described the dial TWX station facilities (Paper No. CP 63-522).

One session on data communication and telegraph systems covered a wide variety of subjects. Automatic testing for maintenance purposes in Telex networks was described by H. J. A. Radler of Siemens and Halske AG (Paper No. TP 63-284). N. A. Jacobs of Teletype Corp. described the design innovations and manufacturing techniques used to achieve low cost in the Models 32 and 33 teleprinters. (Paper No. CP 63-462.) Another paper by W. Y. Lang described numerous minor but significant developments in the printing telegraph field by various companies during 1962 (Paper No. CP 63-484).

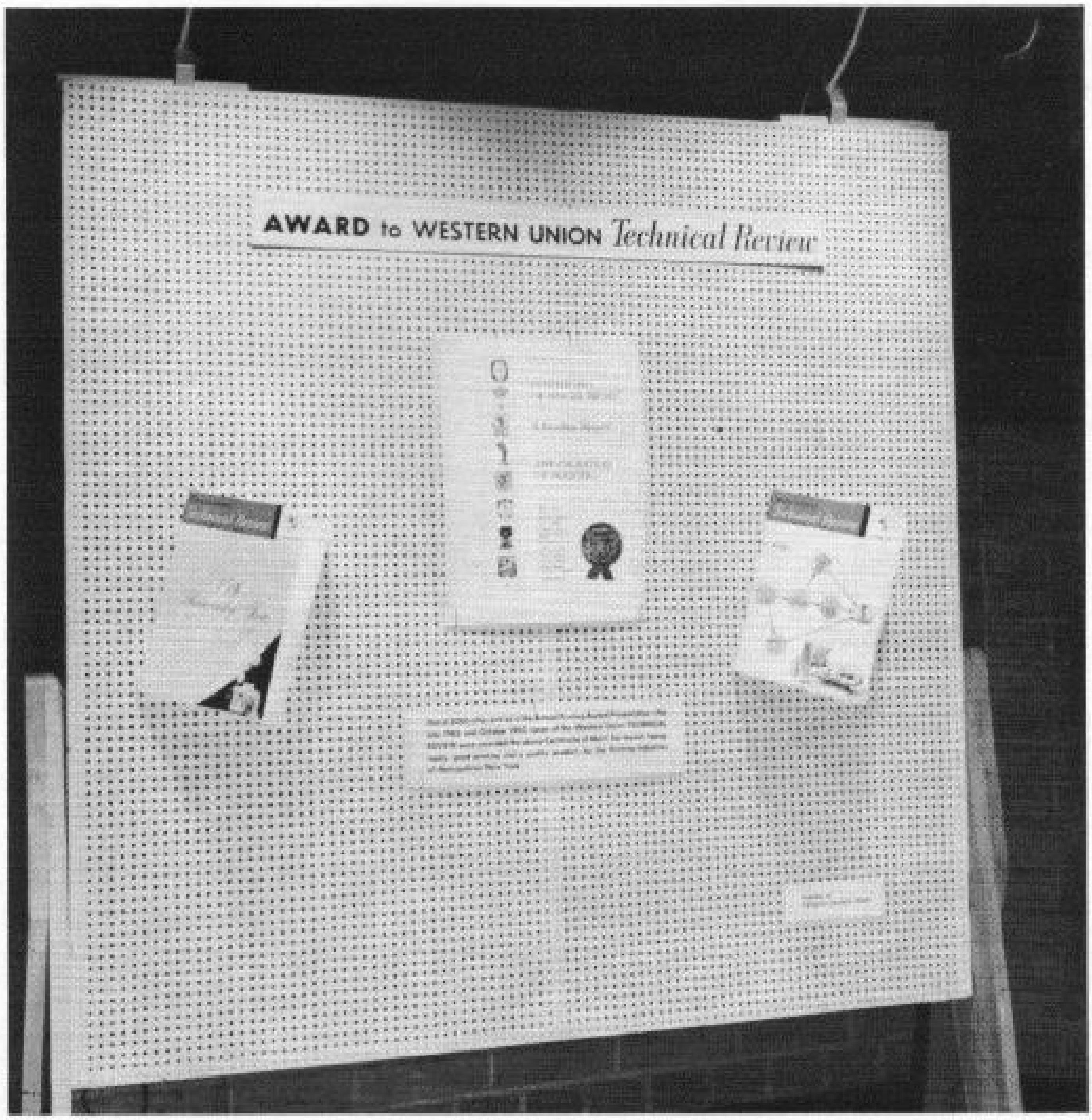
THE EDITOR

Award to the Western Union TECHNICAL REVIEW

Two issues of the Western Union TECHNICAL REVIEW won a Certificate of Merit at the Annual Exhibition of the Printing Industries of Metropolitan New York on January 17, 1963.

The July 1962 and October 1962 issues were selected from 3,000 other entries and judged worthy of honorable mention. The award was given for layout, typography, good printing and a quality product.

The format and layout of the TECHNICAL REVIEW was changed with the July 1962 issue, the 15th Anniversary Issue. This is the first award to Western Union for its technical publication.



Display of Award in Western Union Lobby, 60 Hudson St., N. Y. 13, N. Y.

Radio-Relay Systems
Microwave Measurements
Waveguide Reflections

Arnoff, E.: Measurement of Discontinuities in Waveguides

Western Union TECHNICAL REVIEW, Vol. 17, No. 2 (April 1963). pp 46 to 63

This article describes a frequency-modulation radar method for determining the location and magnitude of waveguide discontinuities. The practical application of this method using commercially available equipment is emphasized. This article was presented as a Transactions paper at the 1963 IEEE Winter General Meeting in January 1963.

Varactors
Semiconductor Devices
Variable-Capacitance Diodes

Ernst, R. L. and Fitzpatrick, J. K.: Varactor Diode Part II—Applications to Microwave Systems

Western Union TECHNICAL REVIEW, Vol. 17, No. 2 (April 1963) pp 65 to 77

The fundamental theory of the varactor diode was explained in Part I, which appeared in the January, 1963 issue of the Western Union TECHNICAL REVIEW.

Part II outlines the most prominent current applications of varactors. Fundamental relations descriptive of the basic circuits are given. The importance of these circuits to microwave system design is explained.

Maintenance Organization
Microwave Systems
Vehicles

Woodman, G. B.: New Ideas in Microwave System Maintenance

Western Union TECHNICAL REVIEW, Vol. 17. No. 2 April 1963) pp 78 to 81

This article is an updating of the maintenance procedures for our microwave system. It briefly covers the changes in organization, vehicular needs, intra-system handling of communications and emergency power requirements.

Test Instruments
Radio Equipment

Ottenberg, E. C.: Delay, Linearity and White Noise Testing of F.M. Radio Relay Systems

Western Union TECHNICAL REVIEW, Vol. 17, No. 2 (April 1963). pp 82 to 89

Delay, Linearity and White Noise are explained in relation to FM radio relay systems. The theory and operation of the test equipment is reviewed. The presentation is for the new worker in the field who may wish to more fully understand the reasons for making these tests and the theory of how the test equipment operates.

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