

TUNNEL

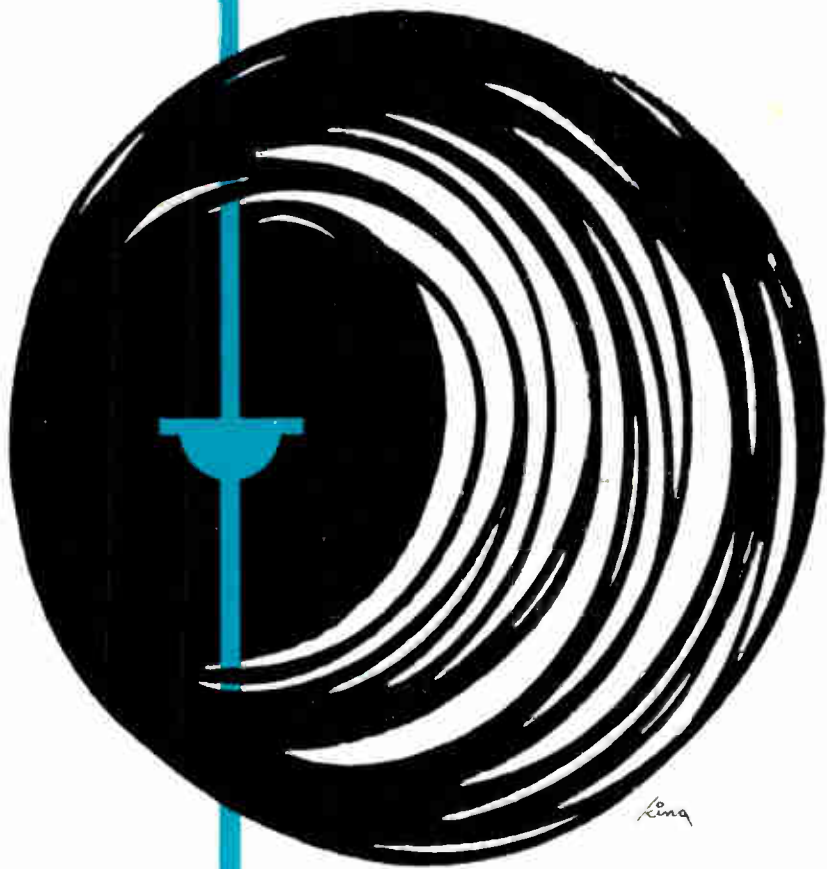


DIODE

MANUAL

CIRCUITS

APPLICATIONS SPECIFICATIONS



Long

GENERAL ELECTRIC

TUNNEL DIODE MANUAL

First Edition

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The circuit diagrams included in this manual are included for illustration of typical tunnel diode applications and are not intended as constructional information. For this reason, wattage ratings of resistors and voltage ratings of capacitors are not necessarily given. Similarly, shielding techniques and alignment methods which may be necessary in some circuit layouts are not included. Although reasonable care has been taken in their preparation to insure their technical correctness, no responsibility is assumed by the General Electric Company for any consequences of their use.

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FOREWORD

The tunnel diode is the newest of continual advancements in semiconductor devices since the introduction of the transistor and silicon controlled rectifier. It is directly responsible for opening up many new areas of application not previously feasible. These additional applications have been made possible by virtue of the tunnel diode's extreme speed (high frequency), stable characteristics that are insensitive to temperature changes, modest power supply requirements, ability to operate in a wide variety of critical environments, low noise level, simplicity, light weight and small size.

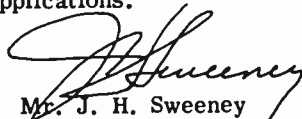
Tunnel diodes will not displace transistors or other active components in existing equipment. In some new circuits this may occur; but generally speaking, tunnel diodes should improve the functional worth of other active components by working with them. Tunnel diodes, together with other semiconductor devices, will make possible the practical design of equipment now either impossible or impractical.

Since General Electric first introduced the tunnel diode in 1959, much energy and talent have been directed toward tremendously improving and greatly increasing the versatility of this newest of semiconductor devices.

An example of this effort is the recent introduction of General Electric's "back" diode and "microwave" tunnel diode. This latter device, housed in a small stripline package barely visible to the naked eye, has an inductance of less than .000,000,000,4 henries. This "microwave" tunnel diode package was designed for microwave communications, radar, very high frequency amplifiers, and oscillator applications.

General Electric again takes pleasure in initiating another semiconductor reference, the Tunnel Diode Manual, with the objective of extending the same service to the industry as has the G-E transistor manual since its introduction in 1957.

This manual, then, is one of our contributions to the better understanding of tunnel diodes and their circuitry for use in low power, ultra-high frequency, low cost applications.



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TO-18



STRIP LINE

G-E TUNNEL DIODES

CHAPTER I

THEORY

The tunnel diode is a revolutionary new semiconductor device. Although it can perform many of the functions of conventional devices its principles of operation are entirely different from those of other semiconductor devices and vacuum tubes.

Such conventional amplifying devices as transistors and vacuum tubes depend on emitting a charge carrier into a region where its motion can be influenced by a signal electrode, and on subsequently collecting the charge carrier on an output electrode. The speed of this process is limited by the time it takes a charge carrier, having left the emitter, to traverse the control region, and appear on the collector.

The basic conduction mechanism in tunnel diodes, however, has a theoretical frequency limit of 10^7 megacycles per second which is several orders of magnitude higher than the drift and diffusion mechanism involved in the operation of conventional diodes and transistors.

1.1 The Tunnel Effect

The tunnel diode takes its name from the tunnel effect--a process wherein a particle (obeying the laws of the quantum theory) can disappear from one side of a potential barrier and appear instantaneously on the other side, even though it does not have enough energy to surmount the barrier. It is as though the particle can "tunnel" underneath the barrier.

In the case of the tunnel diode, the barrier is the space charge depletion region of a p-n junction. This is the same barrier which prevents the current from flowing in the reverse direction in the case of the ordinary rectifier diode. In the tunnel diode, this barrier is made extremely thin (less than a millionth of an inch) -- so thin, in fact, that penetration by means of the tunnel effect becomes possible. This gives rise to an additional current in the diode at very small forward bias which disappears when the bias is increased. It is this additional current that produces the negative resistance in a tunnel diode.

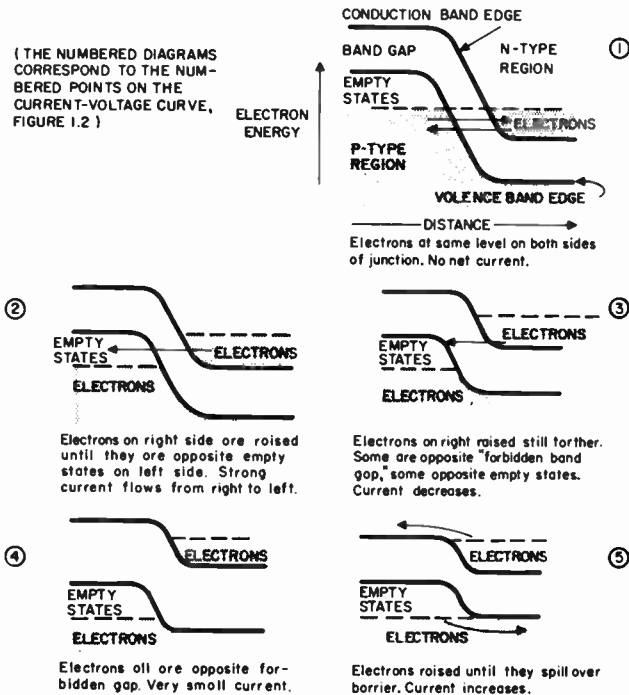
The origin of the tunnel current can be qualitatively understood by considering the changes in the characteristics of a conventional p-n junction diode made with higher and higher concentrations of free carriers in the semiconductor crystal. As the density of the charge carriers is increased, the reverse breakdown voltage decreases. It might be thought that there would be a limiting case when the reverse breakdown voltage was reduced to zero. This is not correct, however; the limit is determined by the solubility of the impurities which determine the carrier concentrations. Experiments have shown that it is

possible to dope many semiconductor materials more heavily than is needed to reduce the breakdown voltage to zero. If this heavy doping is used, the diode can still be in the reverse breakdown condition at a slight forward bias. When a larger forward voltage is applied, the diode goes out of the reverse breakdown condition and the current falls to a small value.

1.2 Physical Description of a Tunnel Diode Junction

A semiconductor has a forbidden region where there are no states available for its electrons. This region is called the band gap. The states below this gap (which comprise the valence band) are almost all filled. The states above it (the conduction band) are almost all empty. The number of empty states in the valence band, or electrons in the conduction band, can be controlled by adding either acceptor impurities or donor impurities to the semiconductor crystal. Each acceptor impurity takes one electron out of the valence band, and each donor gives one electron to the conduction band. In this way p-type (empty states in valence band) and n-type (electrons in conduction band) regions can be built into a crystal. The intersection of these two regions is called a p-n junction.

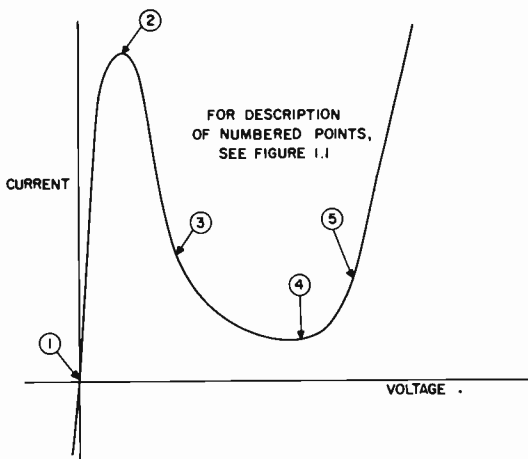
Figure 1.1 and 1.2 represent the conduction and valence bands in



TUNNEL DIODE JUNCTION AT VARIOUS BIAS CONDITIONS

FIGURE 1.1

the vicinity of a junction at different values of applied bias. It can be seen that as the bias is increased the bands which overlap each other at zero bias become uncrossed. Since tunneling is represented by a horizontal transition on this picture, the current decreases as the bands become uncrossed.



TUNNEL DIODE CHARACTERISTIC

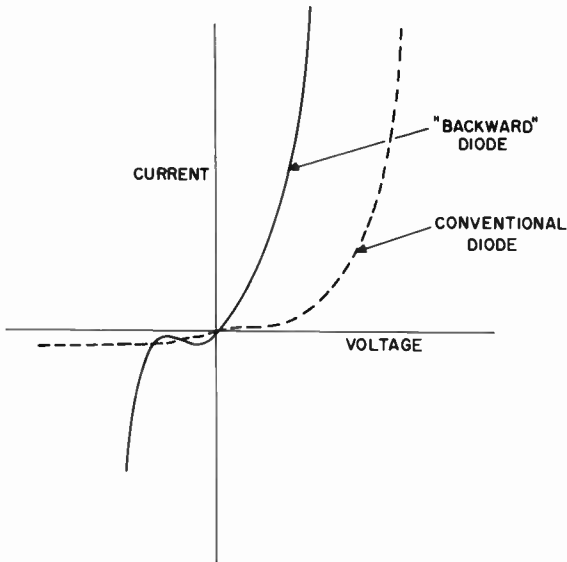
FIGURE 1.2

1.3 Backward Diodes*

In a tunnel diode, the voltages corresponding to points #2 and #4 in Figure 1.2 are essentially determined by the semiconductor material (germanium, silicon, or gallium arsenide) used in the process. The current value corresponding to point #2 is determined, however, by the cross-section area and doping level of the junction. In practice the tunnel junction is made larger in cross-section area than necessary and then etched until the peak current (point #2 in Figure 1.2) has the desired value. Tunnel diodes can be made with peak currents anywhere from a few microamps to several amperes. Tunnel diodes with peak currents of $500 \mu\text{a}$ and higher are usually used as amplifying or switching elements in circuits that take advantage of the fact that with increasing forward bias, the tunnel current first increases and then decreases.

It will be noted in Figure 1.2 that when voltage is applied to a tunnel diode, in the reverse direction, a large current flows which increases continuously as the voltage is increased. If a negative voltage of the magnitude of point 4 in Figure 1.2 were applied to a tunnel diode, the resulting current would be many hundreds of times higher than when this voltage is applied in the positive direction. A tunnel diode used to take advantage of this large change in current with polarity of applied voltage is called a "backward diode". The term "backward" meaning that the diode conducts heavily with negative rather than positively applied voltage. Figure 1.3 shows a comparison of germanium "backward" and conventional diodes.

* "Backward" diode is also referred to as "back" diode throughout this edition.



**COMPARISON BACKWARD AND CONVENTIONAL
DIODE CHARACTERISTIC**

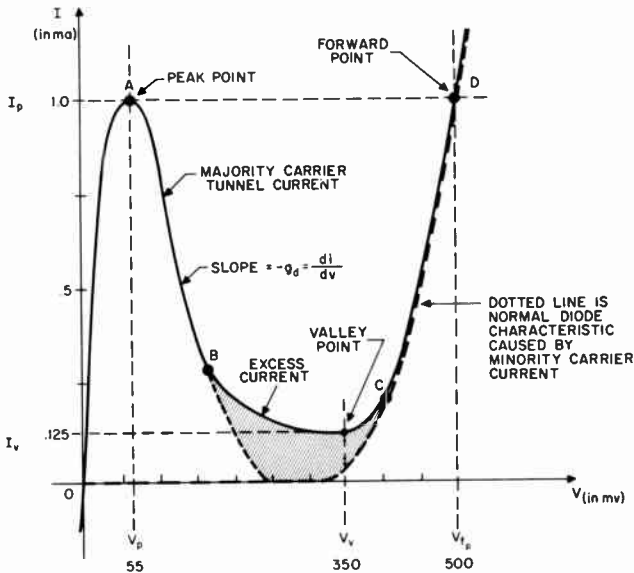
FIGURE 1.3

It can be seen that a backward diode has a lower voltage drop at a given current than a conventional diode. This low forward drop is very advantageous in tunnel diode and some transistor circuits. Backward diodes are designed to have a low ($50 \mu\text{a}$ or less) peak point currents since current flow with forward bias is undesirable in backward diodes for the same reasons that leakage current with reverse applied voltage is undesirable in conventional diodes. In some types of tunnel diode circuits, however, it may be desirable to have a low peak point current and in this case a backward diode can be used.

CHAPTER 2

RATINGS AND CHARACTERISTICS

The voltage-current characteristic of a germanium tunnel diode is shown in Figure 2.1 together with the important DC parameters. The dotted line in this figure shows a normal diode characteristic resulting from minority carrier current. It is seen that the tunnel diode follows this characteristic beyond point C. In the lower voltage region below point C and in the reverse biased state the diode current consists of majority carriers which tunnel through the narrow PN junction.



STATIC CHARACTERISTIC CURVE OF GERMANIUM TUNNEL DIODE

FIGURE 2.1

Electrical Characteristics

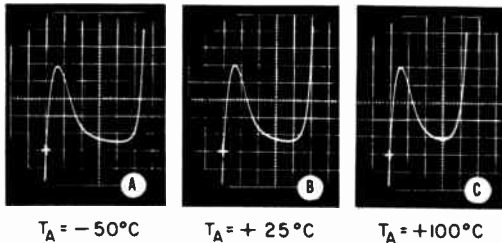
A "relatively" linear negative conductance region exists between point A (the peak point) and point B in Figure 2.1. Between point B and point C, the current is greater than the sum of the theoretical majority and minority currents. The current in this region, identified as the excess current, can not, as yet, be completely explained. Intuitively the excess current or valley current should be low and therefore the highest peak point to valley point current ratio seems desirable. There are some tangible reasons for this also. The greater this ratio, for any given value of peak point current, the greater will be the available output current swing. For example, a tunnel diode with a peak current of one milliampere and a peak to valley current ratio of 8 will have an available current swing of $1.0 - 0.125 = 0.875$ ma.

The "peak point current" of a tunnel diode is determined by an etching process and can be held to within $\pm 2.5\%$ or better on a production basis. However, the peak point voltage, V_p , valley point voltage, V_v , and forward point voltage, V_{fp} , are determined by the semiconductor material and are largely fixed. For germanium these voltages are respectively 55 mv, 350 mv and 500 mv typical at 25°C .

The magnitude of the negative conductance is equal to the slope di/dv of the voltage current characteristic. For a one milliamper germanium tunnel diode the negative conductance is between 0.006 and 0.010 mhos corresponding to a negative resistance between 100 and 160 ohms.

Temperature Characteristics

Variation of the tunnel diode parameters with temperature is a matter of extreme importance to the circuit designer. Figure 2.2 shows the voltage-current characteristic of a typical germanium tunnel diode at temperatures of -50°C , 25°C , and 100°C .



**VOLTAGE-CURRENT CHARACTERISTIC CURVES
OF A TYPICAL GERMANIUM TUNNEL DIODE**

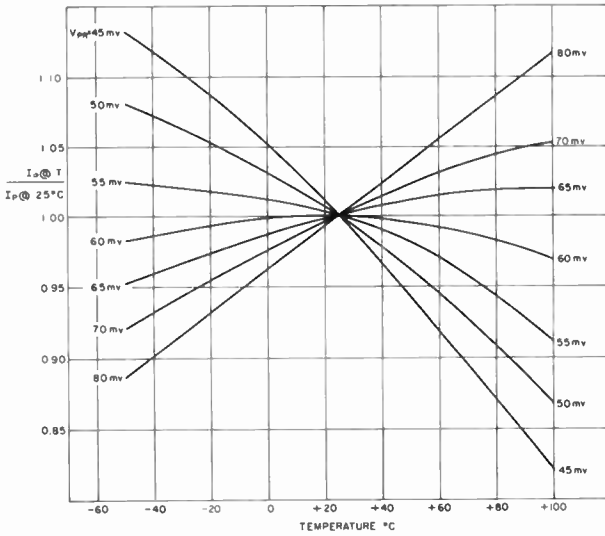
FIGURE 2.2

Note that the peak voltage, valley voltage and forward voltage all decrease with increasing temperatures while the valley current increases with increasing temperature. The peak current may increase or decrease with temperature depending on the doping level and the resistivity of the semiconductor material. This can best be illustrated by Figure 2.3 showing the temperature variations of the peak current for units with different peak voltages. V_p exhibits good correlation with the doping level and hence with the temperature characteristic of the diode.

The parameter V_p is dependent on the dopants and resistivity used. For the diode shown in Figure 2.2, the peak current is a maximum at approximately 25°C and decreases at higher and lower temperatures.

Each application generally has a different temperature problem. For example, in switching circuits the primary concern might be the stability of the peak current since it determines the switching threshold, although the changing forward voltage can affect the amplitude of the output voltage pulse.

In oscillators where direct matching is not required, it may be important only to make sure that at the lowest operating temperatures the device is driven from a voltage source which requires that the re-



VARIATION OF NORMALIZED I_p WITH TEMPERATURE FOR DIFFERENT VALUES OF V_{pp}

FIGURE 2.3

sistance of the source supplying the voltage to the tunnel diode is much less than the negative resistance of the diode. Oscillators have been operated successfully over a temperature range from 4°K to over 573°K, a remarkably wide operating range. In amplifiers where some degree of match between the diode negative conductance and the positive circuit conductances is required, it is obvious that this match must be maintained over the required operating temperature range.

The variation of the important DC parameters between -50°C and +100°C is shown in Figure 2.4 for 1 ma germanium tunnel diodes of the 1N2939 or 1N2940 types.

CHARACTERISTIC	SYMBOL	1N2939	1N2940
PEAK POINT VOLTAGE	V_p	55 MV	55 MV
TEMPERATURE COEFFICIENT	$\Delta V_p / \Delta T$	60 $\mu V / ^\circ C$	60 $\mu V / ^\circ C$
VALLEY POINT VOLTAGE	V_v	350 MV	350 MV
TEMPERATURE COEFFICIENT	$\Delta V_v / \Delta T$	-1.0 MV / °C	-1.0 MV / °C
FORWARD POINT VOLTAGE AT PEAK CURRENT	V_{FP}	500 MV	500 MV
TEMPERATURE COEFFICIENT	$\Delta V_{FP} / \Delta T$	-1.0 MV / °C	-1.0 MV / °C
PEAK TO VALLEY RATIO	I_p / I_v	8 MIN	5 MIN
VALLEY POINT CURRENT TEMPERATURE COEFFICIENT	$\Delta I_v / \Delta T$	+0.75 % / °C	+0.75 % / °C
CONDUCTANCE TO PEAK CURRENT RATIO	g_d / I_p	9.0 MHO / AMP	6.5 MHO / AMP
CONDUCTANCE TEMPERATURE COEFFICIENT	$\Delta g_d / \Delta T$	-0.5 % / °C	-0.5 % / °C
CAPACITANCE TO PEAK CURRENT RATIO	C / I_p	6 pF / ma	4 pF / ma

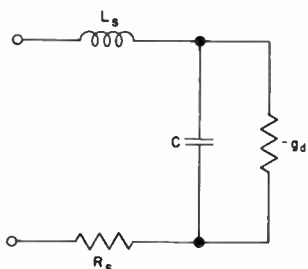
TYPICAL ELECTRICAL CHARACTERISTICS OF GERMANIUM TUNNEL DIODE

FIGURE 2.4

Note that the peak point voltage has a temperature coefficient of -0.06 millivolts/ $^{\circ}\text{C}$ and the forward point voltage has a temperature coefficient of -1.0 millivolts/ $^{\circ}\text{C}$ as compared with a value of -2.5 millivolts/ $^{\circ}\text{C}$ for the forward drop of a conventional diode or transistor. For detailed specifications of individual tunnel diode types, see the specification sheet section of this manual.

Frequency Limitations

The small signal (AC) equivalent circuit for the tunnel diode biased in the negative conductance region is shown in Figure 2.5. The inductance, L_s , in the equivalent circuit is relatively low and is determined primarily by the inductance of the leads. A small amount of series resistance, R_s , is also present which is determined by the bulk resistance of the semiconductor material. The capacity, C , is primarily due to the capacity of the junction although a small portion of the capacity is due to the leads and the package. As the junction capacity is decreased for lower current units, the package capacitance becomes more and more important however. The negative conductance, $-g_d$, in the equivalent circuit is equal to the slope of the voltage-current characteristic at the particular bias point under consideration. The value of the negative conductance can be assumed to be independent of frequency, the chief limitations in the frequency response of the tunnel diode being determined by the parasitic elements in the equivalent circuit (R_s , L_s , C).



TYPICAL VALUES OF IN2939 PARAMETERS	
SERIES INDUCTANCE, L_s	$6\text{m}\mu\text{h}$
TOTAL CAPACITANCE, C	$7\mu\mu\text{fd}$
SERIES RESISTANCE, R_s	$1.5\ \Omega$
NEGATIVE CONDUCTANCE, $-g_d$	$.006 - .01\text{mho}$
NEGATIVE RESISTANCE, r_d	$100 - 150\ \Omega$
PEAK POINT CURRENT, I_p	1ma
VALLEY POINT CURRENT, I_v	$.1\text{ma}$
PEAK POINT VOLTAGE, V_p	55mv
VALLEY POINT VOLTAGE, V_v	350mv
FORWARD POINT VOLTAGE, V_f	500mv
* 1/8 INCH LEAD LENGTH	

**SMALL SIGNAL EQUIVALENT CIRCUIT AND
TYPICAL VALUES OF PARAMETERS**

FIGURE 2.5

Two significant frequency figures of merit can be assigned to the tunnel diode:

(a) resistive cut-off frequency, $f_{ro} = \frac{|g_d|}{2\pi C} \sqrt{\frac{1}{R_s |g_d|} - 1}$ (2.1)

$$(b) \text{ self-resonant frequency } f_{x0} = \frac{1}{2\pi} \sqrt{\frac{1}{L_s C} - \left(\frac{g_d}{C}\right)^2} \quad (2.2)$$

Both these frequencies are derived from the equivalent circuit of Figure 2.5. The resistive cut-off frequency is the frequency at which the real part of the diode impedance measured at its terminals goes to zero. The tunnel diode can not amplify above this frequency. The self-resonant frequency is the frequency at which the imaginary part of the diode impedance goes to zero. It should be pointed out that both frequencies are reduced by external circuit components and therefore the highest possible operating frequency is very circuit dependent. In a transistor package the tunnel diode is limited to frequencies below 1 Kmc, this limit being due primarily to the lead inductance. Micro-strip or microwave packaging, owing to its inherently lower inductance, can potentially raise the frequency capabilities by an order of magnitude or more.

Nuclear Radiation Effects

Encouraging results have been obtained from preliminary investigations of the effects of nuclear radiation on the characteristics of tunnel diodes. Under a dosage of 3×10^{14} NVT (90% thermal, 10% fast), no apparent change in the electrical characteristics was observed except for the noise figure which increased by approximately 20% at the point of maximum negative conductance and by 100% near the valley point.

At a dosage of 5×10^{15} NVT, the valley current increased by about 25% while the other DC characteristics had not changed. The noise figure increased by a factor of 3 at the point of maximum negative conductance while the noise figure in the vicinity of the valley point was immeasurably high.

A Battelle Memorial Institute report¹⁶ shows that a 10 ma I_p germanium tunnel diode with an initial valley point current of 1 ma had a valley point current of 3.6 ma at a radiation flux of 2.4×10^{16} NVT (fast neutron/cm²). Above this flux-density large changes of valley current would occur for small increases in flux. It should be pointed out that although I_v increases while both V_v and V_{fp} decrease with flux density, the characteristic seems to remain unchanged in the reverse direction (important for backward diodes). The same holds true for the positive slope region below the peak. I_p and V_p are totally unaffected while the negative conductance slope does not appear to change radically up to flux levels of 5.5×10^{16} NVT (fast neutrons/cm²).

Maximum Ratings

The "Absolute Maximum Ratings" of tunnel diodes are primarily dependent on the junction temperature produced by the current flowing through it, the physical size of the junction area and the efficiency with which heat is removed.

The high current density material used in tunnel diodes to provide good g_d/C ratios causes the size of the junction to be very small. A one milliampere unit, for instance, might have a junction diameter of less than 3×10^{-4} inch. A large current flowing through such a small cross-sectional area can rapidly create enough heat to melt the alloys used.

As large currents can flow through the tunnel diode junction even at low forward or reverse voltages, an absolute maximum current specification (forward and reverse) is more meaningful than voltage or dissipation limits per se.

Also given in the "Absolute Maximum Ratings" are the storage and operating junction temperature ranges for the device. Actually the device proper can operate from liquid helium temperatures (4°K) to the melting point of the alloys. The package might however limit this operating range, as the glass seals might become leaky or even crack at very low temperatures. Even under those conditions, the device might continue to function as it does not necessarily have to be hermetically sealed. Most packages will readily operate over a -50°C to $+100^{\circ}\text{C}$ range or better.

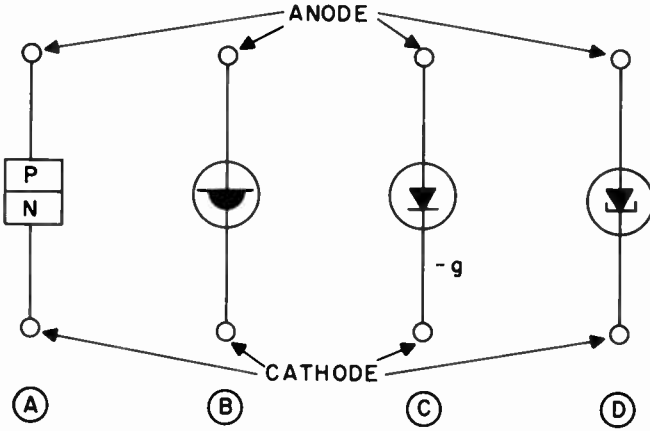
Symbols

Several different symbols for tunnel diodes are in current use. The more popular symbols are shown in Figure 2.6 together with the PN junction equivalent. Many of the other symbols which have been used in the literature are not recommended since they are in direct contradiction of AIEE/IRE standards on graphical symbols.

The anode and cathode designations for each of the symbols are indicated in Figure 2.6. If the anode is positively biased with respect to the cathode the tunnel diode will display its negative conductance characteristic and the high voltage injection characteristic. Figure 2.6a shows schematically the PN junction structure of the device. Figure 2.6b shows the symbol which is used exclusively in this manual. The polarity is easily remembered by imagining the arrow on a conventional diode to "tunnel" through the bar after which it appears on the other side of the bar in the blunt, rounded form as shown. This symbol has the advantage that it can be combined with other symbols to indicate complex devices. For example, a PNP transistor with a tunnel emitter can be indicated by adding a semicircle to the conventional symbol for a PNP transistor. The symbol shown in Figure 2.6c indicates a tunnel diode by means of a -g next to the symbol signifying that the device is a negative conductance device. This symbol tends to imply that the device is used for small signal operation and hence is generally used in amplifiers and low power oscillators. Note that the -g can not be used inside the circle without violating IRE standards. The symbol shown in Figure 2.6d is frequently advocated because it contains the letter T in its structure.

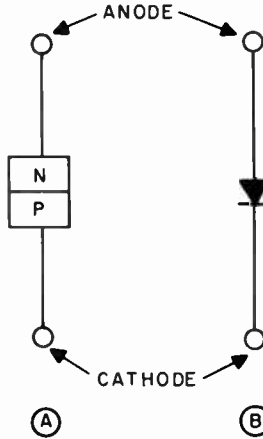
The symbol for the back diode (backward diode) is shown in Figure 2.7 together with the PN junction equivalent. Since the back diode can be considered to be a conventional diode with a low reverse breakdown voltage a conventional diode symbol can be used. Note that the anode is connected to the N material and the cathode is connected

to the P material since the direction of easy current flow is opposite to that of a conventional diode.



COMMONLY USED SYMBOLS FOR THE TUNNEL DIODE

FIGURE 2.6



PREFERRED SYMBOL FOR THE BACK DIODE

FIGURE 2.7

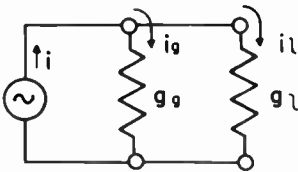
CHAPTER 3

TUNNEL DIODE AMPLIFIERS

Introduction

Since by definition a positive conductance dissipates energy, it follows that a negative conductance generates energy. This is the basis of negative conductance (or resistance) amplifiers.

Consider first the case of an ideal sinusoidal generator, having an internal conductance g_g , delivering a rms current i to a load g_l (see Figure 3.1).



$$P_l \text{ max} = \left(\frac{i}{2}\right)^2 \cdot \frac{1}{g_l} = \frac{i^2}{4g_g} \quad \left\{ \begin{array}{l} g_g = g_l \\ P.G._{AV} = \frac{P_l}{i^2/4g_g} = 1 \end{array} \right.$$

POWER TRANSFER FROM GENERATOR TO LOAD

FIGURE 3.1

Suppose g_l is varied to maximize the power delivered to it. The latter occurs under a "matched" condition, namely when $g_g = g_l$. Under this condition, the generator current splits evenly between the two branches of the circuit. The maximum power delivered to the load is thus given by:

$$P_l \text{ max.} = \left(\frac{i}{2}\right)^2 \cdot \frac{1}{g_l} = \frac{i^2}{4g_g} \quad (3.1)$$

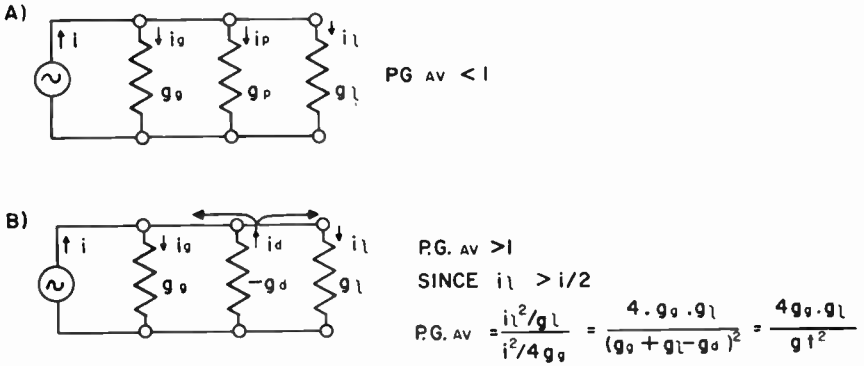
As this is the maximum available power from the generator, one might use this value to find the "transducer gain", defined as the power delivered to an arbitrary load divided by the power available from the generator.

$$P.G. \text{ avail.} = \frac{P_l}{i^2/4g_g} \quad (3.2)$$

This gain can obviously not exceed unity.

Suppose that a positive conductance g is placed in parallel with the generator and load (see Figure 3.2a). Part of the generator current flows into this positive conductance and the transducer gain will be less than unity. If, however, a negative conductance ($-g$) replaces the positive shunt conductance (as in Figure 3.2b), the current flowing in the added branch is out of phase to that of the previous case

and hence current is coming out instead of going into this negative conductance branch. Hence this branch acts as a generator and drives additional current into the circuit. The transducer gain can now exceed unity since the load current i_l can now be greater than $i/2$.



POWER TRANSFER HAVING NEGATIVE CONDUCTANCE IN CIRCUIT

FIGURE 3.2

By proper adjustment of the load, this gain may be infinite as can be seen from the following equation:

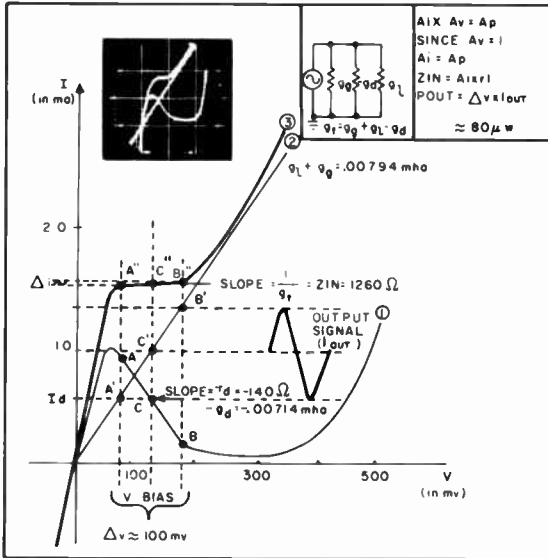
$$P.G. \text{ avail.} = \frac{P_{load}}{P_{av. gen.}} = \frac{i_l^2/g_l}{i^2/4g_g} = \frac{4 \cdot g_g \cdot g_l}{(g_g + g_l - g_d)^2} \quad (3.3)$$

As the sum of the positive conductances ($g_g + g_l$) is made equal to the negative conductance $-g_d$, the denominator becomes zero and the P. G. infinite. Hence, the degree of "Matching" between $g_g + g_l$ and $-g_d$ will determine the gain. If $g_g + g_l$ is smaller than $-g_d$, the denominator is negative and the circuit unstable. Thus for stable values of gain $g_g + g_l$ must be greater than $-g_d$ and for this gain to be large $g_g + g_l$ must only be slightly larger than $-g_d$. Figure 3.3 shows a graphical representation of the tunnel diode V-I curve, the load line and the resultant parallel impedance.

The tunnel diode capacity is in parallel with the negative conductance and as the frequency increases will become a more and more effective shunt. In a parallel type amplifier, an inductance L is put across the tunnel diode in order to resonate the circuit at a frequency f_o , where:

$$f_o = \frac{1}{2\pi \sqrt{L C}} \quad (3.4)$$

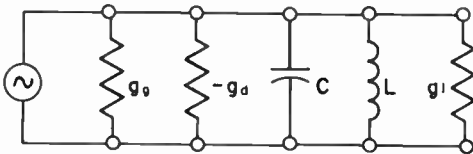
(see Figure 3.4).



GRAPHICAL ANALYSIS OF PARALLEL AMPLIFIER STAGE

FIGURE 3.3

TRANSDUCER GAIN:



$$P.G. = \frac{4 g_o g_l}{g_t^2 + \omega^2 C^2 \left(1 - \frac{\omega \omega_0^2}{\omega^2}\right)^2}$$

AT RESONANCE $\omega_0 = \omega$ AND

$$P.G. = \frac{4 g_o \cdot g_l}{g_t^2}$$

EQUIVALENT CIRCUIT OF A TUNNEL DIODE "PARALLEL" AMPLIFIER

FIGURE 3.4

The generalized gain equation for this circuit now becomes:

$$P. G. \text{ avail.} = \frac{4 g_g g_l}{g_t^2 + \omega^2 C^2 \left(1 - \frac{\omega_0^2}{\omega^2}\right)^2} \quad (3.5)$$

and since at resonance $\omega = \omega_0$, the available power gain at resonance is still:

$$P. G. \text{ avail.} = \frac{4 g_g g_l}{g_t^2} \quad (\text{as per equation 3.4})$$

where $g_t = g \text{ total} = (g_g + g_l - g_d)$ (3.6)

From equation (3.5), it can be seen that gain decreases "off-resonance" and since bandwidth is defined as the difference between the two side band frequencies at which the gain is 3 db down, bandwidth (BW) can be defined as:

$$BW = \frac{\omega_2 - \omega_1}{2\pi} = \frac{g_g + g_1 - g_d}{2\pi C} = \frac{g_t}{2\pi C} \tag{3.7}$$

From equations (3.5) and (3.7), it can be seen that as the gain is increased (by decreasing g_t) the bandwidth decreases, becoming zero at the point where g_t is zero, which coincides with the point of infinite gain (oscillations).

The gain-bandwidth product of the circuit of Figure 3.4 can be expressed fully in terms of the circuit parameters as follows:

$$\sqrt{P. G.} \cdot BW = \sqrt{\frac{g_g g_1}{\pi C}} \tag{3.8}$$

if $g_g + g_1$ is approximately matched to $-g_d$, ($g_g + g_1 \approx |-g_d|$), then:

$$\sqrt{P. G.} \times BW = \sqrt{\frac{g_1 (g_d - g_1)}{\pi C}} \tag{3.9}$$

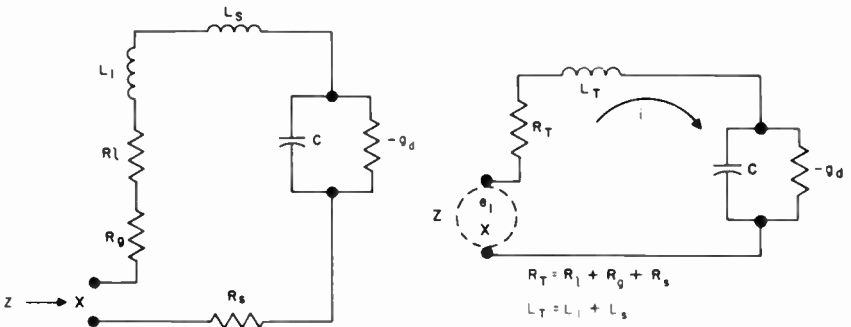
and hence the gain bandwidth product will be largest when $g_1 = g_g = g_d/2$, at which time it is equal to:

$$\sqrt{P. G.} \times BW = \frac{g_1}{\pi C} = \frac{g_g}{\pi C} = \frac{g_d}{2\pi C} \tag{3.10}$$

Other factors influencing the choice of g_g and g_1 for a given g_d are discussed in the section on amplifier noise.

Stability Criteria of Tunnel Diode Amplifiers

Successful linear operation of a tunnel diode amplifier depends on the stability of the complete system, including in particular the internal impedance of the bias supply and the signal source impedance. Con-



(A) SERIES CONFIGURATION OF AMPLIFIER OR OSCILLATOR CIRCUIT

(B) SIMPLIFIED SERIES CONFIGURATION

FIGURE 3.5

sidering the equivalent circuit of Figure 2.5 a basic "series" amplifier circuit can be reduced to that shown in Figure 3.5 where $R_T = R_g + R_1 + R_s$, $L_T = L_s + L_1$, C is the total diode capacity and $-g_d$ the negative conductance of the diode at the operating current and voltage.

To determine the system stability, one can examine the distribution of poles or zeros of the circuit determinant in the complex S-plane.

If the zeros of Z seen at the input fall in the right half of the S-plane, the system is unstable. Conversely, if the zeros fall in the left half side of the S-plane, the circuit is stable. The input impedance is given as:

$$Z(S) = \frac{S^2 L_T C + S (R_T C - L_T |-g_d|) + (1 - R_T |-g_d|)}{SC - |-g_d|} \tag{3.11}$$

and the zeros are:

$$S = -\frac{1}{2} \frac{R_T}{L_T} - \frac{|-g_d|}{C} \pm \sqrt{\frac{1}{4} \left(\frac{R_T}{L_T} - \frac{|-g_d|}{C} \right)^2 - \frac{1 - R_T |-g_d|}{L_T C}} \tag{3.12}$$

then S will have a negative real part only if both:

$$\frac{R_T}{L_T} - \frac{|-g_d|}{C} > 0 \tag{3.13}$$

$$\text{and } 1 - R_T |-g_d| > 0 \tag{3.14}$$

This can be rewritten as:
$$\frac{1}{|-g_d|} > R_T > \frac{L_T |-g_d|}{C} \tag{3.15}$$

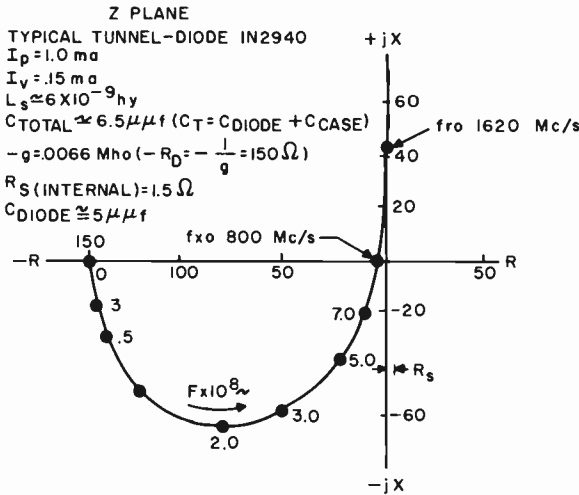
Frequency Limitation

A Nyquist plot of the real and imaginary component of the input impedance divulges two characteristic frequencies. $f_1 = f_{ro}$ is the frequency at which the real part of the circuit impedance goes to zero and $f_2 = f_{xo}$ is the frequency at which the imaginary component of the circuit impedance goes to zero.

$$f_{ro} = \frac{|-g_d|}{2 \pi C} \sqrt{\frac{1}{|-g_d| R_T} - 1} \tag{3.16}$$

$$f_{xo} = \frac{1}{2\pi} \sqrt{\frac{1}{L_T C} - \frac{g_d^2}{C^2}} \tag{3.17}$$

Figure 3.6 is the Nyquist plot of a typical 1N2940 (device only) where R_T and L_T are replaced by R_S and L_S in the above equation.



CHARACTERISTICS OF A TYPICAL TUNNEL DIODE

FIGURE 3.6

Now several conditions could prevail:

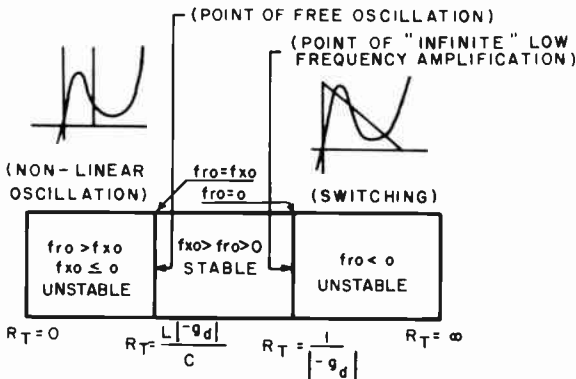
- (1) $f_{ro} < 0$, the device can only switch as $R_T \geq 1/-g_d$.
- (2) $f_{xo} < 0$, the device can only operate in the relaxation oscillation mode.
- (3) If $f_{ro} > f_{xo}$, then at frequencies up to f_{xo} , the circuit sees some real and imaginary values of negative impedance and hence it can oscillate in a relaxation mode.
- (4) If $f_{ro} = f_{xo}$, then at this point both the real and imaginary parts of the input impedance are zero, and the circuit once shocked into oscillation by any random noise or transient will yield free sinusoidal oscillations.
- (5) If $f_{xo} > f_{ro}$ the circuit will be stable and amplification can take place, since the imaginary term of the input impedance does not go to zero until beyond f_{ro} at which time the circuit is already "cut off".

Figure 3.7 is a graphical illustration of the above conditions. In this figure the two roots $R_T = L g_d / C$ and $R_T = 1/-g_d$ are taken from equations 3.11, 3.13, and 3.14.

Summary of Gain and Stability Conditions

The successful design of a tunnel diode amplifier circuit depends on the proper application of the stability criteria. The amplifier circuit must meet the following criteria:

- (1) First, the circuit must be biased stably in the negative conductance region from a voltage source ($R_T < 1/-g_d$ and $f_{ro} > 0$).
- (2) Second, the circuit inductance must be smaller than $1 < R_T C / |-g_d|$ ($f_{xo} > f_{ro}$).



CONDITION FOR STABLE OPERATION SHOWN GRAPHICALLY

FIGURE 3.7

- (3) To obtain large values of stable gain the sum of the positive conductances ($g_g + g_1 + g_x$) must be always only slightly larger than the negative conductance of the diode. ($g_g + g_1 + g_x \geq -g_d$), which also means that the greatest value of gain is obtained very nearly when $R_T = 1/-g_d$.
- (4) The greatest gain bandwidth product is obtained when the load conductance is approximately equal to the generator conductance; in which case, they are both approximately equal to one-half of the diode negative conductance. ($g_1 = g_g = -g_d/2$). At that point $\sqrt{P.G.} \times B.W. = g_d/2 \pi C$. This also points out that to obtain greater gain bandwidth product, diodes with the largest possible figure of merit (g_d/C) should be used.
- (5) The above condition does not take into consideration amplifier noise figure, temperature and bias variations as well as certain other practical considerations explained a little later.

Noise Considerations

It is generally known that the negative resistance of a tunnel diode exhibits shot noise. A tunnel diode also has a parasitic series resistance exhibiting thermal noise. At very low frequencies (in the audio range) the tunnel diode also exhibits $1/f$ noise. Despite these noise sources, tunnel diode amplifiers can be made with noise figures lower than 3 db.

Two requirements must be met to obtain a low noise figure tunnel diode circuit. First, the tunnel diode must itself exhibit low noise and secondly, the circuit must be optimized for low noise performance. The tunnel diode's basic performance can be described by a noise figure of merit which is a function only of the diode parameters, its bias point and the operating frequency.

When the diode is used as an amplifier, the circuit noise figure can be made to asymptotically approach the noise figure of merit of the diode. The latter is normally the minimum limit. The amplifier's

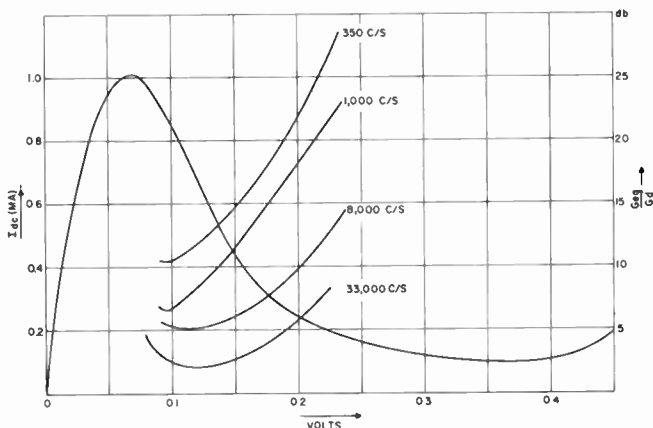
noise figure is made to approach the figure of merit of the diode by suitable adjustment of circuit parameters.

E. G. Nielsen's⁷ paper on the noise performance of tunnel diodes equates the noise figure of merit of the diode as being:

$$F = \frac{\left(1 + \frac{g_{eq}}{g_d}\right)}{\left(1 - \frac{R_S}{|R_D|}\right) \left[1 - \left(\frac{f}{f_{ro}}\right)^2\right]} \quad (3.18)$$

In this equation the ratio of g_{eq}/g_d assesses the shot noise relative to the diode negative conductance. At frequencies above the audio range, where $I_{eq} = I_{DC}$ (DC bias current through the tunnel diode) g_{eq} is equal to approximately $20 I_{DC}$. At lower frequencies, I_{eq} is greater than I_{DC} ($g_{eq} = eI_{eq}/2KT$) and the results can be seen in Figure 3.8 where g_{eq}/g_d is plotted at various frequencies thereby exhibiting the $1/f$ noise contribution of the diode. At 33 Kc/s (which was not quite the break frequency) at a bias of 125 mv, $g_{eq}/g_d = 2$ db which would yield a total minimum noise figure $F = 4.2$ db. The ratio of R_S/R_D assesses the parasitic series diode resistance relative to the negative resistance and hence this would assess the thermal noise. At frequencies near the device "cut off" frequency (f_{ro}) where the real part of the negative resistance becomes very small, the ratio of R_S/R_D becomes quite important. Thus the last ratio assesses the frequency dependence of R_D .

With germanium tunnel diodes $R_S/R_D \approx .01$ and $g_{eq}/g_d \approx 1$. Using these figures a device noise figure of close to 3.0 db can be obtained at medium frequencies, but at $f = .707 f_{ro}$ the minimum would go to 6 db while it would exceed 10 db at $f = .9 f_{ro} \dots f_{ro}$ in this case is the device "cut off" frequency and hence in equation 3.16 R_S replaces R_T .



$1/f$ NOISE IN TUNNEL DIODES

FIGURE 3.8

The noise figure of the amplifier circuit is given by the following equation:

$$N. F. = 1 + \frac{20 I_{DC}}{g_g} + \frac{T_1 g_1}{T_g g_g} \quad (3.19)$$

and can be minimized when g_g is made as large and g_1 as small as possible. Referring back to Equations 3.3 and 3.5, the highest gain is obtained when $g_g + g_1 \approx g_d$. Therefore, the optimum value for g_g would make it equal to g_d , with g_1 by necessity becoming very small. Thus to summarize the noise considerations:

1) The minimum attainable circuit noise figure is close to the noise figure of the device as given by equation 3.18, which can be about 3 db.

2) To obtain a diode with such a low noise figure, the ratios of R_S/R_D and $g_{eq}/g_d \approx 20 I_{DC}/g_d$ must be at a minimum and the operating frequency above the $1/f$ range and not too close to f_{ro} ($< .707 f_{ro}$).

3) The circuit must then be optimized for low noise by matching the generator conductance to the negative tunnel diode conductance while the load conductance must be very small. (Example: A generator (r_g) of 100Ω could be driving a (R_D) of -100Ω looking into a load impedance (r_l) of 1000Ω . Assume $C = 5 \times 10^{-12}$. Under the above condition, the gain bandwidth product is not optimized $g_d/2C = 318 \times 10^6$, but could still be adequate $\sqrt{P. G. \cdot B. W.} = \sqrt{g_g g_1 / \pi C} = \sqrt{10^{-2} \times 10^{-3}} / 3.14 \times 5 \times 10^{-12} = 201 \times 10^6$.

Practical Design Considerations

Several other factors must be taken into consideration before actually designing an amplifier circuit. These are:

- 1) The non-linearity of $-g_d$
- 2) The bias dependance of $-g_d$
- 3) The temperature dependance of $-g_d$

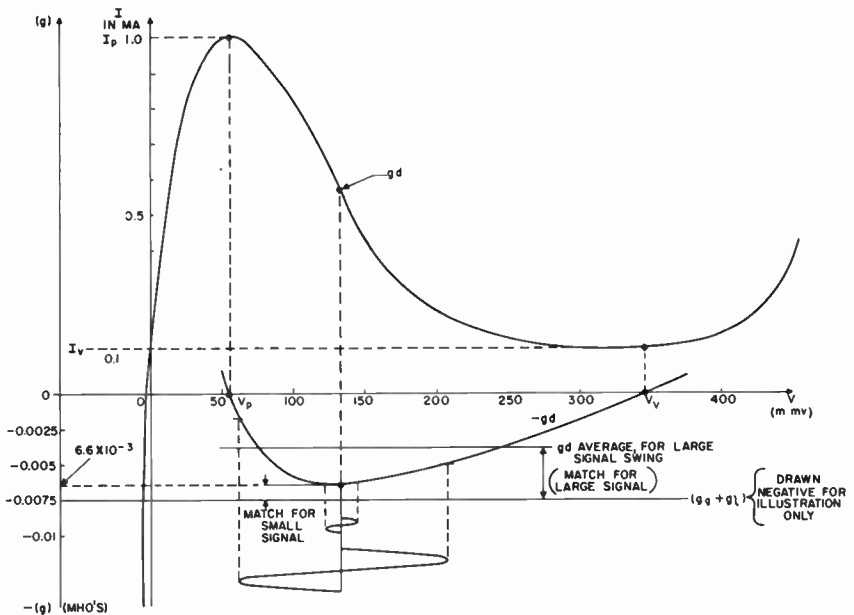
1) Non-Linearity Of The Diode Negative Conductance-

As a result of the non-linearity of $-g_d$ (see g_d vs. V in Figure 3.9), the tunnel diode amplifier is endowed with two additional problems.

- a) this non-linearity will result in some distortion.
- b) another result is a built-in "automatic gain control" effect.

The first problem is evident and fortunately the degree of non-linearity can be kept relatively low. The second problem can best be illustrated by the graph in Figure 3.9.

This graph shows the variation of I and g_d versus voltage. Looking at g_d we see it to be negative between V_p and V_v . Remembering that in an amplifier the resonant gain is obtained through the close matching of the sum of the positive circuit conductances and the negative diode conductance, one might draw these positive conductances as



**CONDUCTANCE VERSUS VOLTAGE —
CHARACTERISTIC**
FIGURE 3.9

a straight line and for illustration purposes only, forget the sign of these positive conductances and draw them as a negative conductance. The distance between line A - B and the negative conductance is the degree of mismatch. Now applying a small signal (say 2mv-rms), one can see that the degree of mismatch is relatively constant over this small voltage range and a certain gain results.

If a larger signal is applied, however, the voltage swing is larger and due to the greater non-linearity of g_d over this range, only an "average" match occurs. Or, one might say the negative conductance is no more equal to $-g_d$ at point OP but is equal to some lesser, average value of $-g_d$. This, of course, increases the distance between the average $-g_d$ line and the $(g_g + g_1)$ line thusly decreasing the gain for larger signal swings.

2) Bias Dependence of the Negative Conductance-

Biasing at the center of the more linear portion of the tunnel diode characteristic (near the inflection point -- point of maximum negative conductance) will allow the greatest voltage swing, hence the greatest dynamic range. For germanium tunnel diodes, this point is around 130 - 140 mv and the more linear portion of the characteristic will be between 80 and 180 mv, thus allowing a maximum signal swing of 100 mv peak to peak.

For high temperature operation, the large signal distortion will increase mainly as a result of the increase in valley current. If the degree of distortion is not acceptable, smaller signal swings will alleviate this problem. The greatest bias problem is that the negative conductance region is not linear. Slight variations in the bias operating point with the resulting change in negative conductance, can cause large changes in gain. It is thus essential to ensure a very stable bias voltage. Some of the methods to obtain a stable, low impedance bias supply are:

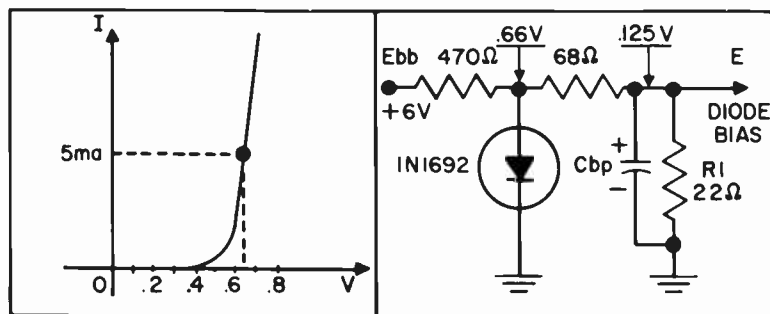
- 1) The use of mercury cells.
- 2) The use of forward biased diodes as voltage regulators.
- 3) The use of breakdown diodes as voltage regulators.

An example of the use of a forward biased diode for bias stabilization is shown in Figure 3.10. Here an inexpensive silicon diode is heavily biased in the forward direction so that it exhibits a low voltage and a low dynamic resistance. A low impedance voltage divider is used to reduce the diode voltage to the value desired for biasing of the tunnel diode. Large variations of supply voltages will cause only minor variations in tunnel diode supply voltage. At first, an apparent shortcoming of this circuit seems to be the fact that this voltage regulator is somewhat temperature sensitive and hence the operating point will shift with temperature. This might be desirable however in view of the next problem.

3) Temperature Dependence of $-g_d$

Early work on germanium tunnel diodes indicated that the negative conductance varies roughly linearly with temperature at a rate of $-0.5\%/^{\circ}\text{C}$. Such a shift would again cause gain changes in an amplifier circuit. Several methods could be used to compensate for this effect:

- a) The use of temperature sensitive resistors in series or parallel with one of the circuit legs to keep the same degree of match.
- b) The use of diodes in the bias supply (as per Figure 3.10) to produce small changes in bias resulting in a changing $-g_d$.



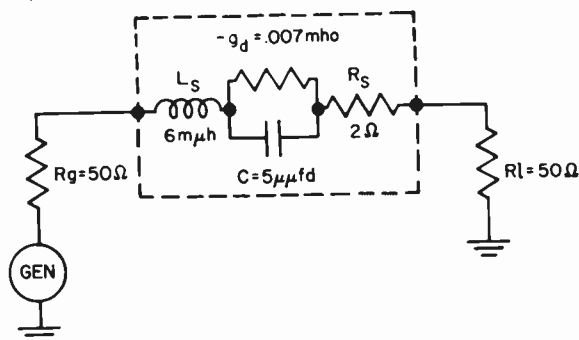
SILICON DIODE VOLTAGE REGULATOR

FIGURE 3.10

The voltage across the 1N1692 decreases at about $2.5 \text{ mv}/^{\circ}\text{C}$ as the temperature is increased. In the tunnel diode, $-g_d$ decreases at $0.5\%/^{\circ}\text{C}$ as the temperature is increased. Subsequently, if the operating point is chosen slightly beyond the inflection point (say 150 mv or so) as $-g_d$ decreases with increasing temperature, the bias decreases and the operating point moves on the characteristic towards an increased $-g_d$, thereby compensating for the original change. Furthermore, if the $2.5 \text{ mv}/^{\circ}\text{C}$ is too fast a change in bias (overcompensation) then a tunnel diode can be substituted for the 1N1692 which will result in a change of only $1 \text{ mv}/^{\circ}\text{C}$ (see Figure 2. 4).

Amplifier Design Procedure

In this circuit (see Figure 3.11), the source is a 50 ohm generator, the load is also 50 ohms while the series resistance (R_S) of the device is 2 ohms . Hence $R_T = 50 + 50 + 2 = 102 \text{ ohms}$. Use is made of a 1N2939 having a $5 \mu\mu\text{fd}$ capacitance and a negative conductance of 7 millimhos ($-rd = 143 \text{ ohms}$) at the inflection point.



A.C. SERIES LOOP CIRCUIT

FIGURE 3.11

In order to abide by the previously mentioned stability criteria, the real part of the negative conductance must be made equal to zero at the operating frequency. This also means that the circuit cut-off frequency is made equal to the operating frequency. Hence,

$$R_T - \frac{|-g_d|}{|-g_d|^2 + \omega^2 C^2} = 0, \text{ thus } R_T = \frac{1}{|-g_d| \left(\frac{1 + \omega^2 C^2}{g_d^2} \right)} \quad (3.20)$$

R_T must be therefore made equal to:

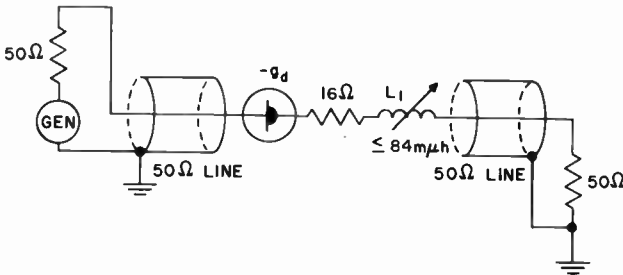
$$R_T = \frac{143}{1.21} \approx 118 \Omega$$

Since the present series loop only exhibits a $R_T \approx 102 \Omega$, a 16Ω series resistance must be added to meet the previously outlined gain and stability criteria.

The last part of the AC circuit design procedure is the choice of the tuning inductance L_1 . To get the highest value of stable gain L_T total must be only slightly smaller than the oscillation criteria $L_T < R_T C / |-g_d|$ which here must be:

$$L_T < \frac{118 \times 5 \times 10^{-2}}{7 \times 10^{-3}} = 84.3 \text{ m}\mu\text{h}$$

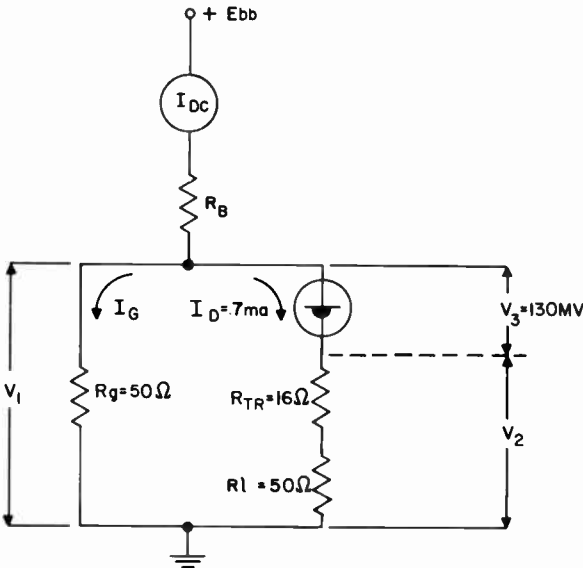
Since 2-12 $\text{m}\mu\text{h}$ are inherent in the leads of the device (depending on lead length) and some stray circuit inductance will be found in the circuit, the actual coil (L_1) will have to present a slightly smaller inductance value.



AC CIRCUIT OF 100 Mc/s AMPLIFIER STAGE

FIGURE 3.12

The bias arrangement can be derived in the following manner:



DC BIAS CIRCUIT FOR 100 Mc/s AMPLIFIER STAGE

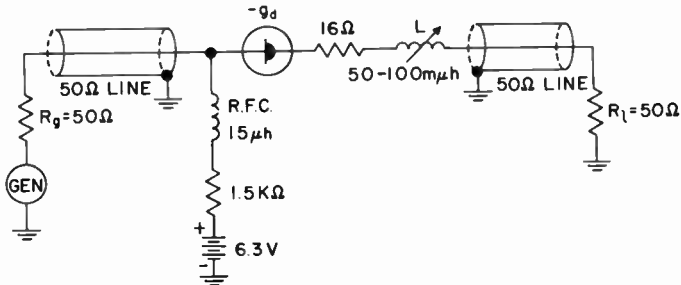
FIGURE 3.13

Assuming that the inflection point occurs at 130 mv and .7 ma, then $V_2 = 130$ mv and I_D is .7 ma and V_2 is $(R_{TR} + R_L) I_D = (16 + 50) \cdot 7 \times 10^{-3} = 44$ mv; therefore, $V_1 = 130 + 44 = 174$ mv.

I_G therefore is $174 \times 10^{-3} / 50 = 3.48$ ma, and the total DC current $I_{DC} = I_G + I_D = 3.48 + .7 = 4.18$ ma. If one were to use a 6.3 v battery then $R_B = 6.3 - .174 / 4.18 \times 10^{-3} = 6.126 / 4.18 \times 10^{-3} \approx 1.5K\Omega$. In order to decouple the DC supply from the amplifier by at least a 10 K Ω inductive reactance,

$$L_{RF \text{ choke}} > \frac{X_L}{\omega} = \frac{10^4}{6 \times 10^8} = 15 \mu\text{h}$$

Figure 3.14 shows the complete circuit.



COMPLETE 100 Mc/s "SERIES" AMPLIFIER CIRCUIT

FIGURE 3.14

The measured results were 32 db gain at 100 Mc/s with a 20 Mc/s symmetrical bandwidth. As L_1 is increased toward $L_1 = R_T C / |-g_d|$, the gain increases at the expense of bandwidth magnitude and symmetry. Further refinements in bias and temperature stabilization would complete such an amplifier design.

Conclusions

Tunnel diode "Series" and "Parallel" type amplifiers can be stable, can have low noise, high gain, wide bandwidth providing all stability criterias are satisfied. Some of the static requirements are that the device be properly biased in the negative conductance region from a stiff voltage source, that the sum of the positive conductances just barely exceed the negative conductance (or if the signal swing is greater than a few millivolts -- the average negative conductance). Furthermore, f_{x0} must be made to exceed f_{r0} and be greater than zero. This permits the circuit to be "cut off" at the operating frequency while the imaginary component is still greater than zero, hence the circuit cannot oscillate. Although the circuit is "cut off" (stable) the device in the circuit pumps energy into the load, thus gain can be obtained.

In operation, these conditions must remain satisfied, hence the parameters must remain stable under changes of temperature and

supply voltages (or be stabilized for such environmental changes).

Cascading tunnel diode amplifier stages is made difficult by the bilateral properties of this two terminal device. The amplifier is basically a positive feedback type circuit and unilateralization is impossible. Isolation between stages can be obtained by the use of isolators and circulators.

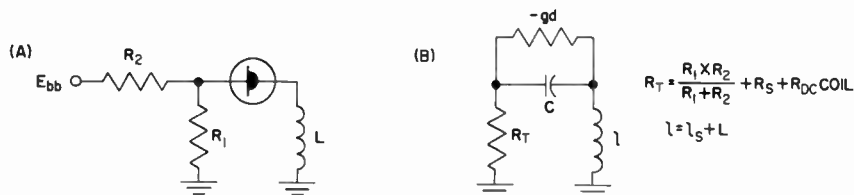
In a 1960 WESCON Paper, Dr. Schaffner⁶ described a 450 Mc/s UHF amplifier circuit having 15 db gain, 12 Mc/s bandwidth and 5.5 db noise figure (he quoted further improvements in N. F. down to 4.2 db). The antenna VSWR change was only 1 - 1.12 db, while with an isolator the circuit could withstand a change in 1 - 1.8 db VSWR (with 3 db gain reduction). Several papers in the June and July 1960 issues of the IRE Proceedings dealt with the same general problem. In the July issue, John J. Sie¹³ uses two ZJ-56 (1N2939) in conjunction with a SAGE 750 quarter wave stripline coupled Hybrid to obtain stable gain over a very wide frequency range (210 - 628 Mc/s). His results were a gain of 8.2 db \pm .6 db with a noise figure of approximately 2 db.

OSCILLATORS

4.1 Tunnel Diode Oscillators

Because of its inherent negative conductance characteristic, the tunnel diode is ideally suited for oscillator operation. The temperature stability of its characteristics is excellent and the frequency stability of the resulting oscillator circuit primarily depends on the associated circuit components.

The simplest tunnel diode oscillator circuit is the "series" type (see Figure 4.1). In this circuit the diode, appropriately biased in the negative conductance region, is put in series with an external inductor.



SIMPLE "SERIES" OSCILLATOR AND EQUIVALENT CIRCUIT

FIGURE 4.1

In this arrangement, the analysis¹⁰ shows that the circuit will oscillate freely in a sinusoidal manner when:

$$\frac{R_T}{|g_d|} = \frac{L}{C} \quad (4.1)$$

at a frequency:

$$f_o = \frac{1}{2\pi} \sqrt{\frac{1 - R_T \cdot |g_d|}{L \cdot C}} \quad (4.2)$$

Both the frequency and the stability of this circuit heavily depend on $|g_d|$ and since this conductance is not a constant, but is a time-average value which also varies with voltage and temperature, such a circuit is rather unstable.

Another circuit approach, much less subject to the various causes of instability, can be seen in Figure 4.2. The operating frequency of this "Series-Parallel" circuit is primarily dependent on the L & C of the tank circuit, and can therefore be quite stable. A de-

tailed analysis for this circuit shows that the operating frequency is determined by:

$$f_o = \frac{1}{2\pi} \sqrt{\frac{1}{L(C+C_1)} - \frac{g_d^2}{C_1(C+C_1)}} \tag{4.3}$$

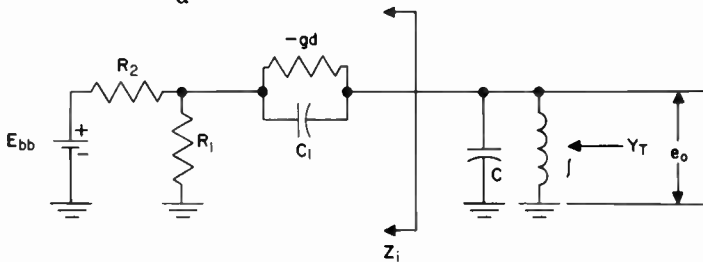
and that for stable sinusoidal oscillations,

$$R_T = \frac{g_d}{\omega^2 C_1^2 + g_d^2} \tag{4.4}$$

It should be pointed out that several approximations have been made in the analysis. First, the diode series resistance and series inductance have been neglected as they will generally be small compared to the external lumped constants. Secondly, the term g_d , which in reality should be a time-average value where g_d can be expressed in function of time as:

$$g_d(t) = g_o + g_1 \cos \omega t + g_2 \cos 2\omega t + \dots$$

is only approximated. Presently, only an "average" value of g_d is given by the manufacturer as it is the only parameter that is currently being specified. While observing the V-I characteristic, the tunnel diode is shunted by a variable resistance. When $R = 1/g_d$, the slope of the negative conductance becomes zero over the "active" range and since R is known, g_d is thus established.



SIMPLIFIED EQUIVALENT CIRCUIT OF "SERIES-PARALLEL" SINEWAVE OSCILLATOR

FIGURE 4.2

Limitations

The use of the tunnel diode as an active device in oscillator circuits involves certain problems. Obviously the first of these is the small amount of power output obtainable from such a circuit. It is apparent from a rough approximation of the V-I characteristics that:

$$P_{o \text{ max.}} = \frac{\#V(I_p - I_v)}{2\sqrt{2}} \approx \frac{I_p}{28} \tag{4.5}$$

hence it would take 28 amp- I_p to obtain one watt of power output or

#V is a coarse approximation of the "relatively" linear portion of the negative conductance slope - 100mv for Ge & 200mw for GaAs units.

2.8 amps to even get 100 mw. With a 2.8 amp unit, the negative resistance is only around 0.1Ω or less. Driving such a resistance from a stiff voltage source is difficult and inefficient.

Another problem is that such a high power unit has a rather large junction area, hence a relatively large junction capacitance. The magnitude of this capacitance seriously limits the frequency at which this power is available.

Further limitations can be found in coupling this oscillator to a load. The tunnel diode oscillator circuit only delivers a fraction of a volt (rms) to the tank circuit. Generally, as this circuit is coupled into other low impedance semiconductor circuits, a stepdown transformer only serves to lower this output swing.

Finally, the efficiency of such a circuit, driven from present conventional power supplies (relatively high DC voltages) is quite low. An efficient tunnel diode supply would provide a fraction of a volt with a large current capacity -- just the opposite of conventionally used supplies. On the other hand, solar cells and thermoelectric generators could probably be used to advantage in tunnel diode circuits.

Oscillator Design Procedure

Given the frequency and required output power, the choice of diode can be made. For one milliwatt of rms power, a 28 ma peak current unit is required, since $I_p = 28 P_o$. One consideration in the choice of diode is its capacity and inductance for microwave applications.

Another consideration, is of a purely practical nature. At low frequencies, the size of the inductor becomes electrically and physically large. To minimize physical dimensions, smaller wire is used resulting in larger values of DC resistance. The tunnel diode must remain stably biased in the negative conductance region. Therefore, R_T must be smaller than $1/g_d$, and if the DC resistance of the coil were very large, g_d would have to be very small. This means the use of a very low current tunnel diode since $g_d = \Delta_i/\Delta_v$ and Δ_v is fixed.

For our design example, let us use a 1 Kc/s sinusoidal oscillator having a limited power output of about one microwatt. A rough approximation tells us that a unit with I_p over $28 \mu a$ will do the job. As tunnel diodes with $I_p < 1$ ma are not generally available, a backward diode (generally this device can be used as a low current tunnel diode) is utilized. Such a unit is the ZJ69 having a negative conductance of approximately 3×10^{-4} mhos. R_T can now be 1 - 2 K Ω without seriously impairing the DC stability of the circuit.

Once g_d is determined, the choice of R_T can be made. The limits are that $R_T < 1/g_d$, but if R_T is made small, the DC power consumption of the circuit becomes too large. Hence R_T is chosen somewhere between .3 and .7 of $1/g_d$, as this will yield adequate stability with reasonable power consumption.

If we choose $R_T = .3/g_d$, R_T will be about 1000 Ω for our example. At this point, C_1 can be calculated, where: (Per Eq. 4.4)

$$C_1 = \sqrt{\frac{g_d (1 - R_T \cdot g_d)}{R_T \omega^2}} = \sqrt{\frac{3 \times 10^{-4} (1 - .3)}{10^3 \times 4 \times 10^7}} = .071 \text{ mfd}$$

Since:

$$C + \frac{C_1}{1 - R_T \cdot g_d} = \frac{1}{L \cdot \omega^2}$$

if L is chosen somewhat arbitrarily at 100 mh with a $R_{DC} = 100 \Omega$;

$$C + \frac{C_1}{1 - R_T \cdot g_d} = \frac{1}{.1 \times 4 \times 10^7} = 0.25 \text{ mfd}$$

hence:

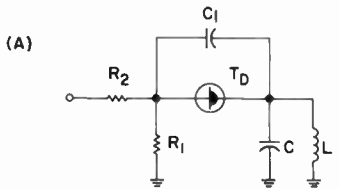
$$C = 0.25 - \frac{.071}{1 - .3} \text{ mfd} = 0.15 \text{ mfd}$$

The DC bias circuit can now be calculated. Assume the circuit supply voltage $E_{bb} = 3.0$ volts. If R_1 (see circuit in Figure 4.3) is 1000Ω , then:

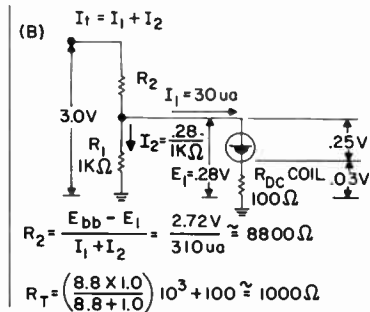
$$R_2 = \frac{E_{bb} - E_{diode} + E_{RDC \text{ Coil}}}{I_{total}} = \frac{3 - (.25 + .03)}{310 \times 10^{-6}} \approx 8800\Omega$$

and I_{total} is: $I_{diode} + E_{diode}/R_1 = 310 \mu a$

$$R_T = \frac{R_1 \times R_2}{R_1 + R_2} + R_{DC \text{ Coil}} + R_s = \frac{8800 \times 1000}{9800} + 100 + 2 \approx 1000 \Omega$$



- $R_1 = 1K\Omega$
- $R_2 = 8.8K\Omega$
- $C = .15\text{mfd}$
- $C_1 = .071\text{mfd}$
- $L = 100\text{mh} (R_{DC} = 100\Omega)$
- $T_D = ZJ69 (\text{BACK DIODE})$



$$R_2 = \frac{E_{bb} - E_1}{I_1 + I_2} = \frac{2.72V}{310\mu a} \approx 8800\Omega$$

$$R_T = \left(\frac{8.8 \times 1.0}{8.8 + 1.0}\right) 10^3 + 100 \approx 1000\Omega$$

1 Kc/s OSCILLATOR CIRCUIT

FIGURE 4.3

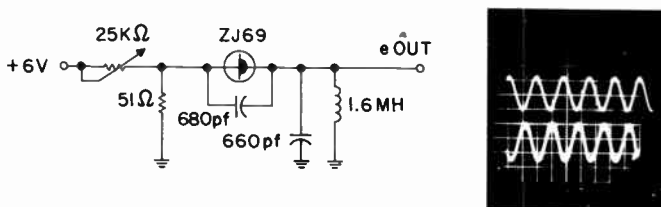
Another example at 100 Kc/s, would simply divide L, C and C_1 by 100 thus:

$$C_1 = 710 \text{ pf}$$

$$C = 1500 \text{ pf}$$

$$L = 1 \text{ mh}$$

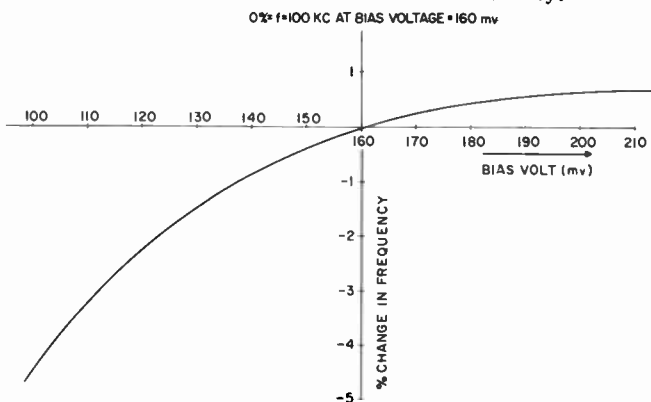
If the same tunnel diode is used, the bias circuit remains the same as in Figure 4.3. At 100 Kc/s, the size of the inductance will easily permit the use of larger peak current tunnel diodes however, and hence the circuit can be readjusted to accommodate a 1 ma I_p conventionally available unit with greater power output. The performance characteristics of a 100 Kc/s oscillator is illustrated in Figures 4.4 through 4.6. Figure 4.4 shows the circuit diagram and the oscilloscope presentation of the output waveform across the tunnel diode (upper) and across the tank circuit (lower).



**100 Kc/s SINEWAVE OSCILLATOR CIRCUIT
AND OUTPUT WAVEFORMS**

FIGURE 4.4

Figure 4.5 shows the frequency variation vs. bias voltage and Figure 4.6 shows the frequency variation vs. temperature characteristic of the diode. In these tests only the tunnel diode was heated in order to establish its contribution to circuit instability.



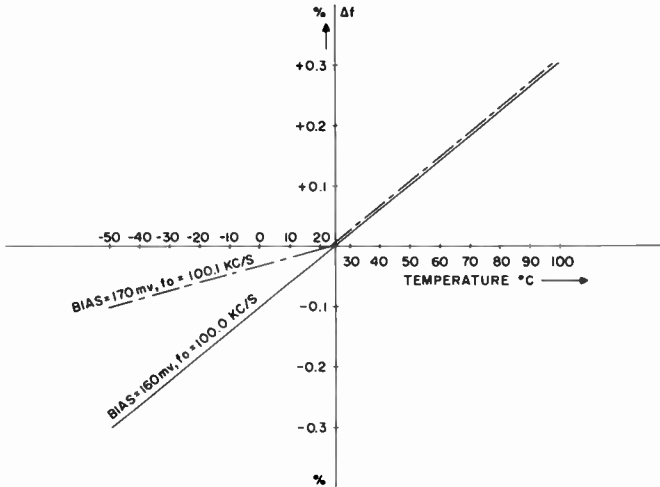
FREQUENCY VARIATION VS. BIAS VOLTAGE AT 25°C

FIGURE 4.5

As can be seen from the data, the frequency stability of this circuit is quite good over a wide range of temperature and bias voltage.

1 Mc/s Oscillator

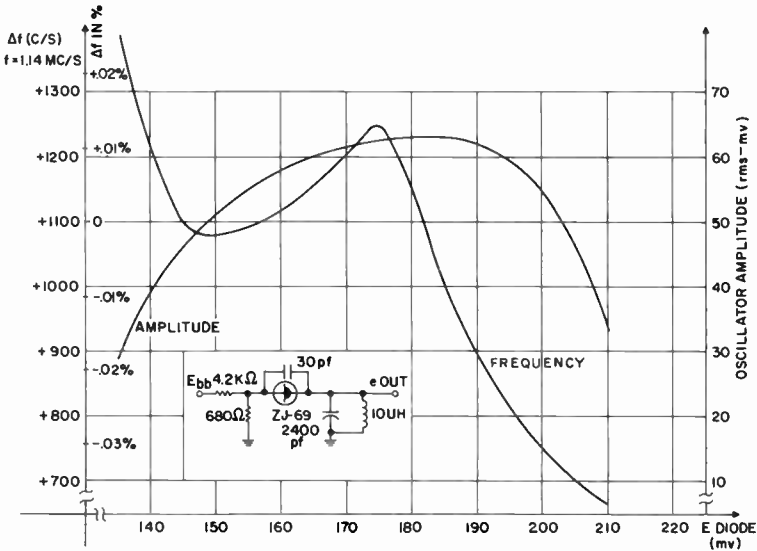
The next circuit, a 1 Mc/s oscillator, was tested and showed considerable improvement, as illustrated in Figures 4.7 through 4.9.



**FREQUENCY VARIATION VS. TEMPERATURE
OF 100 Kc/s OSCILLATOR**

FIGURE 4.6

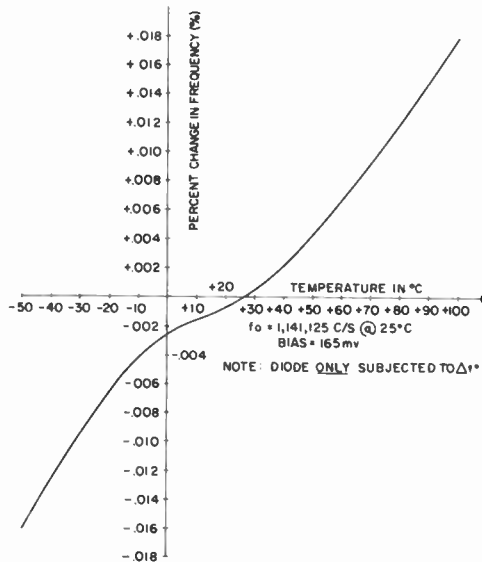
Figure 4. 7 illustrates the fact that by changing the bias voltage, the variation of frequency is only in the order of a few hundred cycles out of 1.14 Mc/s, a change in the order of 0.05%. Over the range of 155 to 200 mv, the amplitude of the output voltage varies only from 63 mw-rms down to 55 mv-rms.



**FREQUENCY STABILITY VS. BIAS VOLTAGE
OF 1 Mc/s OSCILLATOR**

FIGURE 4.7

Figure 4.8 shows the temperature stability of this oscillator circuit over a range of -50°C to $+100^{\circ}\text{C}$. From this figure it can be seen that the frequency varies only $+18 - 16$ millipercents over this wide temperature range.



**FREQUENCY STABILITY VS. TEMPERATURE
OF 1 Mc/s OSCILLATOR**
FIGURE 4.8

Frequency Stability

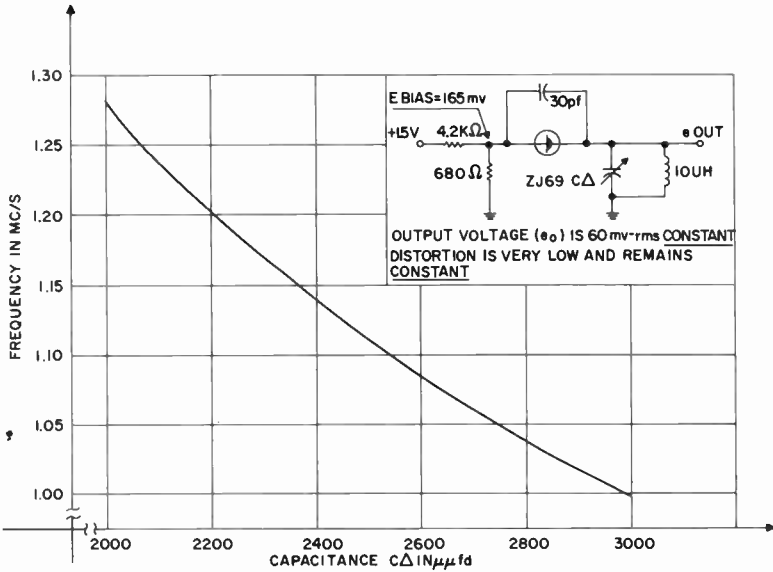
The main variable creating the frequency shift versus bias and temperature is the negative conductance ($-g_d$). As the operating frequency is increased, the variation of frequency (Δf) caused by Δg_d remains the same, but since f changes, $\Delta f/f \times 100$ decreases, thus causing a lower percentage of frequency shift.

Finally, Figure 4.9 shows that varying the capacitor in the parallel tank circuit over an appreciable range accomplishes smooth tuning without change in amplitude or distortion. This latter point is especially important since in the "series" type oscillator circuit a change in l or C only, affects stability and distortion greatly.

Frequency-Power Product

At higher power output requirements (in the milliwatt region) further limitations cause the aforementioned design and the design procedure to become impractical. After having chosen a tunnel diode with a sufficient peak current to furnish the required power output, g_d is essentially established. For example, to get 3 mw of power output (rms), it will take a 100 ma unit to furnish this amount of power. A

100 ma unit has a negative conductance g_d in the order of 1 mho ($-rd=1\Omega$). If one keeps the product of $R_T \cdot g_d$ at 0.5 then C_1 is established by Equation 4.4.



FREQUENCY TUNING BY MEANS OF C_Δ

FIGURE 4.9

The problem arises in the heretofore somewhat arbitrary choice of L . This inductance becomes much more restrictive as g_d and the operating frequency increase. The inductance is determined by:

$$L = \frac{1}{\omega^2 \left| C + \frac{C_1}{1 - R_T \cdot g_d} \right|} \tag{4.6}$$

The maximum value of L is realized when C becomes zero, at which time:

$$L_{max.} = \frac{1}{\omega^2 \left(\frac{C_1}{1 - R_T \cdot g_d} \right)} \tag{4.7}$$

If one now determines this value by using a higher power example at 1 Mc/s:

$$L_{max.} = \frac{1}{39.6 \times 10^{12} \times .32 \times 10^{-6}} = 78 \text{ m}\mu\text{h}$$

If a small capacitance is now added and/or if the operating frequency is further increased, L will quickly reach the package limitations.

For example, if $f = 10 \text{ Mc/s}$ and C is 10 pf, then:

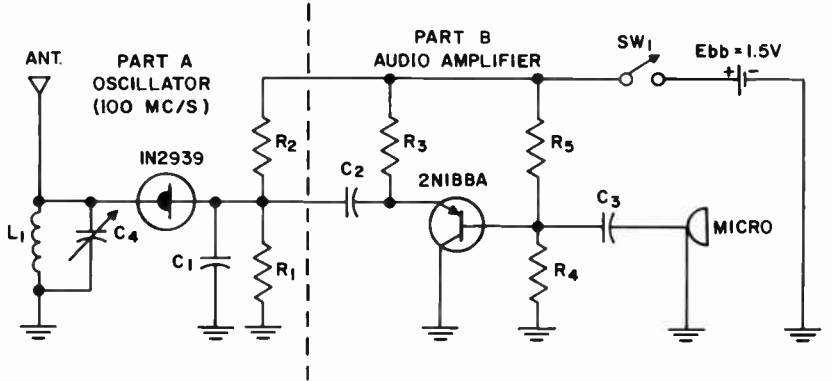
$$C_1 = \sqrt{\frac{1(1 - .5)}{.5 \times 39.6 \times 10^{14}}} = .016 \text{ mfd}$$

$$L = \frac{1}{39.6 \times 10^{14} \times 10 \times 10^{-12} + \frac{(16 \times 10^{-9})}{.5}} = 7.88 \text{ m}\mu\text{h}$$

For the TO-18 packaged devices, this is the magnitude of the case and lead inductance, and a further reduction of the inductance would be difficult to achieve. Naturally lower inductance tunnel diode configurations as well as some techniques of tuning out some of the inductance (in coaxial or triplate arrangement for microwave applications) might extend these limits somewhat.

FM Transmitter

A simple micropower FM transmitter using the 1N2939 tunnel diode is shown in Figure 4.10.



- | | | |
|-------------------------------|--|--|
| R ₁ 22Ω -1/2 WATT | C ₁ .001 CERAMIC DISC | L ₁ 6T NO.16 WIRE - 3/8" OIAM. OPEN AIR (L≈2μh) |
| R ₂ 270Ω -1/2 WATT | C ₂ 50 μfd ELECTROLYTIC | ANT. ANTENNA 4 3/4" LENGTH NO. 14 WIRE |
| R ₃ 470Ω -1/2 WATT | C ₃ 5 μfd ELECTROLYTIC | MICRO MICROPHONE "SHURE BROTHERS" MOEEL NO. 420 OR EQUIV |
| R ₄ 10KΩ -1/2 WATT | C ₄ 1.5-5.0 μμfd AIR VARIABLE | |
| R ₅ 10KΩ -1/2 WATT | | |

88-108 Mc/s WIRELESS F.M. MICROPHONE

FIGURE 4.10

Operation may be best explained by separating the circuit into two portions. Part A is a basic tunnel diode oscillator whose frequency is primarily determined by the resonant circuit in the cathode. Resistors R1 and R2 provide a stable low impedance voltage for the anode of approximately 150 mv. Capacitor C₁ is the RF bypass for the anode.

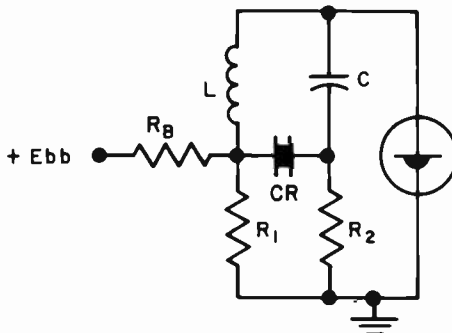
Part B is a transistor emitter follower stage to amplify the audio signal from the microphone. The amplified audio is fed through capacitor C2 to the anode of the tunnel diode. FM modulation is ac-

completed by the audio signal instantaneously changing the anode bias. Since the characteristic curve is not perfectly linear in the negative resistance region, the negative conductance changes slightly with bias. As can be seen from the self-resonant frequency equation, f_{xo} is a function of $-g_d$ and therefore the resonance of the circuit is affected. FM deviations of ± 75 KC are readily obtainable with this type of circuit.

The transmitter shown in the diagram has been successfully used as a wireless portable microphone. Its great advantage is that it allows complete mobility on the part of the speaker, and of course has no wires or cords. When used with an average FM receiver having a sensitivity of $10 \mu v$, an operating range in excess of 100 feet was obtained.

Crystal Controlled Oscillator

The circuit in Figure 4.11 works basically as per the previous description with the exception of the criteria for R_T . R_1 and R_2 are identical and are chosen to be about twice the value required for R_T . As a result, oscillations are not possible "off resonance". At resonance however, the crystal becomes a short circuit and R_1 is in parallel with R_2 , essentially halving R_T . This new value of R_T permits the circuit to oscillate freely at a frequency accurately governed by the crystal.



CRYSTAL CONTROLLED OSCILLATOR

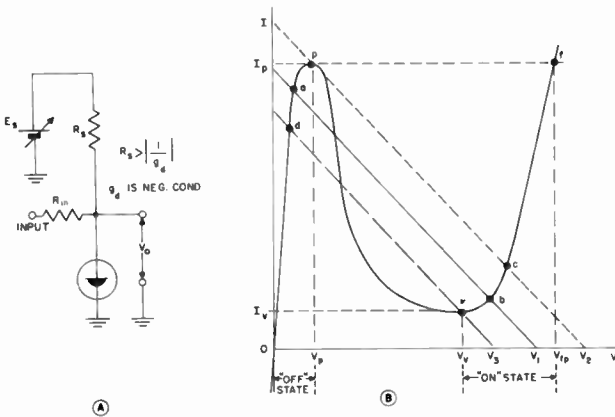
FIGURE 4.11

CHAPTER 5

SWITCHES

One of the most promising modes of operation for the tunnel diode is as a switch where it can be used to function as an astable, monostable, and bistable oscillator, or it can be used in performing logic and memory functions. In addition, it can be used with other semiconductor devices to perform a variety of functions.

As discussed in Chapter 3, the stability criteria of the series circuit dictates that the negative resistance portion of the tunnel diode characteristic is an unstable region if the total circuit positive resistance is larger than the negative resistance of the tunnel diode. Consequently, the tunnel diode can only switch through this region to either of its stable states. Figure 5.1a shows a simple tunnel diode bistable switching circuit. This requires that $R_s > |1/g_d|$ where R_s is the load resistance, and $|g_d|$ is the magnitude of the negative conductance. Figure 5.1b shows the conductance load line of R_s superimposed on the tunnel diode characteristic. For a supply voltage V_1 ,



SIMPLE TUNNEL DIODE BISTABLE CIRCUIT

FIGURE 5.1

the load line intersects the tunnel diode characteristic at points "a" and "b". If the stable point is initially at point "a" and the supply voltage is increased to V_2 , the circuit current exceeds the peak current of the tunnel diode, and it switches to point "c". If the supply voltage is reduced to V_3 , the circuit current becomes less than the valley current, and the tunnel diode will switch back to point "d". On the other hand, if the supply voltage is held constant at V_1 , the tunnel diode can be switched from point "a" to point "b" by a positive current pulse at the input terminals which momentarily increases the tunnel diode current to a value in excess of the peak current. The diode can

be switched from point "b" to point "a" by applying a negative current pulse of such magnitude at the input terminals as to momentarily reduce the tunnel diode current to a value less than the valley current.

The low voltage or "off" state of the tunnel diode is the region from zero to the peak voltage, and the current through the diode consists entirely of majority carriers transported across the junction by the tunneling mechanism. The high voltage or "on" state is considered to be the region from the valley voltage to the peak forward voltage, and the current through the diode consists entirely of minority carriers transported across the junction by diffusion.

The speed of switching between the two states is very high and is determined chiefly by the junction capacitance and the amount of charge available from the trigger pulse. If a trigger of minimum amplitude is used in conjunction with a constant current bias source (the load line in Figure 5.1b is horizontal), the rise time of the tunnel diode voltage between the 10 and 90 percent points is approximately given by:

$$t_r = \frac{(V_{fp} - V_p) C}{(I_p - I_v)} \quad (5.1)$$

where C is the tunnel diode capacitance, V_{fp} and V_p are the forward and peak voltages, and I_p and I_v are the peak and valley currents respectively. Equation 5.1 can also be written:

$$t_r = \left(\frac{V_{fp} - V_p}{I_p \frac{\eta - 1}{\eta}} \right) C \quad (5.2)$$

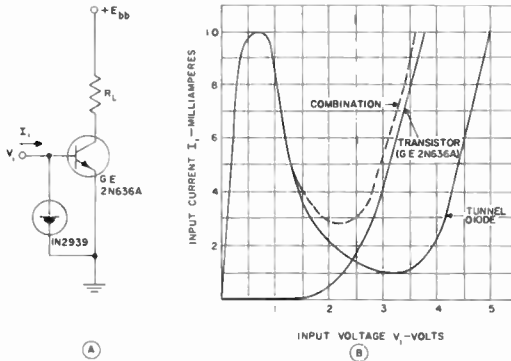
where η is the peak-to-valley current ratio. Since V_{fp} , V_p and $C/I_p - I_v$ are more or less independent of I_p , the rise time is also independent of the peak current, and the rise time can be decreased by reducing $C/I_p - I_v$. This is equivalent to reducing C/g_d . Using the parameters of the 1N2939 listed in Chapter 2, the rise time of this device is calculated to be 3.5 μsec . which is in close agreement with measured values. Switching speeds for 10 ma germanium tunnel diodes have been measured to be less than 1 μsec ., and the correlation with the calculated values has been excellent.

5.1 Hybrid Circuits

Many simple and practical switching circuits are possible when the tunnel diode is used in conjunction with a transistor or controlled rectifier.

A transistor-tunnel diode hybrid circuit can be formed by paralleling the base-emitter junction of the transistor with a tunnel diode as seen in Figure 5.2a. When the tunnel diode is switched to the high voltage state, the transistor is turned on since the V-I characteristic of the tunnel diode is similar to that of the base-emitter junction of

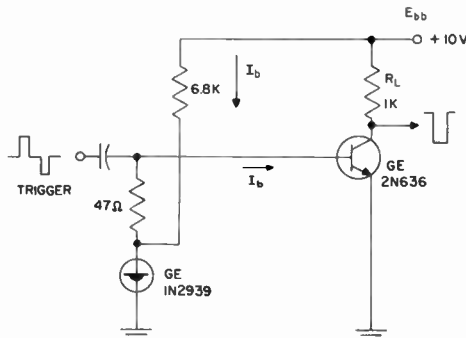
the transistor. A comparison of the tunnel diode and transistor characteristics as well as the parallel combination is given in Figure 5.2b. The net input characteristic can then be analyzed by means of load lines for bistable operation as desired.



HYBRID TRIGGER CIRCUIT AND CHARACTERISTIC
(Using Germanium Alloy Transistor and Germanium Tunnel Diode)

FIGURE 5.2

A simple bistable circuit or flip-flop is shown in Figure 5.3. In this case, the tunnel diode is biased in the low voltage state by the current I_b which is slightly less than the peak current. Since the transistor is in the "off" condition, the collector is at the supply voltage, E_{bb} . If a positive trigger pulse is supplied at the input such that the tunnel diode current increases above the peak current, the tunnel diode switches to the high voltage state. The tunnel diode will remain in the high voltage state with a major portion of the bias current being diverted into the base of the transistor (how much is diverted can be obtained from the input characteristic shown in Figure 5.2b). The transistor collector voltage will then fall to ground potential if $I_b h_{FE} > E_{bb}/R_L$.

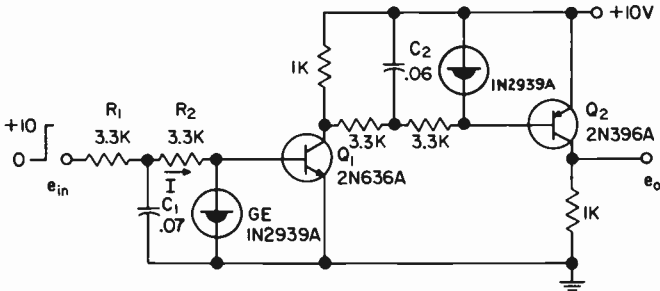


**BISTABLE CIRCUIT USING TUNNEL DIODE
AND NPN GERMANIUM ALLOY TRANSISTOR**

FIGURE 5.3

Likewise, a negative pulse at the input, which causes the tunnel diode current to fall below the valley current, will switch the tunnel diode back to the low voltage state, thus turning the transistor off. The 47Ω resistor which is in series with the tunnel diode in Figure 5.3 serves to insure that the tunnel diode will be biased above the valley point when in the high voltage state, and it also prevents the loading of the trigger pulses by the tunnel diode.

A simple hybrid time delay circuit which permits any number of consecutive time delays is shown in Figure 5.4. In this case, the bias supply for the tunnel diode has been eliminated, and the input is supplied through a charging circuit. In Figure 5.4, the timing cycle starts with the application of a positive 10 volt step at the input. C_1 charges through R_1 , with the current I being proportional to the voltage across C_1 . The time constant of the charging circuit is approximately $(R_1 R_2 / R_1 + R_2) C_1$. When the current I reaches the peak current value of the tunnel diode, the tunnel diode switches to its high voltage state and turns Q_1 on. The time delay for a step change at the



TUNNEL DIODE TIME DELAY CIRCUIT WITH TWO CASCADED COMPLEMENTARY STAGES

FIGURE 5.4

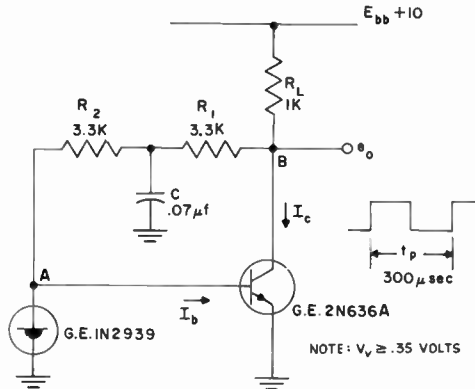
input of e_{in} is given by:

$$t_d \approx R'C_1 \ln \left[1 - \frac{(R_1 + R_2) I_p}{e_{in}} \right] \quad (5.3)$$

where $R' = R_1 R_2 / R_1 + R_2$. Because the collector voltage of Q_1 falls from +10 volts to a very low value, a similar timing sequence is initiated for the second stage since the second stage is a complementary version of the first stage. At the end of the second timing sequence Q_2 will turn on, and its collector voltage will rise to 10 volts. For the component values shown in Figure 5.4, the time delay of each stage is approximately 120 microseconds. A multiphase oscillator results if an odd number of stages are connected in a closed loop.

Using the time delay scheme described above, a square wave oscillator can be built by returning the input to the collector of the transistor as shown in Figure 5.5. A cycle of operation begins with C charging through R_1 and R_L . The transistor is turned off because the tunnel diode is in the low voltage state. When the current through

R_2 reaches the peak current value of the tunnel diode, the tunnel diode switches to the high voltage stage. The base current I_b then drives the transistor to saturation, shorting the collector and point B to ground. The capacitor then discharges through R_1 and R_2 until the voltage at point A falls below the valley voltage of the tunnel diode, and the tunnel diode then returns to the low voltage state and turns the transistor off. The cycle then repeats itself. The symmetry of the output can be changed if a small current is fed in or taken out of point A. In order that the circuit operate properly, several require-



ASTABLE HYBRID OSCILLATOR

FIGURE 5.5

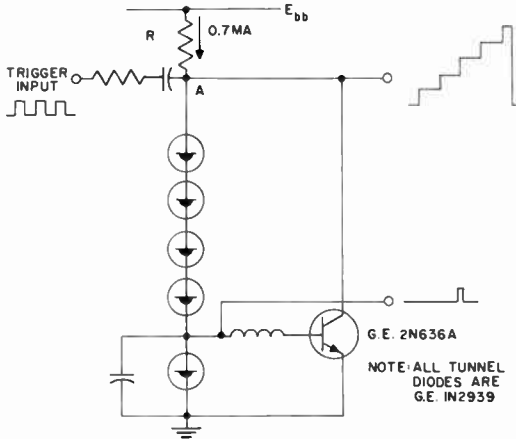
ments must be met:

1. When the transistor is in the off state (the tunnel diode is in the low voltage state), the current through R_2 must exceed the peak current of the tunnel diode before the capacitor finishes charging to its steady-state value.
2. At the valley voltage of the tunnel diode, the product $h_{FE} I_b$ must be larger than I_c or E_{bb}/R_L . This is the same as saying the transistor must remain saturated when the base emitter voltage is equal to the valley voltage of the tunnel diodes.
3. The switching times of the transistor must be less than the time required for one cycle.

With the values of components shown in Figure 5.5, the period is about 300 microseconds.

A simple pulse frequency divider of N:1 can be made using the technique shown in Figure 5.6. In this case, five tunnel diodes are connected in series to form a 5:1 frequency divider. They are biased from a current source whose magnitude is less than the peak currents of the tunnel diodes; thus, they are all in the low voltage state. The bottom diode is selected to have a higher peak current than the other tunnel diodes in the circuit. Each time a positive pulse occurs at the input, one of the upper four tunnel diodes switches to the high voltage state. When the fifth pulse occurs, the bottom tunnel diodes switches

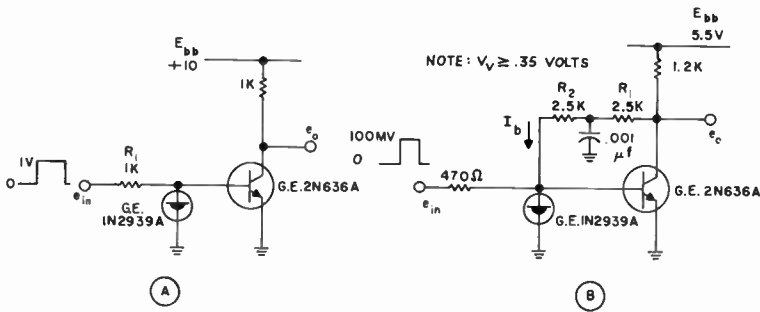
to the high voltage state and turns on the NPN transistor, which resets the circuit by essentially grounding point A. When point A is grounded, the current through the tunnel diodes is reduced to a value less than the valley current, and the tunnel diodes switch back to the low voltage state. The capacitor across the bottom tunnel diode and the inductor in series with the transistor base delay the signal to the transistor so that complete switching can occur. The waveform appearing across the string of tunnel diodes is a staircase with the risetime of each step being determined by the rise time of the trigger pulse. The operating frequency of the circuit is limited chiefly by the switching speed of the reset transistor.



**SERIES CONNECTED TUNNEL DIODES USED FOR 5:1 PULSE
FREQUENCY DIVIDER OR STAIRCASE WAVE GENERATOR**

FIGURE 5.6

The tunnel diode also makes an excellent level detector. Figure 5.7a shows a simple tunnel diode-transistor level detector which has a very stable switching point, with the stability of the switching point being determined primarily by the stability of the peak current. Since the drift of the peak current with temperature for germanium tunnel diodes is a function of peak voltage, it can be seen from Figure 2.3 that tunnel diodes with peak voltages of 60 mv have the least drift with temperature. If these tunnel diodes are used in Figure 5.7, the drift of the level of switching can be less than ± 50 mv from -50 to 100°C . If the level of the voltage to be detected is less than a volt, R_1 in Figure 5.7a may be decreased; however, the minimum value it can have is 200 ohms. The circuit in Figure 5.7b allows the detection of voltages down to about 70 mv without the use of a negative supply. I_b is adjusted to be 90 to 95% of the peak current, and the time constant $R_1 R_2 C / R_1 + R_2$ must be made larger than the width of the pulse for the circuit to operate properly. With the values of components shown, an output is obtained when the input pulse reaches 100 mv, and the circuit will function properly if the pulse widths are between 0.5 and 5 microseconds. A 5.5 volt supply voltage is used since reference diodes

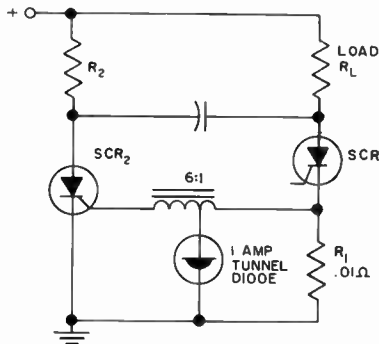


HYBRID LEVEL DETECTOR CIRCUITS

FIGURE 5.7

with this voltage have nearly a zero temperature coefficient. For larger pulse widths, C must be increased. If the circuit is to operate at high temperatures (71°C), a drift in the level of switching will result with the circuit as shown because of the transistor leakage current. For this case, the reference diode should be tied from the collector to ground and the supply voltage raised so that the reference diode is operating well in the breakdown region.

Figure 5.8 shows a tunnel diode-silicon controlled rectifier current limiting circuit for power equipment. In this case, the tunnel diode has a peak current of one ampere. When the load current exceeds the limit value, the voltage across the .01 ohm sensing resistor exceeds the peak voltage of the tunnel diode, and it switches to the high voltage state. The change of voltage across the tunnel diode is stepped up by the auto transformer to a value which is sufficient to fire the controlled rectifier, SCR2. When SCR2 fires, a negative voltage is applied by C_1 across SCR1 which causes SCR1 to turn off in $20 \mu\text{sec.}$ or less, thus interrupting the load current. The advantage of using the tunnel diode in this application is its ability to be triggered at a very low voltage level with the resultant very low power loss in the current sensing resistor.

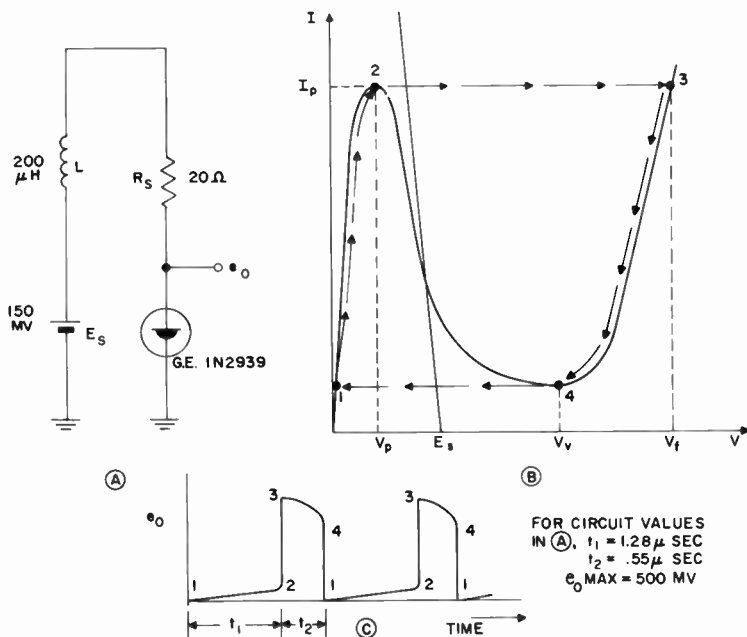


TUNNEL DIODE USED AS CURRENT SENSING ELEMENT IN SILICON CONTROLLED RECTIFIER CIRCUIT BREAKER

FIGURE 5.8

5.2 Astable Oscillators

Figure 5.9a shows a simple tunnel diode relaxation oscillator circuit. For proper operation of the circuit, the DC load line must intersect the tunnel diode characteristic in the negative resistance region as shown in Figure 5.9b. Consequently, the magnitude of E_S must be less than V_V and greater than V_P .



BASIC TUNNEL DIODE RELAXATION OSCILLATOR CIRCUIT

FIGURE 5.9

The circuit operation begins when power is applied. The circuit current increases exponentially, and when it reaches the peak current value of the tunnel diode (point 2 in Figure 5.9b), the tunnel diode switches to the high voltage state (point 3). Since the voltage drop across the tunnel diode is larger than the supply voltage, the circuit current begins to decrease exponentially. However, when the circuit current reaches the valley current of the tunnel diode (point 4), the tunnel diode switches to the low voltage state (point 1). One cycle of operation is completed, and another cycle just like the one just described begins.

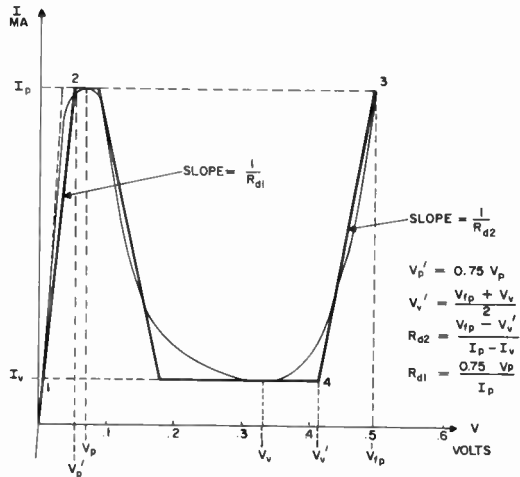
Because the current build up from point 1 to point 2 requires a longer time than when it is decreasing from point 3 to point 4, the output voltage across the tunnel diode will not be symmetrical. This is shown in Figure 5.9c. The trailing edge of the output is quite rounded--this is due to the non-linear V-I characteristic of the tunnel diode in the high voltage region.

In order to avoid a non-linear differential equation in solving for

the period of the relaxation oscillator, the V-I characteristic must be linearized since current-voltage relationships of the diode in both the low and high voltage regions are non-linear. This linearization is shown in Figure 5.10. If the linear portion of the diode characteristic in the low voltage region is extended, it will intersect the I_p constant current line at approximately $V_p/2$. Therefore, an "average" value of voltage at this current is $0.75 V_p$, and the linearized dynamic resistance is then:

$$R_{d1} = \frac{0.75 V_p}{I_p} \quad (5.4)$$

**LINEARIZATION OF THE
TUNNEL DIODE
CHARACTERISTIC**
FIGURE 5.10



In the high voltage state, between V_v and V_{fp} , there is no linear region. Therefore, an "average" value of voltage at the valley current is:

$$V_v' = \frac{V_{fp} + V_v}{2} \quad (5.5)$$

and the linearized dynamic resistance in this region is:

$$R_{d2} = \frac{V_{fp} - V_v'}{I_p - I_v} \quad (5.6)$$

The time required by the circuit current to go from point 1 to point 2 then is given by:

$$t_1 = L/R_{T1} \ln \left(\frac{E_s - R_{T1} I_v}{E_s - R_{T1} I_p} \right) \quad (5.7)$$

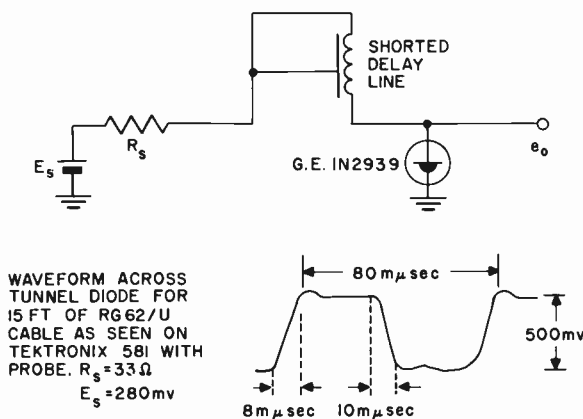
where $R_{T1} = R_s + R_{d1}$, R_s is the external series resistance, and E_s is the supply voltage.

The time required by the circuit current to decrease from point 3 to point 4 is given by:

$$t_2 = L/R_{T2} \ln \left(\frac{V_{fp} + I_p R_s - E_s}{V_v + I_v R_s - E_s} \right) \quad (5.8)$$

where $R_{T2} = R_s + R_{d2}$. The total period is then $t_1 + t_2$. Correlation between the calculated values using equations (5.7) and (5.8) and actual measurement was within 10% on 0.5, 1 and 2.2 ma germanium tunnel diodes with wide ranges of supply voltage and series resistance.

The period of the simple relaxation oscillator is very sensitive to supply voltage, with the sensitivity being primarily due to t_1 . If a shorted delay line is used in place of the inductance, a square wave results, whose frequency is insensitive to supply voltage changes.¹⁷ Such a circuit is shown in Figure 5.11.



**RELAXATION OSCILLATOR USING
A SHORTED DELAY LINE**

FIGURE 5.11

5.3 Monostable Oscillator

If the supply voltage is decreased or increased such that the DC load line intersects the tunnel diode characteristic in a positive resistance region as shown in Figure 5.12b, a monostable oscillator results. A supply voltage of E_{s1} will bias the tunnel diode in the low voltage state. A positive input current pulse of magnitude $I_p - I_a$ will switch the tunnel diode to the high voltage state (point 3 in Figure 5.12b). The current in the inductance begins to decrease exponentially from I_a until it reaches the valley current at which time the tunnel diode switches back to the low voltage state. The switching time from point 2 to point 3 (rise time) and from point 4 to point 1 (fall time) are less than ten nanoseconds for one milliamperere germanium tunnel diodes. The time required for the current to decrease from point 3 to

point 4 (the output pulse duration) is given by:

$$t_a = \frac{L}{R_T} \ln \left[\frac{V_{fa} + I_a R_s - E_{s1}}{V_v' + I_v R_s - E_{s1}} \right] \quad (5.9)$$

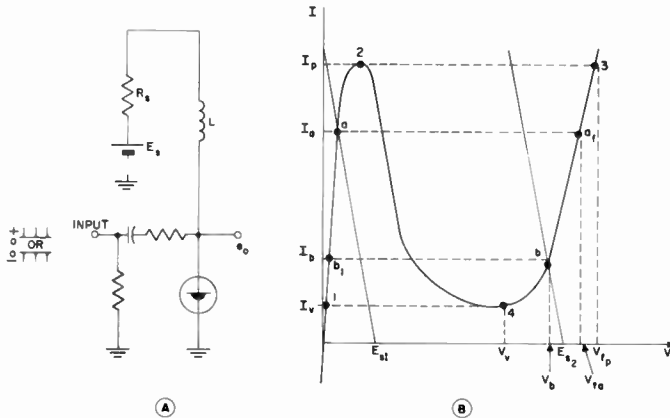
where $R_T = R_{d2} + R_s$ and $V_v' = (V_{fp} + V_v)/2$. R_{d2} is derived as discussed in the above section on astable oscillators. The value of I_a can be determined by measurement, graphical analysis, or it can be derived as:

$$I_a \approx \frac{E_{s1}}{R_s + R_{d1}} \quad (5.10)$$

where R_{d1} is derived as discussed in the above section on astable oscillators. V_{fa} is given by:

$$V_{fa} = V_{fp} - R_{d2} (I_p - I_a) \quad (5.11)$$

Correlation between measured and calculated values using equation 5.9 were within 10%. The maximum repetition rate of the trigger pulses is limited by the length of time required for the circuit current to increase from point 1 to point "a" in Figure 5.12b. This time is given by equation 5.7 with the exception that I_p is replaced by I_a .



TUNNEL DIODE MONOSTABLE OSCILLATOR

FIGURE 5.12

If the tunnel diode is biased in the high voltage state (point b in Figure 5.12b), then negative trigger pulses of a magnitude $I_b - I_a$ are required at the input. The output pulse will then be one going from a high voltage to a low voltage, and its width is given by:

$$t_b = \frac{L}{R_T} \ln \left[\frac{R_T I_b - E_{s2}}{R_T I_p - E_{s2}} \right] \quad (5.12)$$

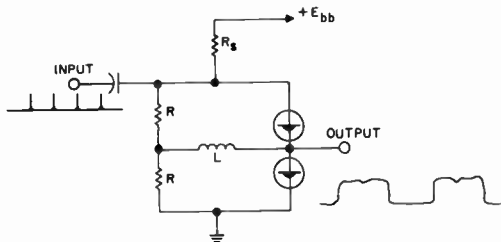
where $R_T = R_s + R_{d1}$. The value of I_b is given by:

$$I_b \approx \frac{E_{s2} - (V_v' - I_v R_{d2})}{R_s + R_{d2}} \tag{5.13}$$

Again, correlation between measured and calculated pulse widths is excellent. The maximum repetition rate is limited by the time required for the circuit current to decrease from point 3 to point b in Figure 5.12b. This time is given by equation 5.8 with the exception that I_v is replaced by I_b and V_v' is replaced by V_b .

5.4 Tunnel Diode Flip-Flop^{3,14}

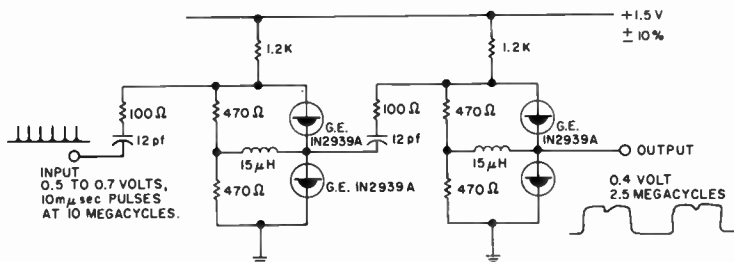
Figure 5.13 shows a tunnel diode flip-flop circuit which requires only trigger pulses of one polarity. The supply voltage is restricted to a magnitude such that only one tunnel diode can be in the high voltage state. The difference between the two tunnel diode currents flows through the inductance. When a positive trigger pulse turns the diode which is in the low voltage state to the high voltage state, the voltage induced in the inductance (because of the decreasing current through it) is of a polarity as to reset the other tunnel diode to the low voltage state. Each pair of trigger pulses completes one switching cycle.



TUNNEL DIODE FLIP-FLOP

FIGURE 5.13

The basic flip-flop circuit can be inter-connected to form a counter as shown in Figure 15.14. With the values of components and input pulses shown, the counter will operate successfully up to 10 megacycles with supply voltage tolerances of $\pm 10\%$.



TUNNEL DIODE TWO STAGE COUNTER

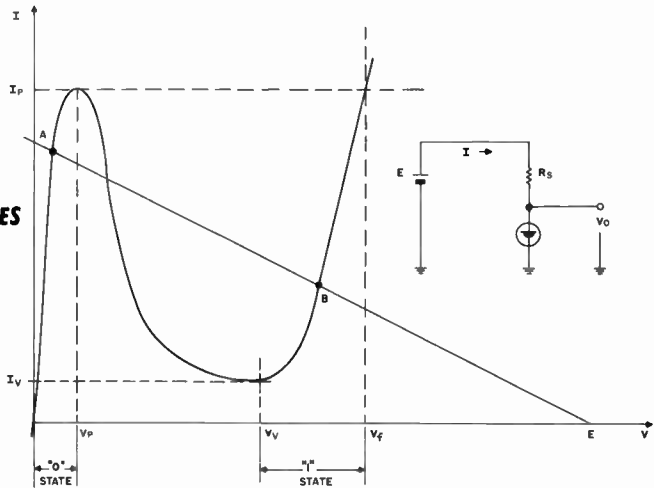
FIGURE 5.14

LOGIC CIRCUITS

The fast switching speed of tunnel diodes coupled with their potential low cost and small size make attractive their use in computer logic.

In Chapter 5 it was shown that the tunnel diode can be switched between its low and high voltage states by controlling the current through it. This resultant change in voltage can be used in logic circuits to distinguish between a "1" and a "0". As an example (see Figure 6.1), the low voltage state can be called a "0", and it can have a value ranging from zero to the peak voltage V_p , depending upon the magnitude of the current. The high voltage state can be called a "1", and it can have a value ranging between the forward voltage V_f and the valley voltage, V_v . The time required to switch between the low and high voltage states can be less than a nanosecond.

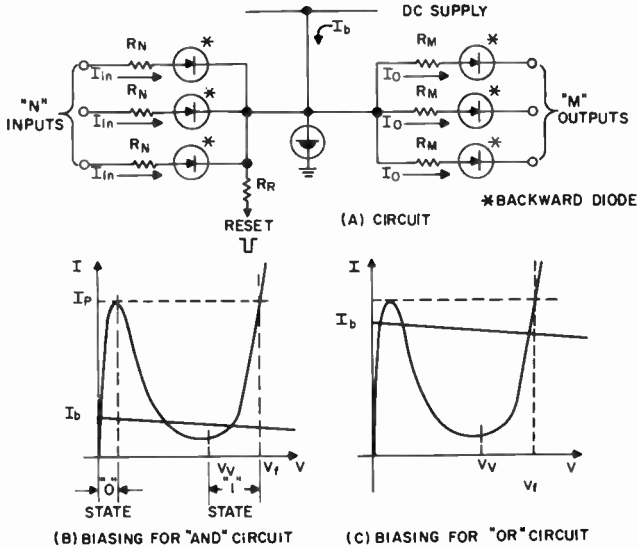
**LOGIC VOLTAGE STATES
OF THE
TUNNEL DIODE
FIGURE 6.1**



6.1 Simple Analog-Threshold Logic

A very simple circuit for performing both "and" and "or" logic is shown in Figure 6.2. This has been called analog-threshold logic rather than threshold logic because the inputs are connected together and summed before reaching the threshold device - the tunnel diode¹⁴. In normal diode threshold logic circuits each input circuit has its own thresholding device - the diode. With analog-threshold logic the allowable tolerance in the amplitude of each signal is a function of the number of inputs: the larger the number of inputs, the smaller the allowable tolerances.

The same circuit configuration (Figure 6.2a) can be used to perform both "and" and "or" functions. In either case, the tunnel diode is normally biased in the low voltage state with the magnitude of the bias current I_b determining whether one or all the inputs must be present in order to drive the tunnel diode to its high voltage state.



ANALOG-THRESHOLD LOGIC

FIGURE 6.2

Once in the high voltage state, the tunnel diode can be returned to the low voltage state by reducing the supply voltage to zero or by a negative reset pulse.

In the analysis which follows, it is assumed that the logic stage is driving and is being driven from similar tunnel diode logic stages. The input coupling resistors R_N may or may not be equal to the output coupling resistors R_M . The number of input circuits or the "fan-in" is N , while the number of output circuits or the "fan-out" is M . Unilateral flow of information is obtained by the use of backward diodes.

"And" Circuit Analysis

The tunnel diode in Figure 6.2 is in the low voltage or "0" state. For an "and" circuit, each of the tunnel diodes driving the "fan-in" resistors must be in the high voltage state for the tunnel diode under consideration to switch to the high voltage state. However, the value of voltage of the input tunnel diodes may vary from V_f to V_v , depending on the number of loads each is driving. In addition, tolerances of supply voltage, resistors, and tunnel diode parameters will affect the values of I_{in} , I_b , and threshold current I_p . Two requirements which must be met for the circuit to function properly are:

$$(N-1) I_{in \max.} + I_b \max. < I_p \min. \tag{6.1}$$

$$(N) I_{in \min.} + I_b \min. > I_p \max. \tag{6.2}$$

When the tunnel diode has switched to the "1" or high voltage state,

the bias current must be large enough to supply the load current in addition to the tunnel diode current, and the tunnel diode current must be larger than the valley current, otherwise the tunnel diode will switch back to the low voltage state. Thus, the requirement to be met when the tunnel diode is in the high voltage state is:

$$I_b \text{ min.} - M I_o \text{ max.} > I_v \text{ max.} \quad (6.3)$$

If the resistance tolerance is α , the supply tolerance Δ , and the tunnel diode parameter tolerance is β ; then, in the limiting case, equations 6.1, 6.2, and 6.3 can be written respectively:

$$(N-1) \left[\frac{V_f (1 + \beta) - V_p (1 - \beta)}{R_N (1 - \alpha)} \right] + I_b \frac{(1 + \beta)}{(1 - \alpha)} = I_p (1 - \beta) \quad (6.4)$$

$$N \left[\frac{V_v (1 - \beta) - V_p (1 + \beta)}{R_N (1 + \alpha)} \right] + I_b \frac{(1 - \Delta)}{(1 + \alpha)} = I_p (1 + \beta) \quad (6.5)$$

$$M \left[\frac{V_v (1 + \beta) - V_p (1 - \beta)}{R_M (1 - \alpha)} \right] - I_b \frac{(1 - \Delta)}{(1 + \alpha)} = - I_v (1 + \beta) \quad (6.6)$$

By setting R_N equal to R_M , by solving equations 6.4, 6.5, and 6.6 simultaneously, and by eliminating R , I_b and I_p , the relationship between N and M can be obtained in terms of the tunnel diode voltages, the tolerances, and the peak-to-valley ratio. For example, the above equations are solved for various tolerances using a germanium tunnel diode with $I_p/I_v = 10$, $V_p = .055$ volts, $V_v = .35$ volts, and $V_f = .5$ volts:

(a) Case of zero tolerance ($\alpha = \beta = \Delta = 0$)

$$.295N + .295M = 1.1 I_p R$$

N and M are not dependent upon one another and M can be made large by increasing R . The maximum "fan-in", however, is 3. (This is determined by eliminating I_b and I_p when 6.4 and 6.6 are solved simultaneously.)

(b) 2% Tolerance ($\alpha = \beta = \Delta = .02$)

$$.193N + .062M = .45$$

For $M = 1$, N can only be 2

(c) 3% Tolerance ($\alpha = \beta = \Delta = .03$)

$$0.20N + 0.093M = .45$$

For $M = 1$, N can only be 1

Thus the tolerance problem is very severe for a worse case design of the "and" circuit. In actuality, it is more severe than indicated since ideal backward diodes are assumed and capacitance and overdrive effects have not been considered.

"Or" Circuits

The circuit in Figure 6.2 can perform the "or" function if the bias current is increased sufficiently such that only one input is required to switch the tunnel diode into the high voltage state. If backward diodes are used in the coupling circuits, then the requirements to be met are:

$$I_{in \text{ min.}} + I_{b \text{ min.}} > I_{p \text{ max.}} \tag{6.7}$$

and
$$I_{b \text{ max.}} < I_{p \text{ min.}} \tag{6.8}$$

Equations 6.7 and 6.8 can be written respectively in terms of the tolerances in the limiting case as:

$$\frac{V_v (1 - \beta) - V_p (1 + \beta)}{R_N (1 + \alpha)} + I_b \frac{(1 - \Delta)}{(1 + \alpha)} = I_p (1 + \beta) \tag{6.9}$$

and
$$I_b \frac{(1 + \Delta)}{(1 - \alpha)} = I_p (1 - \beta) \tag{6.10}$$

R can be calculated from 6.9 and 6.10 if the tolerances, peak current and voltages are known. If backward diodes are not used, then the current which flows in the N-1 inputs which are in the low voltage state must be taken into account in equation 6.9.

After the tunnel diode has switched to the high voltage state, the "fan-out" equation is the same as given in equation 6.6. If backward diodes are not used in the "fan-in" circuits, then the current drain of the input circuits must be added to equation 6.6.

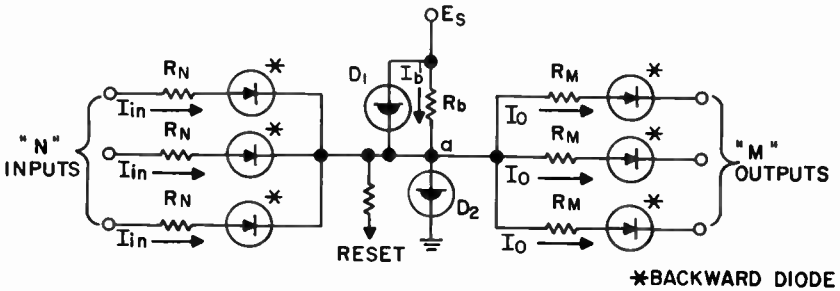
6.2 Chow's Circuit

Figure 6.3 shows a circuit developed at the G-E Electronics Laboratory by W. F. Chow which reduces the dependence of N on M as compared to the analog-threshold "and" circuit. The difference between this and the previous circuit is the addition of the tunnel diode D_1 across R_b and the reduction of the supply voltage, E_s , to a value which is large enough to permit only one of the tunnel diodes to be in the high voltage state. Thus, $E_s \approx V_v + V_p$. The value of R_b is determined by the number of inputs. Before each logic operation D_2 is set into the low voltage state and D_1 is set in the high voltage state. The bias current in D_2 is then the current through R_b and the valley current of D_1 . When the sum of the input and bias currents exceeds the peak current of D_2 and switches D_2 into the high voltage state, D_1 switches to the low voltage state, essentially returning point "a" to the supply voltage. (The dynamic resistance of D_1 in the low voltage stage is much less than R_b .) Consequently, the total output current, MI_o , is limited only by the difference between the peak current of D_1 and the valley current of D_2 . It is assumed that backward diodes are used to insure the unilateral propagation of information. The general requirements which must be met to insure that D_2 will only switch to the high voltage state when all the inputs are present are given by

equations (6.1) and (6.2). I_b , however, is now equal to $I_v + I_b'$. The requirement to be met after D_2 has switched to the high voltage state is:

$$I_{v2 \text{ max.}} + M I_o \text{ max.} < I_{p1 \text{ min.}} \tag{6.11}$$

where $I_{v2 \text{ max.}}$ is the maximum valley current of D_2 and $I_{p1 \text{ min.}}$ is the minimum peak current of D_1 . (The peak current of D_1 and D_2 need not be the same.) In the analysis that follows, it is assumed that the tunnel diode stage under consideration (Figure 6.3) is driving and being driven by similar types of logic circuits.



CHOW'S ANALOG-THRESHOLD LOGIC CIRCUIT

FIGURE 6.3

If the supply voltage tolerance is Δ , the resistor tolerances are α , the tunnel diode parameter tolerances are β , and the bias current tolerance is Ω , then the limit worst case design expressions of 6.1, 6.2 and 6.11 are respectively for the "and" operation:

$$(N-1) \left[\frac{E_s (1 + \Delta) - V_p (1 - \beta)}{R_N (1 - \alpha)} \right] + I_b (1 + \Omega) = I_{p2} (1 - \beta) \tag{6.12}$$

$$N \left[\frac{E_s (1 - \Delta) - 2 V_p (1 + \beta)}{R_N (1 + \alpha)} \right] + I_b (1 - \Omega) = I_{p2} (1 + \beta) \tag{6.13}$$

$$M \left[\frac{E_s (1 + \Delta) - 2 V_p (1 - \beta)}{R_M (1 - \alpha)} \right] + I_{v2} (1 + \beta) = I_{p1} (1 - \beta) \tag{6.14}$$

where:

$$I_b = I_{v1} + \frac{E_s - V_p}{R_b}$$

and

$$E_s = V_v + V_p$$

If the tolerances are set equal to zero, equations 6.12 and 6.13 can be solved to give the bias current in terms of N and the other parameters. Equations 6.12, 6.13 and 6.14 can be solved to give the relationship between M and N in terms of the tolerances, peak voltage, supply voltage, peak-to-valley ratio, and the ratio of the peak currents of the

two tunnel diodes. By making the peak current of D_1 larger than that of D_2 , a larger "fan-out" results for a given "fan-in".

As an example, if D_1 and D_2 are respectively 2.2 ma and 1ma germanium tunnel diodes with peak-to-valley ratios of 10, and $E_s = .45$ volts, $V = .055$ volts and the tolerances are $\alpha = \beta = \Delta = 3\%$ and $\Omega = 6\%$, then^p the relationship between M and N is:

$$0.64N + 0.29M = 3.5 \tag{6.15}$$

Thus, for $M = 1, N = 5$; and for $M = 3, N = 4$.

If the tolerances are set equal to zero, equations 6.12 and 6.13 can be solved for the maximum possible "fan-in":

$$N = \frac{E_s}{V_p} \quad \text{or} \quad N = \frac{V_v}{V_p} + 1 \tag{6.16}$$

For germanium tunnel diodes, the maximum possible "fan-in" (zero tolerance conditions) is 7. As the tolerances are increased, the "fan-in" is reduced.

6.3 The Goto or "Twin" Circuit¹⁵

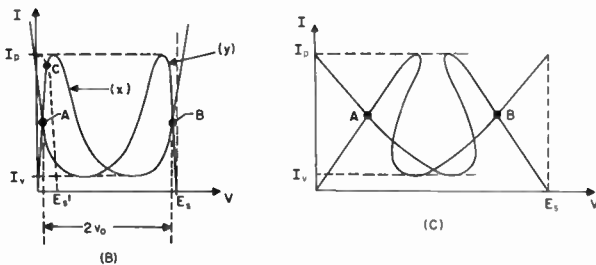
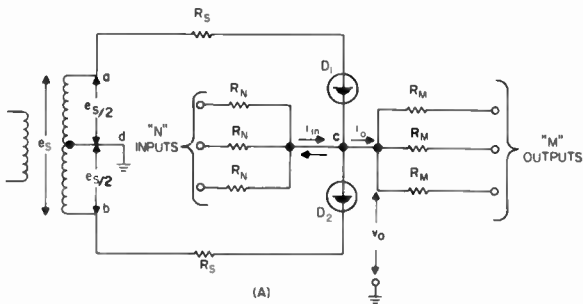
Binary "1" 's and "0" 's can be represented respectively by positive and negative unit voltages instead of positive unit voltages and small or zero voltages, and the logic operation can be based on an analog sum or majority of the voltages. Thus, the output of a gate used in performing majority logic will be dependent upon whether a majority of the inputs is either positive or negative. Consequently, the number of inputs for the majority gate must be an odd number if ambiguity in the output circuit is to be avoided.

Figure 6.4 shows a circuit which will give both a positive and negative output, and which can be used as a majority gate. The supply voltage, which can be in the form of a square wave or sinusoid, is obtained from the secondary of a center-tapped transformer. The tunnel diodes D_1 and D_2 are in series with the supply, and the input and output voltages are taken with respect to the junction of the tunnel diodes and the grounded center-tap of the transformer.

The tunnel diodes are selected to have closely matched peak currents. During the negative half-cycle (when the point "b" is positive with respect to point "a"), both tunnel diodes are conducting in the reverse direction. Resistors or non-linear devices such as backward diodes must be added in series with the tunnel diode in order to limit the current flow during this half-cycle. Point "c" will be at ground potential since the transformer, tunnel diodes, and resistors form the arms of a balanced bridge. As the supply voltage goes positive (point "a" positive with respect to point "b"), both diodes begin conducting in the forward direction. However, point "c" will remain approximately at ground potential until the current in one or the other tunnel diode exceeds its peak current. Which tunnel diode reaches its peak current first is dependent upon whether the majority of input signals is positive or negative with respect to ground. If the analog sum of the inputs is positive, the signal current will flow into point "c", and its

direction is such that it adds to the current in D_2 and subtracts from the current in D_1 . Therefore, D_2 will reach its peak current first and will switch to the high voltage state. If the supply voltage is limited such that only one tunnel diode can be in the high voltage state, the other tunnel diode is forced to remain in the low voltage state for the remainder of the half-cycle. The output voltage at "c" then is positive with respect to ground. If the majority of input signals had been negative (current flowing out of point "c"), D_1 would have reached its peak current first and consequently the output voltage at "c" would have been negative for the remainder of the half-cycle. Thus, the Goto circuit produces an output of the same polarity as the majority of the input signals.

Composite V-I characteristics of the Goto circuit are shown in Figure 6.4b for the case where R_s is neglected and in 6.4c where it is taken into account. Thus in Figure 6.4b, tunnel diode (y) is considered as the "load" and its curve is the image of the V-I curve of the other tunnel diode (x) reflected through the current axis and translated along the positive voltage axis by a distance E_s . The stable operating points are the intersections "a" and "b" in the positive resistance regions of the characteristics. As the supply voltage E_s increases from zero, the (y) characteristic or "load" is shifted horizontally along the voltage axis. Thus at a supply voltage E_s' , the stable operating point is "c".



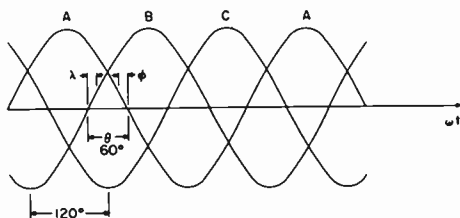
"GOTO" OR "TWIN" CIRCUIT
FIGURE 6.4

The composite characteristic of 6.4c is the same as 6.4b except that each individual characteristic curve is now a composite V-I characteristic of the tunnel diode and resistance R_s .

The Goto gate can also be used to perform "and" and "or" functions. If a positive input is considered a "1" and a negative input a "0", then the circuit in Figure 6.4a will be a two input "and" gate if one of the inputs is returned to a negative bias supply. The reason is that both of the other two inputs must be positive in order to obtain a positive output. Likewise, if one input is returned to a positive supply the circuit becomes a two input "or" gate since only one of the other inputs is required to be positive in order to obtain a positive output.

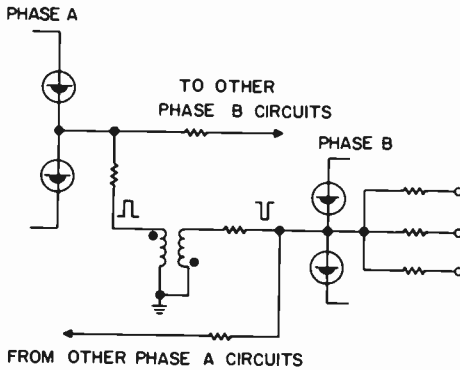
Backward diodes cannot be used with the Goto circuit to obtain a unilateral flow of information since the inputs and outputs can be of either polarity. One of the most simple ways of achieving this is to use a three phase clock. The over-lapping of the phases must be such that at any instant in time one group of circuits is capable of providing an output, a second group is sensitive to its inputs, and the third group is neither sensitive to the inputs nor is it capable of providing an output. The three states of operation can be called respectively: active, receptive, and passive.⁵

If the three phases of the supply voltage are displaced from each other in time by 120° , then information can be transferred from the circuits supplied by one phase to the circuits supplied by the next phase during the 60° of overlap (see Figure 6.5). Thus the circuits supplied by phase A drive the circuits supplied by phase B; those supplied by phase B drive those circuits supplied by phase C; which in turn drive circuits supplied by phase A. The flow of information is unilateral because during the positive overlap of two phases of the supply voltage, the voltage of the third phase is negative. The tunnel diodes supplied by this phase are inactive during this period since they are biased in the reverse state.



THREE PHASE SUPPLY FOR MAJORITY LOGIC
FIGURE 6.5

Inversion or the logical "not" function can be accomplished by the use of a transformer as shown in Figure 6.6. One of the difficulties encountered in using this method is that the magnitude of the secondary voltage is dependent upon the history of the signals applied to the primary. Thus, if the signal applied to the primary is 101010 (alternate positive and negative pulses), the voltage at the secondary will be the inverse of those pulses. (If the turns ratio is 1:1, the amplitude of the primary and secondary voltages will be the same.) On the other hand if the input is all one's or zero's (all positive or all negative pulses), the secondary voltage will be inverted, but the amplitude will be reduced because the transformer will not pass the DC component of the primary voltage. This variation of output amplitude must be taken into account when using a transformer to perform inversion.



INVERSION OR LOGICAL NOT

FIGURE 6.6

If the Goto or twin gate is to function properly as a logic element, the following conditions must be met:

1. During the receptive period and just prior to the switching of one of the tunnel diodes, the majority input current must be large enough to overcome any expected supply voltage unbalance and tunnel diode peak current mismatch.
2. During the active period when one of the tunnel diodes is in the high voltage state, the "fan-out" current and the maximum amplitude of the supply voltage must be limited so that the current through the tunnel diode in the low voltage state is less than its peak current.
3. The overlap of the phases of the supply voltage must be large enough so that the stage in the active mode can transfer its information to the succeeding stages before the information is lost. The information is lost when the supply voltage falls to a value such that the tunnel diode current is less than the valley current. Overlap is an important consideration when the supply voltage is sinusoidal.

The equivalent circuit in Figure 6.7 is used to establish the circuit requirement for meeting the first of the above conditions. For the polarity of input voltage shown (a positive majority) diode D_2 should switch to the high voltage state. The minimum available input voltage results when there is only a majority of one. For this case v_{in} becomes:

$$v_{in} = \frac{1}{N} \left[e_{np} \frac{(N+1)}{2} - e_{nn} \frac{(N-1)}{2} \right] \quad (6.17)$$

where N is the "fan-in" or number of inputs, e_{np} the positive input voltage, and e_{nn} the negative input voltage. If $e_{np} = e_{nn} = v_o$ (Figure 6.4b), then v_{in} becomes v_o/N . The source impedance R_{in} is $R_{N/N}$

where R_N is the coupling resistor of each input.

Since D_2 is to switch to the high voltage state, I_2 must reach the peak current of D_2 before I_1 reaches the peak current of D_1 . Under limiting conditions I_1 and I_2 become equal to I_{p1} and I_{p2} respectively at the same instant. Therefore, the circuit requirement under these conditions becomes:

$$I_2 - I_1 = I_{p2} - I_{p1} \tag{6.18}$$

If the per unit tolerance in peak current is β , then equation 6.18 becomes:

$$I_2 - I_1 = 2\beta I_p \tag{6.19}$$

where I_p is the nominal value of peak current.

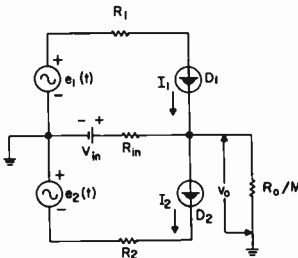
The expression relating $I_2 - I_1$ to the circuit voltages and resistances is given in Figure 6.7. This expression is obtained by solving the loop equations and by assuming the voltage drop across the tunnel diodes to be zero. Notice that the first term of the expression is a function of the supply unbalance, while the second term is a function of the input majority.

The relationship between the "fan-in" and "fan-out" in terms of the circuit tolerances can be obtained by substituting 6.19 for $I_2 - I_1$ in the equation of Figure 6.7; by letting $e_1 = e(1 + \Delta)$, $e_2 = e(1 - \Delta)$ and $v_{in} = v_{in}(1 - \Omega)$; and by assigning resistance tolerances so as to make the right hand side of the equation as small as possible. If $R_N = R_M$, and $R_N/M + N \gg R_1 R_2 / R_1 + R_2$, then the relationship becomes:

$$2 I_p \beta \approx - \frac{e(\alpha + \Delta)(M + N)}{R_N} + \frac{N v_{in}(1 - \Omega)}{R_N} \tag{6.20}$$

If the Goto circuit is to operate over a wide temperature range, the value of v_{in} at the highest operating temperature should be used in 6.20.

**EQUIVALENT
CIRCUIT
OF THE
GOTO STAGE
DURING THE
RECEPTIVE
PERIOD**



$$I_2 - I_1 = \frac{R_1 e_2 - R_2 e_1}{R_2 R_1 + \frac{R_{in} R_0 (R_1 + R_2)}{R_0 + M R_{in}}} + \frac{R_0 v_{in}}{R' (R_0 + M R_{in}) + R_{in} R_0}$$

$$R' = \frac{R_1 R_2}{R_1 + R_2}$$

$$R_{in} = \frac{R_N}{N}$$

$$v_{in} = \frac{v_0}{N} \text{ FOR A MAJORITY OF } I$$

SEE FIG. 6.4 (B) FOR V_0

After diode D_2 has switched, the peak value of the supply voltage must be limited so that the current through D_1 does not exceed its peak current. The equivalent circuit of Figure 6.7 can be used to determine the circuit requirements to meet this condition if v_{in} is set equal to zero. If I_L is the total current which flows through fan-in resistors R_N and the fan-out resistors R_M , and if the current in D_2 is I_2' , then

$$I_{p1} \text{ min.} > I_2' \text{ max.} + I_L \text{ max.} \quad (6.21)$$

If $R_N = R_M$, $R_1 = R_2 = R$, then under limiting conditions 6.21 can be written:

$$I_{p1}(1-\beta) - I_2'(1+\beta') \left[1 - \frac{R}{R + \frac{R_M(1-\alpha)}{M+N}} \right] = \frac{e_m(1+\Delta)}{R + \frac{R_M(1-\alpha)}{M+N}} \quad (6.22)$$

where β is the per unit tolerance of I_2' and e_m is the peak value of the supply voltage if a sinusoidal supply is used. From 6.22 the maximum permissible value of supply voltage is given in terms of M , N , the circuit parameters and tolerances. If the peak value of the supply voltage is to be made as large as possible, I_2' should be a minimum.

When a sinusoidal supply voltage is used, sufficient overlap must exist so that the information is passed to succeeding stages before the logic stage under consideration loses its information. The information is lost because at some angle θ the supply voltage has decreased to a value where it is unable to sustain a current through D_2 in excess of the valley current. The angle θ (Figure 6.5) where the logic stage is reset is given by:

$$\sin \theta = \frac{V_v}{e_m \left[2 - \frac{R}{R + \frac{R_M}{M+N}} \right]} + \frac{R I_v}{e_m} \quad (6.23)$$

where e_m is the peak value of the supply voltage with respect to ground and V_v and I_v are the valley voltage and currents respectively of the tunnel diode in the high voltage state. Equation 6.23 is written assuming that $R_N = R_M$ and $R_1 = R_2 = R$. For worst case conditions, the tolerances of the circuit parameters are included in 6.23 so as to increase θ . The value of e_m can be obtained from 6.22.

Before the logic is reset (as given by θ), the succeeding stages must receive the information and switch to the proper state. The angle λ (Figure 6.5) at which they can switch (assuming a positive input) occurs when the supply voltage has increased sufficiently so as to make I_2 equal to I_{p2} . Thus for $R_1 = R_2 = R$, $R_N = R_M$ and $e_1 = e_2 = e_m \sin \lambda$:

$$\sin \lambda = \frac{R I_{p2}}{e_m} - \frac{M R v_{in}}{2 R_M + (M+N)R} \quad (6.24)$$

For worst case design conditions, the circuit tolerances are included in 6.24 so as to increase λ . Thus, for proper operation of the Goto circuit, $\lambda + \theta$ must be less than the overlap angle, ϕ , as shown in Figure 6.5.

At high frequencies (above a megacycle) or when a square wave is used, the capacitance mismatch of the tunnel diodes becomes important, and it is reflected in the performance of the stage as a mismatch in peak currents.

TUNNEL DIODE TEST CIRCUITS

The measurement of tunnel diode parameters requires considerable care to insure that the test circuit is in a stable state while performing the test. Oscillation or switching is usually avoided by appropriate "loading" or "damping" of the circuit while in some special cases it may even be encouraged in order to measure a specific parameter more accurately.

The parameters to be tested depend on the application. For example, if the device is used as an oscillator or amplifier, the negative conductance $|g_d|$, the capacitance C , the inductance L_S and the series resistance R_S must be known (see the tunnel diode equivalent circuit in Figure 2.5).

As a switching circuit element, it might be more desirable to know I_p , I_v , V_p , V_v , V_{fp} and switching speed (t_r). The following test circuits are designed to measure these parameters. Some of these circuits will yield readings with accuracies of only +10 to 20% and are quite simple in nature. Others are designed for high accuracy and naturally will be more complex.

7.1 V-I Curve Tracer

Observing the tunnel diode V-I characteristic is not always an easy task. Conventional curve tracers usually have enough series resistance in their sweep circuits as to appear as a current source to the tunnel diode under test, that is,

$$R_T > \frac{1}{|g_d|} \quad (7.1)$$

where R_T is the total DC resistance in series with the tunnel diode being tested. As a result, the displayed V-I characteristic lacks the negative conductance portion as shown in Figure 7.1.

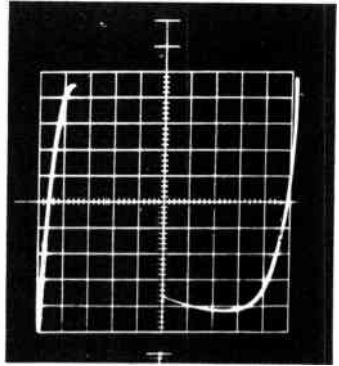
This figure shows that as the sweep passes the peak point current, the diode switches very quickly from V_p to some value of forward voltage dependent on the load line, without ever going through the negative conductance region.

This problem becomes even worse when testing larger peak current units as they will exhibit smaller values of negative resistance. The sweep circuit will thus look more like a current source to such units enabling them to "switch" more readily.

Even when the sweep circuit resistance is small compared to $|rd|$, still another problem might prevent the undistorted display of the V-I curve. Figure 7.2 illustrates this problem which is caused by ex-

**V-I CHARACTERISTIC AS DISPLAYED
BY CURVE TRACER WITH LARGE
INTERNAL SERIES RESISTANCE**

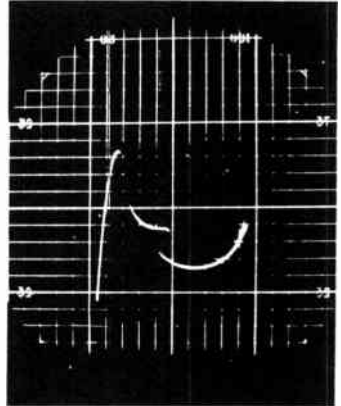
FIGURE 7.1



cessive series inductance in the test circuit. As the sweep goes through the negative conductance region, this inductance will resonate with the diode and circuit capacitances, making the circuit oscillate and causing a display as shown in Figure 7.2.

**V-I CHARACTERISTIC OF OSCILLATING
TUNNEL DIODE**

FIGURE 7.2

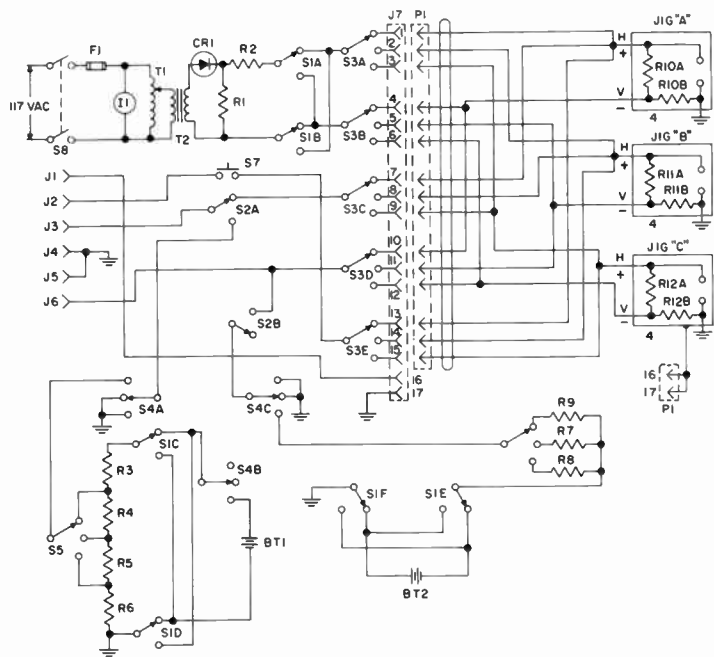


If the total circuit inductance (including the diode inductance L_S) is kept at a minimum, this oscillation will not occur. Actually, if the circuit self-resonant frequency f_{xo} (circuit) is made larger than its cut-off frequency, f_{ro} (circuit), no oscillations can occur. To meet this criterion, the total circuit inductance must be smaller than:

$$L_{\text{total}} < \frac{R_{\text{Total}} \times C_{\text{total}}}{g_d} \quad (7.2)$$

Here again larger peak current units will be more troublesome to test, as one can see from equation 7.2 that if $|g_d|$ is increased, the maximum permissible circuit inductance is further reduced.

The main features of a well working V-I curve tracer capable of displaying the full V-I characteristic of tunnel diodes with a wide range of peak currents, are thus:



PART NO.	NAME	DESCRIPTION
BT1	BATTERY	1.34 VDC MALLORY #RM-12R *
BT2	BATTERY	1.34 VDC MALLORY #RM-12R *
CR1	DIODE	1N1695
F1	FUSE	1AMP
L1	LAMP	NE-5
J1	BINDING POST	SUPERIOR ELECTRIC TYPE DF30BC *
J2	BINDING POST	SUPERIOR ELECTRIC TYPE DF30RC *
J3	BINDING POST	SUPERIOR ELECTRIC TYPE DF30RC *
J4	BINDING POST	SUPERIOR ELECTRIC TYPE DF30BC *
J5	BINDING POST	SUPERIOR ELECTRIC TYPE DF30BC *
J6	BINDING POST	SUPERIOR ELECTRIC TYPE DF30RC *
J7	SOCKET	AMPHENOL #77-MIP-20, 20 PINS *
P1	PLUG	AMPHENOL #70-20, 20 PINS *
R1	RESISTOR	47Ω, 2W
R2	RESISTOR	20Ω, 10W
R3	RESISTOR	340Ω, 1/2 W, 1/4 %
R4	RESISTOR	500Ω, 1/2 W, 1/4 %
R5	RESISTOR	400Ω, 1/2 W, 1 %
R6	RESISTOR	100Ω, 1/2 W, 1 %
R7	RESISTOR	264.6Ω, 5 %
R8	RESISTOR	1353Ω, 5 %
R9	RESISTOR	131.4Ω, 5 %
S1	SWITCH	6 POLE, 2 POSITION
S2	SWITCH	2 POLE, 2 POSITION
S3	SWITCH	5 POLE, 3 POSITION
S4	SWITCH	3 POLE, 3 POSITION, MOMENTARY CONTACT
S5	SWITCH	1 POLE, 3 POSITION
S6	SWITCH	1 POLE, 3 POSITION
S7	SWITCH	N.O., MOMENTARY CONTACT, PUSH BUTTON
S8	SWITCH	DPST, TOGGLE
T1	TRANSFORMER	POWERSTAT TYPE 108 *
T2	TRANSFORMER	STANCOR #P-6134 *
XF1	FUSE HOLDER	FUSETRON TYPE HJM *
X11	LAMP HOLDER	OIALCO #95408-931 *
1	ASSEMBLY	
2	CABINET	PREMIER #SFC-502 - 8"X14"X8" *
3	CHASSIS	PREMIER #ACH-426-5"X7"X2" ALUMINUM *
4	CURVE TRACING JIG	FOR DETAILED DRAWING SEE FIG. # 7.5
R10A B B	RESISTORS	10Ω PYROFILM RESISTOR CO #D375 CARBON FILM MICROWAVE RES.
R11A B B	RESISTORS	4Ω PYROFILM RESISTOR CO #D375 CARBON FILM MICROWAVE RES.
R12A B B	RESISTORS	2Ω PYROFILM RESISTOR CO #D375 CARBON FILM MICROWAVE RES.

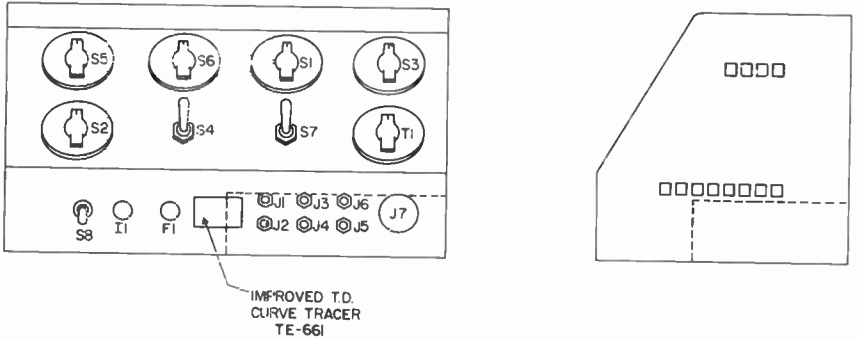
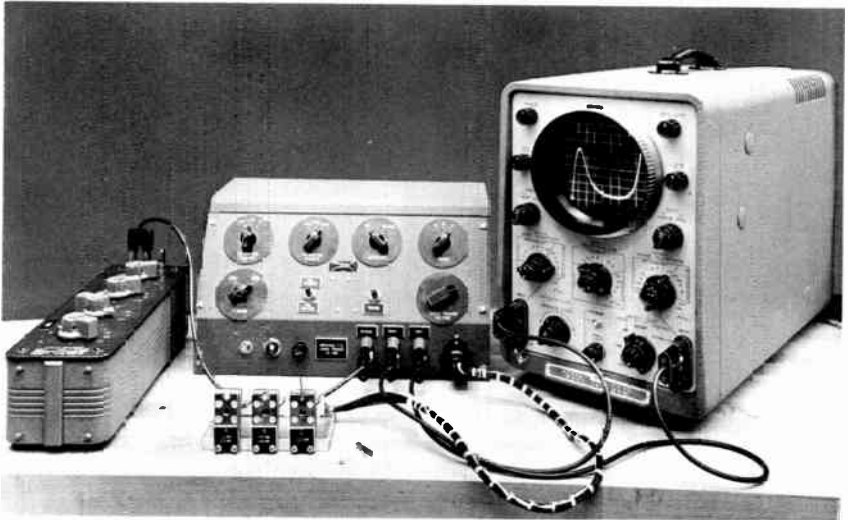
* OR EQUIVALENT

V-I CURVE TRACER CIRCUIT DIAGRAM

FIGURE 7.3

1. A low series resistance sweep circuit and,
2. Low inductance test heads.

The curve tracer circuit shown in Figure 7.3 and pictured in 7.4 covers a range of units from a fraction of one milliampere to 22 ma. With appropriate calibration (see Table of Calibration Procedures), this test set will determine, I_p , I_v , I_p/I_v , V_p , V_v , V_{fp} , V_r as well as to give a rough idea of the positive slopes and the negative



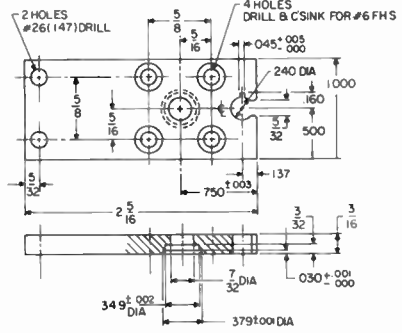
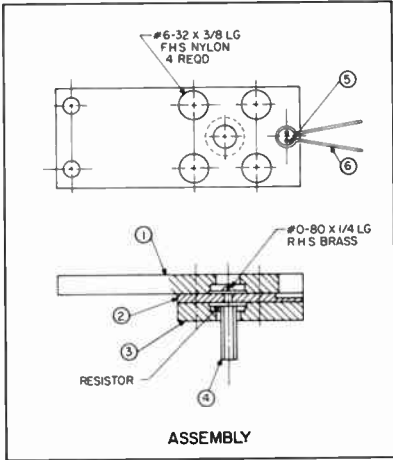
PHOTOGRAPH OF CURVE TRACER TEST SET UP

FIGURE 7.4

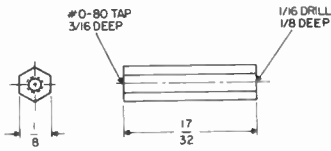
conductance slope. By the use of the shunting decade box, a rough idea can also be obtained of the average value of the negative resistance ($1/g_d = d_v/d_i$).

Figure 7.5 is a mechanical drawing detailing the construction technique of the jigs.

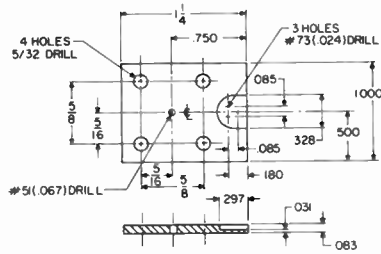
TUNNEL DIODE TEST CIRCUITS



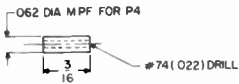
① BASE PLATE MATERIAL : BRASS



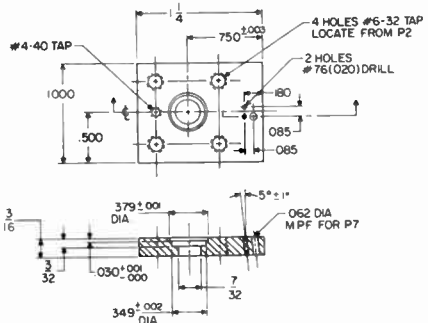
④ POST MATERIAL : BRASS HEX



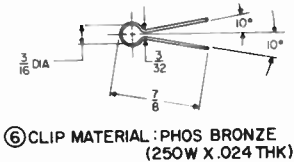
② SPACER MATERIAL : TEFLON*



⑤ BUSHING MATERIAL : TEFLON



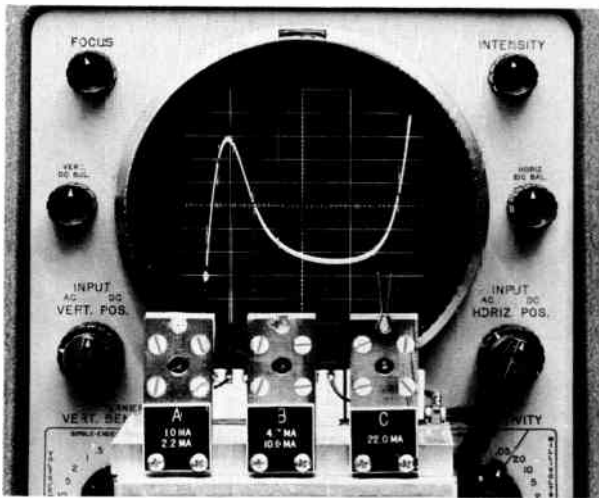
③ BOTTOM PLATE MATERIAL : BRASS



⑥ CLIP MATERIAL : PHOS BRONZE (250W X .024 THK)

NOTE: FOR JIGS B & C USE ONLY 3 MIL THICKNESS MILAR TAPE AS SPACER

OUTLINE DRAWING OF LOW INDUCTANCE TEST JIGS
FIGURE 7.5



CLOSE-UP PHOTOGRAPH OF JIG AND SCOPE PRESENTATION
FIGURE 7.6

This instrument is made more versatile by the small size, low inductance test heads. Since the bias resistors are incorporated in the test heads, these may be connected with long cables to the main chassis and the decade box. As a result, one can conveniently and stably test tunnel diodes in a temperature chamber at some distance from the test set. Figure 7.6 is a photograph of the V-I curves of a 22 ma germanium tunnel diode.

Operating Instructions for Curve Tracer

The three outputs of the curve tracer are connected to the Vertical and Horizontal inputs of a scope (HP 130B or equivalent) and a Resistance Decade Box. Note: To display the V-I characteristic "right side up", the vertical scope input must be inverted and the ground strap removed from what is then the "hot" lead.

The Selection Switch must be in the germanium (Ge) or gallium arsenide (GaAs) position, depending upon the type of unit under test.

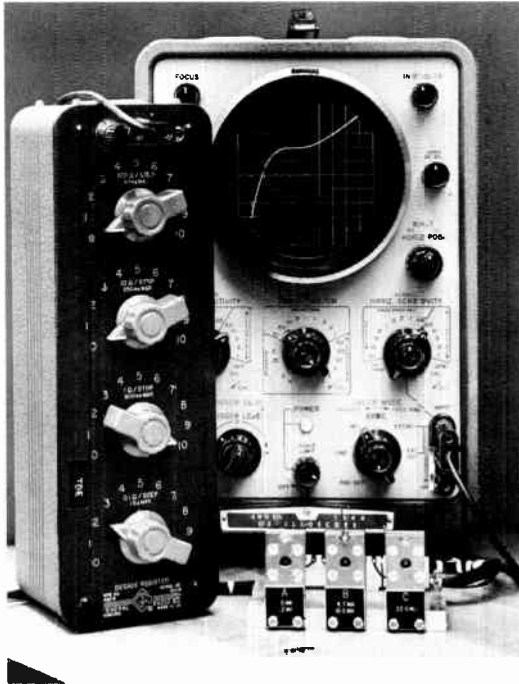
The Range Switch must be in the A, B or C position, depending upon the jig in use.

Jig A is for 1.0 and 2.2 ma units. Jig B is for 4.7 and 10.0 ma units. Jig C is for 22.0 ma units.

The Function Switch must be in operating position.

With the Curve Tracing Voltage Control in the counter-clockwise position, insert the diode. Increase the voltage until the trace is beyond the valley region.

Press the Decade Switch and adjust the decade box until the negative slope is parallel with the horizontal axis. This setting on the decade box is then the terminal negative resistance of the unit being tested. Figure 7.7 shows the resulting curve trace and the setting of the resistance box for a 10 ma unit.



DETERMINATION OF TERMINAL NEGATIVE RESISTANCE

FIGURE 7.7

Scope Calibration

To calibrate the horizontal axis of the scope, the Function Switch must be in the calibration position. In this position, the scope output terminals are shorted and the calibration voltage switch in either the .1v, .5v or 1.0v position. The "Calibrate Voltage-Calibrate Current" switch is then pressed in the "Cal. Voltage" position. The calibrated voltage will then appear across the output terminals.

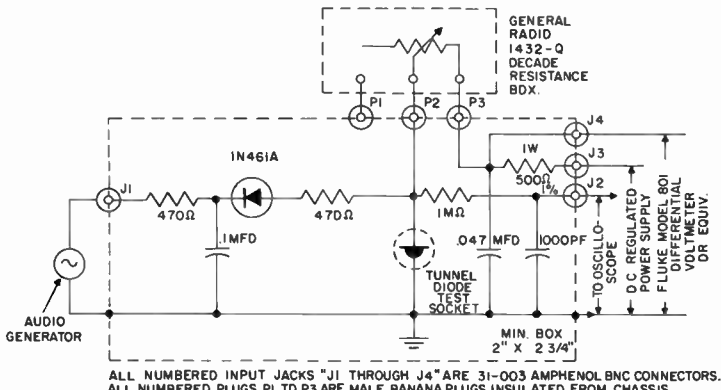
The same applies to the calibration of the vertical axis, with the exception that, if a 1 ma calibration is desired, the Range Switch must be in the A position and the diode in Jig A removed. If the 5 ma calibration is desired, the Range Switch must be in the B position and the diode in Jig B removed. For 10 ma calibration, the Range Switch must be in the C position and the diode in Jig C removed.

7.2 High Precision Peak Point Current (I_p) Test Set

Some applications require very accurate peak point current control. For instance, a threshold switch or a coincidence gate circuit require a stable switching threshold. To meet this demand, some tunnel diode types are specified with a peak point current rating of plus or minus 2.5%. To test accurately to these limits, the peak current test set must be capable of measuring the test parameters

within a fraction of one percent. The following peak point current test set is capable of such accuracy.

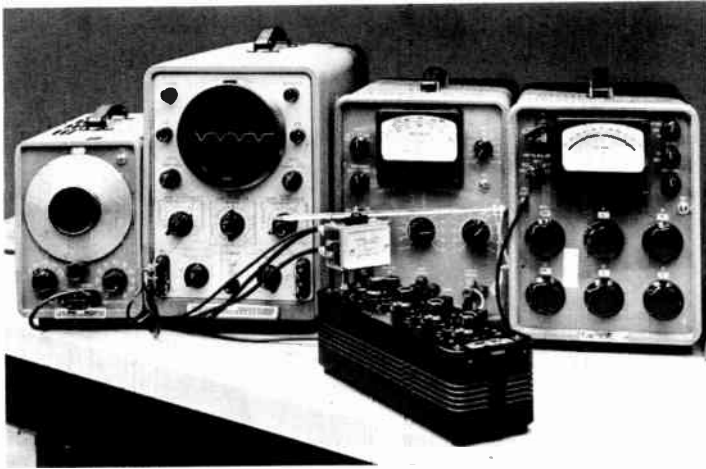
The circuit (see Figure 7.8) is rather simple and consists basically of a tunnel diode driven from an accurately known current source. The DC input is increased to a point just exceeding the peak point current. One can observe this point visually on a scope connected across the tunnel diode. As the peak current is exceeded, the diode switches to its high voltage state which can be observed on the scope as a change in DC level (using a DC scope). At this point, one would still have only a rough approximation of the peak point and hence a small audio sine wave signal is applied. The test set rectifies this signal and applies it to the tunnel diode in a direction subtracting from the applied DC. When the AC signal is large enough to bring the "composite" bias just below the valley point, the tunnel diode will switch back to its low voltage state again, a fact immediately evident on the scope. Subsequently, the applied DC is very gradually reduced and the same is done with the AC signal. The idea is to "fine tune" the DC level as close as possible to the peak point so that the smallest possible AC signal will switch the circuit to its low voltage state. When this is satisfactorily accomplished, the circuit will turn "ON" and "OFF" erratically, but continuously. At this point, a differential voltmeter will very accurately determine this critical DC supply voltage. The use of a high precision resistance decade box will, by simple use of Ohm's Law, yield the current reading.



PEAK CURRENT TEST SET

FIGURE 7.8

The absolute accuracy of this current reading is about $\pm 0.1\%$, however, the degree of indecision as to the exact point at which the circuit turns "ON" and "OFF" consistently, adds about 0.2% potential judgement error to the reading. It is important, in order to achieve high accuracy, that the circuit be totally shielded, properly grounded and that the power supply be well regulated. Since I_p does change somewhat with temperature, a thermometer can be placed near the test socket for temperature readings. Figure 7.9 is a picture of the complete test set up.



PHOTOGRAPH OF PEAK CURRENT TEST SET UP

FIGURE 7.9

7.3 Tunnel Diode Junction Capacitance Test Set

In previous chapters the tunnel diode equivalent circuit has been analyzed and it can be shown that the apparent capacity looking into the device terminals is:

$$C_{\text{apparent}} \cong C_{\text{junction}} + C_{\text{strays}} - L_s g_d^2 \quad (\text{when } \omega \ll g_d/c)$$

In addition, if a bias resistor suitable for biasing the tunnel diode into the negative-conductance region is used, and the inevitable inductance of the leads coupling the measuring node to the biased device is considered, then the capacitance measured by the bridge may be shown to be:¹⁰

$$C_{\text{measured}} = C_j + C_{\text{strays}} - L_{\text{leads}} (g_{\text{bias}} + g_d)^2 - L_s g_d^2$$

Since the inductance of the leads to the bridge can easily be the dominant reactance in the circuit, the method described is seen to be subject to large potential errors even when stringent efforts are made to keep all inductances to the irreducible minimum. Unfortunately this method is necessary in order to measure the variation of junction capacitance with forward voltage in the negative-conductance region.

One further item will be considered before discussing the measurement method to be proposed.

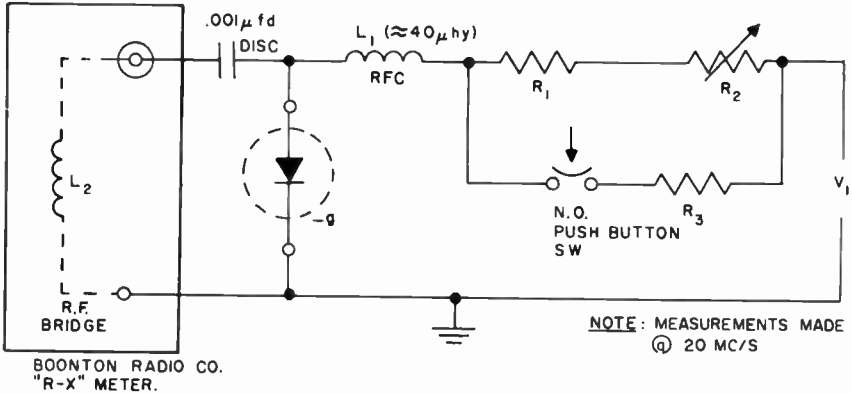
From the stability analysis of the equivalent circuit one finds that for the tunnel diode to be stable in the negative-conductance region,

$$\frac{L_s |-g_d|}{C} < R_s < \frac{1}{|-g_d|}$$

In other words, if the circuit inductance cannot be made less than $R_s C / |-g_d|$ the circuit must oscillate, and any capacitance measured bears only a fortuitous resemblance to the actual capacity. (Even in the valley region the measuring signal will drive the tunnel diode slightly into a region of some negative-conductance for part of the time, and the possibility of oscillations exists.)

Consider an increase of the bias resistance to the limit of $1/|-g_d|$ in order to improve the circuit stability. Then L_s must be less than C/g_d^2 (or conversely $C_j > L_s g_d^2$) or oscillations will persist and the circuit cannot be stabilized. Since the ultimate limits being considered vary as the square of the negative-conductance (a direct function of peak point current) it is apparent that a test circuit which may be stabilized for a low-current tunnel diode will eventually become unstable for some higher current device, due to the limitations of a fixed circuit and/or package inductance.

With the preceding considerations in mind, one sees that the design of a test circuit for measuring the capacitance of tunnel diodes over a large range of peak point currents and g_d/c ratios, can only be accomplished by deliberately violating both stability criteria simultaneously, and forcing stability by using a switching load line to bias the device. It should be also noted that the sum of junction and distributed capacitance can only be measured over a small voltage range around the valley point voltage. Figure 7.10 illustrates the resultant test circuit.



BASIC TUNNEL DIODE CAPACITANCE TEST CIRCUIT

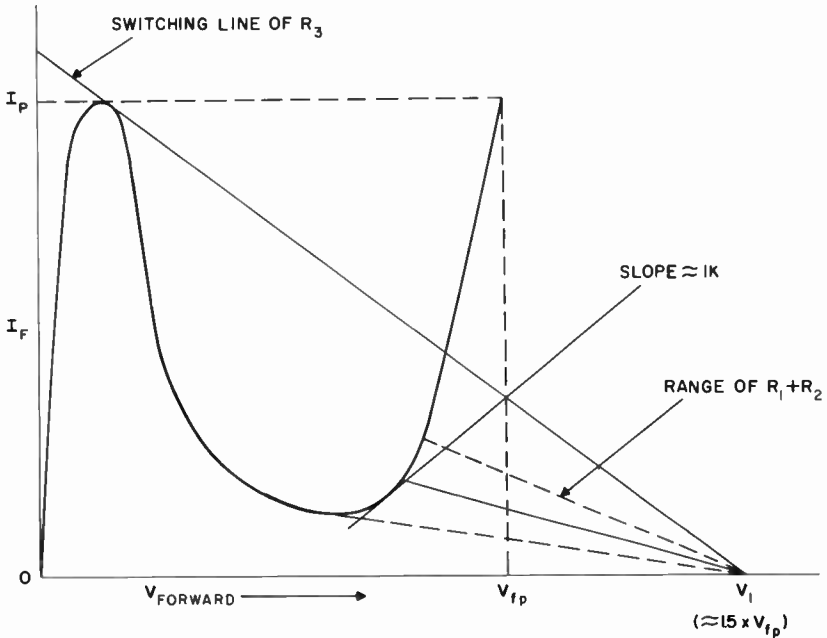
FIGURE 7.10

Figure 7.11 shows the region of operation.

Based on the results of measurements of capacitance variation with voltage made using the former method, one can expect that the junction capacitance will vary approximately as:

$$C \propto \frac{K}{V_{\text{Band Gap}} - (V_{\text{Forward}} + V_{\text{Degeneracy}})^{1/2}}$$

so that the somewhat smaller junction capacitance in the negative



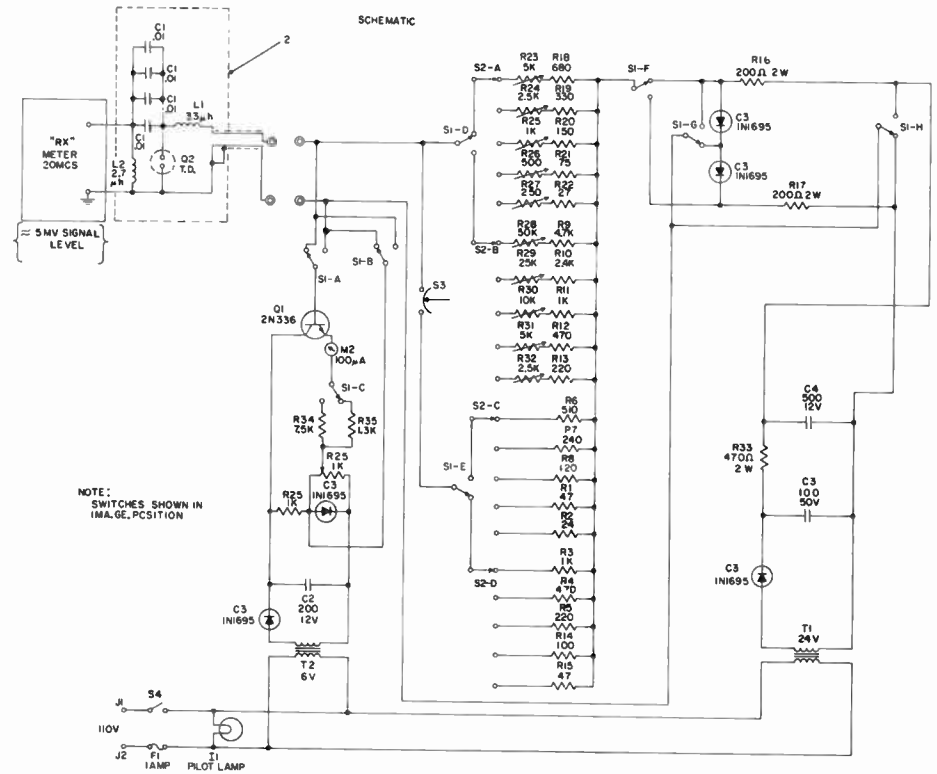
OPERATING REGION AND LOAD LINE
FIGURE 7.11

conductance region may be easily calculated. (The case capacitance may be separately measured and subtracted from the measured quantity.)

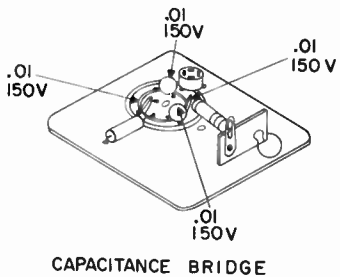
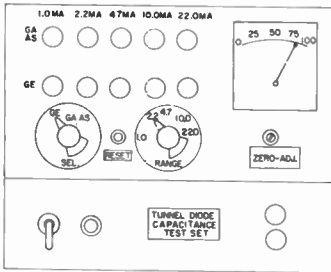
It cannot be over-emphasized that the measuring signal across the tunnel diode must be kept to less than 10 millivolts to prevent errors due to the non-linearity of the diode characteristics.

The RF choke (L_1) is used to decouple the bias network from the bridge. Since the tunnel diode is switching, R_3 is used to supply a current somewhat greater than the peak point current for the particular range of units being measured, R_1 need only be $> 1/|-g_d|$, but is usually made several times this value to prevent oscillation near the valley region. R_2 is used to adjust the intercept point on the device characteristics so that the bridge measures a parallel positive resistance in the order of 1K ohm or greater. (The bridge lead inductance will still reduce the measured capacity somewhat by Lg_d , but if the parallel resistance is 1K ohm or larger it would require $1\mu h$ of lead inductance to cause an error of $1\mu\mu f$ in the reading. If care is taken in keeping lead lengths to less than one inch, the resultant error will be less than $0.2\mu\mu f$ and can be neglected.

The supply voltage (V_1) is usually in the order of 0.75v for germanium and 1.5v for gallium arsenide and is supplied by one silicon diode or two in series as required. Figure 7.12 shows a complete test circuit suitable for tunnel diodes in the range of 1 to 25 ma peak currents.



CR-3	RECTIFIER	IN1695	R21	RESISTOR	75Ω	T2	TRANS	5V. STANCOR P-6469
R1	RESISTOR	47Ω	R22	RESISTOR	27Ω	L1	CHOKE	33µHY
R2	RESISTOR	24Ω	R23	RESISTOR	5KΩ	L2	CHOKE	2.7µHY
R3	RESISTOR	1KΩ	R24	RESISTOR	2.5KΩ	S1	SWITCH	GE-GLAS. SWITCH 2 POS. B CIR.
R4	RESISTOR	470Ω	R25	RESISTOR	1KΩ	S2	SWITCH	12-RANGE SWITCH 5 POS. 4 CIR.
R5	RESISTOR	220Ω	R26	RESISTOR	100Ω	S3	SWITCH	N.O. PUSH BUTTON RESET 5W
R6	RESISTOR	510Ω	R27	RESISTOR	250Ω	F1	FUSE	1AMP
R7	RESISTOR	120Ω	R28	RESISTOR	50KΩ	S4	SWITCH	SPST. TOGGLE POSITIVE ACTION
R8	RESISTOR	240Ω	R29	RESISTOR	25KΩ	J1	TIP JACK	BRUNNBACK CAT # 6280
R9	RESISTOR	4.7KΩ	R30	RESISTOR	10KΩ	J2	TIP JACK	E.F. JOHNSON CAT # 105-520
R10	RESISTOR	2.4KΩ	R31	RESISTOR	5KΩ	M1	DC METER	E.F. JOHNSON CAT # 103-520
R11	RESISTOR	1KΩ	R32	RESISTOR	2.5KΩ	M2	METER	100µA
R12	RESISTOR	470Ω	R33	RESISTOR	470Ω-2W	Q1	TRANSISTOR	4UD4C, 2N336
R13	RESISTOR	220Ω	R34	RESISTOR	75KΩ	Q2	TEST SOCKET	FOR TUNNEL DIODE
R14	RESISTOR	100Ω	R35	RESISTOR	1.3KΩ	I1	LAMP	NE-51
R15	RESISTOR	47Ω	R36	RESISTOR	1.3KΩ	XF1	FUSE HOLDER	BUSSMAN TYPE HJM-CC
R16	RESISTOR	200Ω-2W	C1	CAPACITOR	0.01µFD			
R17	RESISTOR	200Ω-2W	C2	CAPACITOR	2.00µFD-12V			
R18	RESISTOR	680Ω	C3	CAPACITOR	100µFD-50V			
R19	RESISTOR	350Ω	C4	CAPACITOR	500µFD-12V			
R20	RESISTOR	150Ω	T1	TRANS	24 V. STANCOR P-6469			



CAPACITY TEST SET CIRCUIT DIAGRAM
FIGURE 7.12

L_2 is used to offset the bridge if capacitances larger than $20 \mu\mu\text{f}$ are to be measured ($0.27 \mu\text{h}$ will offset $\approx 20 \mu\mu\text{f}$ at the measuring frequency of 20 mc). Note: When the bias is adjusted so that the parallel resistance measured is much greater than $1-5 \text{ Kohms}$, the measuring signal will tend to switch the device, or the bridge readings will tend to drift, because of the relatively large excursions of intercept voltage with signal, and "averaged" values of R and C will prevent a sharp null.

Figure 7.13 shows a photograph of the test set up.



PHOTOGRAPH OF CAPACITY TEST SET UP
FIGURE 7.13

7.4 R_s Test Set

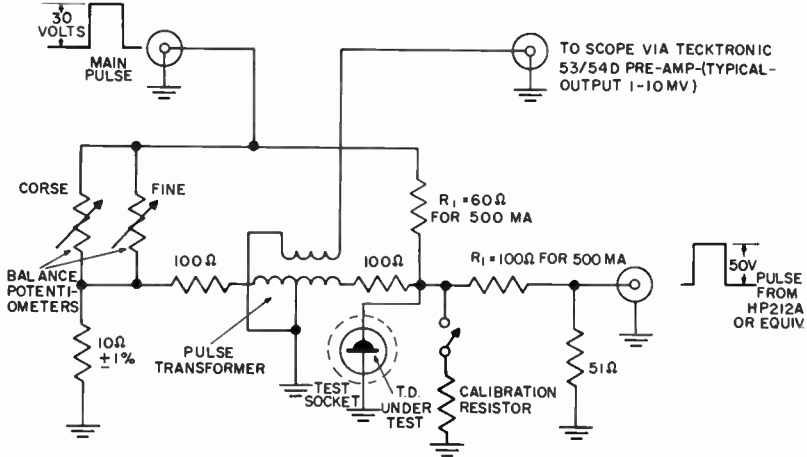
The "Series Resistance" (R_s) is the total internal resistance of the device including contact and lead resistance, as well as the ohmic bulk resistance of the semiconductor material.

In order to obtain a true R_s measurement, it is necessary to measure the incremental slope resistance of the device in its "ohmic" region. This region exists in the high current reverse direction (at least ten times I_p).

The simplest test method would be to apply a large steady state reverse DC current and then measure the incremental slope of R_s by applying a small AC signal. Tunnel diode junction areas are very small however, and the excessive heating which will result at large currents is potentially capable of damaging the units. As a result, a low duty cycle pulse technique was found preferable and this system adopted in the following test set.

A $40 \mu\text{sec}$. high current pulse at 60 c/s (0.24% duty cycle) is applied to bias the tunnel diode in the reverse direction. A shorter, lower amplitude pulse is superimposed on top of the main pulse; the latter acting as a pedestal, the former as an "interrogating" or "sampling" pulse.

These pulses are fed into a bridge network containing the tunnel diode as per Figure 7.14. The series R_1 and R_2 in this figure are chosen for a main pulse-to-incremental pulse ratio of 10:1. Since the main pulse feeds both arms of the bridge, the balance potentiometers are used to balance this main pulse out. The output appearing across the pulse transformer is caused by the incremental pulse only. By use of accurate calibration resistors, the pulse output in millivolts can be calibrated directly into ohms for R_s .



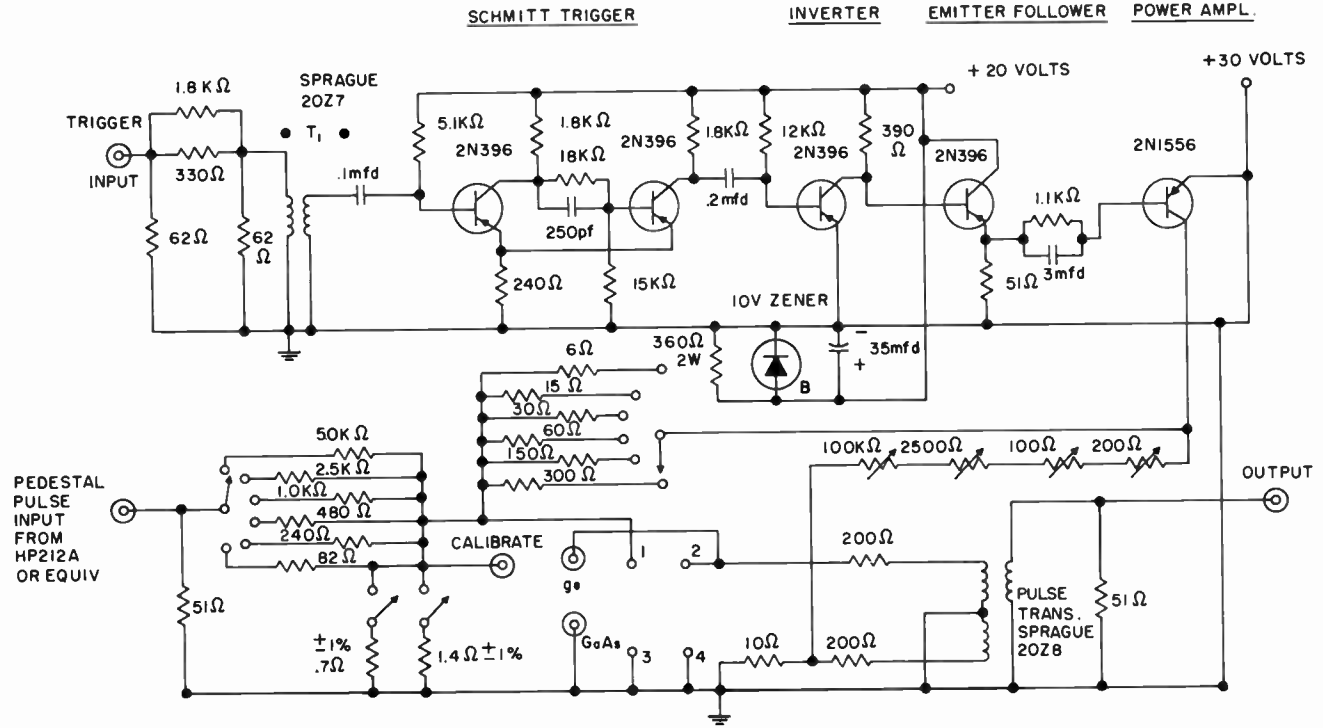
R_s TEST BRIDGE
FIGURE 7.14

Figure 7.15 is the diagram of this test set.

Test Set Up and Calibration Procedure

The pulse generator used in this "test set up" is a Hewlett-Packard 212A or equivalent. The oscilloscope must be rather sensitive and a Tektronix scope with a 53/54D Pre-Amplifier was used, set up for "External Positive Synchronization". The power supply is set for a 30 volt output. The amplitude of the pulse from the 212A generator should be set to 50 volts when connected to the test set.

- 1) Place a jumper from the "Calibrate" jack to the Ge (germ.) jack beside the tunnel diode socket. With the tunnel diode removed from its socket, switch one of the calibration resistors into the circuit.
- 2) Select the desired current test level. Germanium devices can be tested at $100 \times I_p$ to give the most accurate indication of R_s . Because of higher current density and possibly also because of greater power dissipation resulting from a larger voltage at any given current, gallium arsenide devices should only be tested at $20 \times I_p$.



R_s TEST SET DIAGRAM
FIGURE 7.15

- 3) Switch in one of the calibration resistors and balance the bridge so that the only voltage drop across the bridge at balance is that of the $10 \mu\text{sec}$. "interrogation" pulse, using the scope as visual indicator.

This output, divided by the value of calibration resistance used, will now give a calibration in millivolt per ohm.

Operating Procedure

Following calibration, disconnect the jumper and insert a clip lead to make a connection to the diode cap (TO-18 package) in the jack corresponding to the type of device being tested, Ge or GaAs. Next insert the tunnel diode, making sure that the tab for PN devices (such as all General Electric germanium tunnel diodes) point to the left. (Pin 3 of device into socket pin #1). With the tunnel diode in its socket, make the clip lead connection to the cap.

After balancing of the bridge, read the output amplitude of the interrogation pulse.

All that remains is to compute R_s from the calibrated millivolt/ohm reading.

7.5 Measurement of the Device Inductance (L_s)

A precise measurement of " L_s " is a rather difficult measurement to make. The magnitude of this parameter for TO-18 packaged devices is of the order of 1 - 10 nanohenries depending on the method of connection made to the device. For microwave packages (stripline or cartridge types) the device inductance will be an order of magnitude smaller.

The most necessary item of a L_s test circuit would hence be an extremely low inductance adapter for the unit under test. Also, since the magnitude of this inductance to be tested is so small, a very elaborate test set must be used. The test set used here was the General Radio Transfer-Function and Immittance Bridge, Type 1607-A.

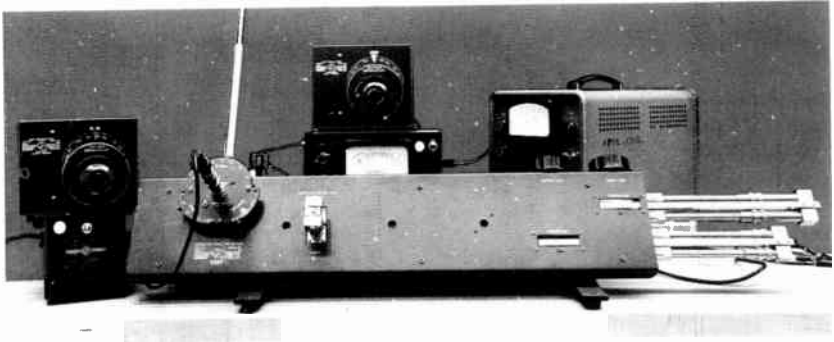
Performing an impedance test at 500 Mc/s, on a "squared off U" piece of wire about 1/4" long, one can bring the plane of measurement slightly above the socket of the transistor mount 1607-P101. By recording the admittance term with the wire protruding 1/8" from the socket, the reference plane is established at this distance from the socket.

Now one can substitute an internally shorted tunnel diode case with the lead length one intends to use plus about 1/4". By subtracting the first reading from the second one, one has established the inductance of the device above the reference plane.

Another method to determine L_s is to measure the tunnel diode impedance at a low frequency when biased in the valley region. This establishes the capacitive reactance since the inductive reactance is negligible at low frequencies. Another impedance measurement at a higher frequency ($\geq 500 \text{ Mc/s}$) where the inductive reactance is appreciable will enable the equivalent inductance to be calculated from

the measured impedance and previously determined capacitance.

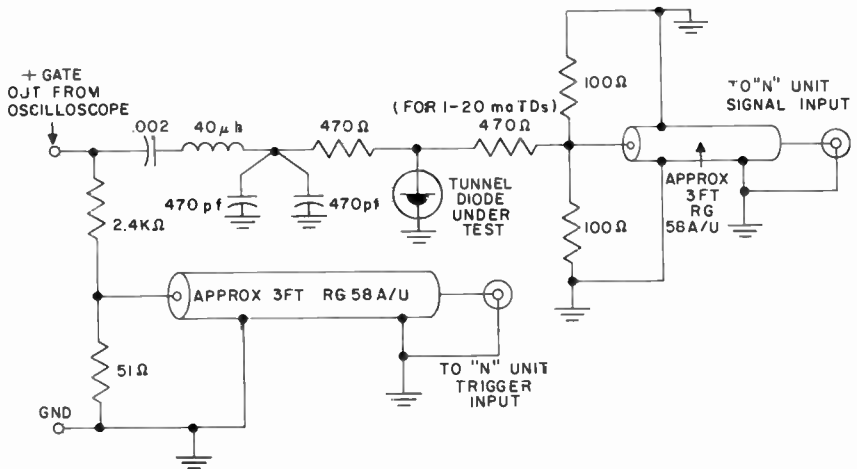
For UHF or microwave application, even for some lower frequency application where L_S must be minimized, a test adapter should be built. General Radio's "Experimenter" brochure (Vol. 34, Nos. 7 & 8, July-August, 1960) or their Reprint #E109-October, 1960 has an outline drawing of such an adapter for TO-18 packaged devices. These two brochures also contain measurement procedures for most tunnel diode parameters, using the aforementioned Type 1607-A bridge.



PHOTOGRAPH OF L_S TEST SET UP
FIGURE 7.16

7.6 Risetime Measurement

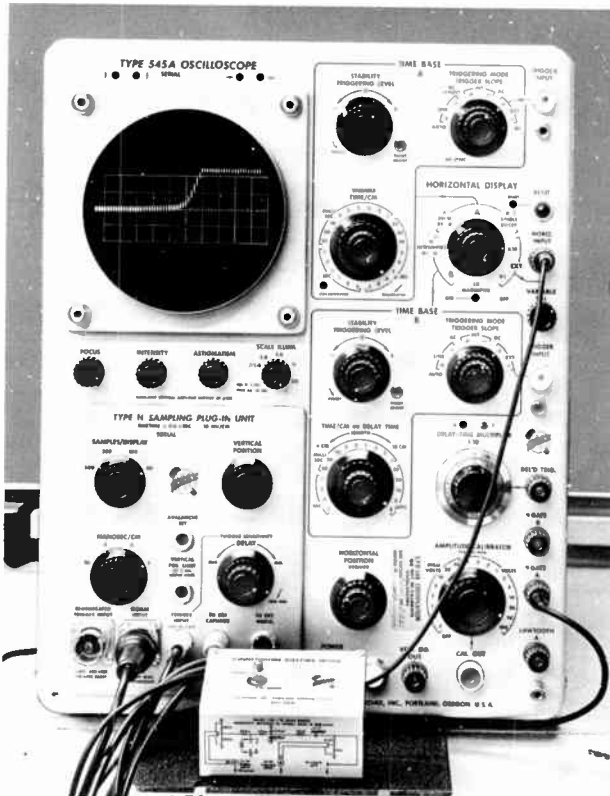
A convenient method of testing switching speed (risetime) is shown in Figure 7.17.



RISETIME TEST SET CIRCUIT DIAGRAM
FIGURE 7.17

A Tektronix plug-in oscilloscope provide both the current ramp source for the tunnel diode and the pre-trigger for the Type "N" sampling scope unit. The use of this sampling scope (or other means of presenting extremely fast speed information) is essential since the risetime of tunnel diodes is much too fast for conventional oscilloscopes.

The Type "N" unit is set up conventionally while the oscilloscope main sweep generator is allowed to "free run" at $1 \mu\text{sec}/\text{cm}$. The "+ GATE OUT" triggers the "N" unit and also provides a current ramp with a low rate of rise which allows the tunnel diode to switch at essentially its own rate. The rise time of the scope is in the order of 0.6 nanoseconds and should be accounted for in the measurement. Figure 7.18 is a photograph of the test set up displaying a 1.4 nanosecond risetime pulse. This corresponds to an actual circuit risetime of 1 nanosecond since the total risetime is the square root of the squares of the scope and the circuit risetimes.



PHOTOGRAPH OF RISE TIME TEST SET UP

FIGURE 7.18

The unit shown in Figure 7.18 is commercially advertised by Tektronix under their part #013-029.

TUNNEL DIODE SPECIFICATIONS

Definition of Terms

Tunnel and Back Diode characteristics are defined with reference to the static characteristic curves of Figure 1 and Figure 2 respectively.

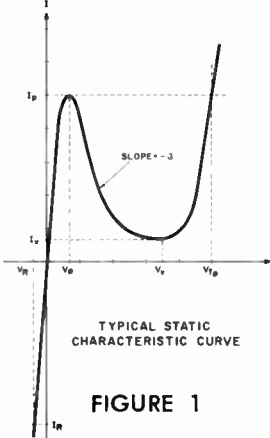


FIGURE 1

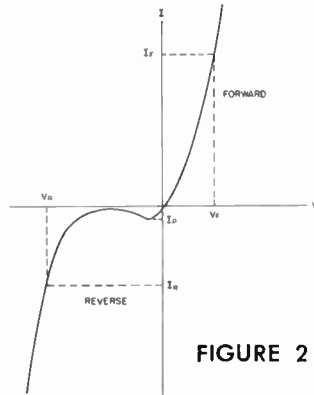


FIGURE 2

Terminology

Symbol

Definition

Peak Point

The peak point is that point on the forward characteristic of a tunnel diode corresponding to the lowest positive voltage at which $d_I/d_V = 0$.

Reverse Peak Point

The peak point is that point on the reverse characteristic of a tunnel diode corresponding to the lowest reverse voltage at which $d_I/d_V = 0$.

Valley Point

The valley point is that point on the forward characteristic which corresponds to the second lower positive voltage at which $d_I/d_V = 0$.

Peak Point Current

I_P

The current flowing at the peak point.

Reverse Peak Point Current

I_P

The current flowing at reverse peak point.

Peak Point Voltage

V_P

The voltage at which the peak point occurs.

Valley Point Current

I_V

The current flowing at the valley point.

<u>Terminology</u>	<u>Symbol</u>	<u>Definition</u>
Valley Point Voltage	V_V	The voltage at the valley point.
Peak-to-Valley Point Current Ratio	I_P/I_V	The ratio of the peak point current to the valley point current.
Forward Peak Point Current Voltage	V_{fp}	The voltage corresponding to that point on the forward characteristic where the current is equal to the maximum specified peak point current.
Forward Voltage	V_F	The voltage corresponding to a point on the forward characteristic at a specified current.
Forward Current	I_F	The current flowing in the first quadrant-conducting region.
Reverse Voltage	V_R	The voltage corresponding to a point on the reverse characteristic curve at a specified current.
Reverse Current	I_R	The current flowing when operated in the third quadrant.
Series Inductance	L_S	The total series equivalent inductance.
Series Resistance	R_S	The total series equivalent resistance.
Capacity	C	The barrier capacity of the intrinsic diode.
Negative Conductance	$-G$	The negative conductance of the intrinsic diode.
Resistance Cut-off Frequency	f_{ro}	The frequency at which the real part of the diode admittance measured at its terminals goes to zero. The resistance cut-off frequency is given by: $f_{ro} = \frac{g}{2\pi C} \sqrt{\frac{1}{gR_S} - 1}$
Self-Resonant Frequency	f_{xo}	The frequency at which the imaginary part of the diode admittance goes to zero. The self-resonant frequency is given by: $f_{xo} = \frac{1}{2\pi} \sqrt{\frac{1}{L_S C} - \frac{g^2}{C}}$
Pico-farad	pf	Micro-micro farad ($\mu\mu f$) i. e. 10^{-12} farads.

<u>Terminology</u>	<u>Symbol</u>	<u>Definition</u>
Nano-henry	nh	Milli-micro henry ($m\mu h$) i. e. 10^{-9} henries.
Nano-second	nsec.	Milli-micro second ($m\mu sec.$) i. e. 10^{-9} seconds.

The "Tunnel" Diode is a new semiconductor device which makes use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristic and very high frequency performance. Its static characteristics are shown in Figure 3.

A "Back" Diode is essentially a "tunnel" diode operated in the reverse direction. For example, you take a "tunnel" diode characteristic curve (Figure 3) and flip it over interchanging quadrants 3 and 4 which results in a "back" diode characteristic curve (Figure 4). The polarity is changed on the back diode since it is operated in the same mode as a conventional diode whose characteristic curve has the conducting region (forward voltage) in the first quadrant and blocking region (reverse voltage) in the third quadrant.

The tunnel diode is designed with emphasis on peak point control and the negative resistance region. The "back" diode is designed with emphasis on forward voltage and its control. In most applications the reverse peak point current of a "back" diode would correspond to a leakage current in a conventional diode. Naturally, the smaller the leakage, the better. The peak point current on back diodes is usually much lower than the peak point current on tunnel diodes. Some "back diodes" are low peak current tunnel diodes and as such do not offer the fine control on conducting voltage that is desired. General Electric Back Diodes are designed for specific diode applications. In order to cover a wide range of applications General Electric offers several types, each of which has minimum and maximum forward voltages for a specific current level of operation.

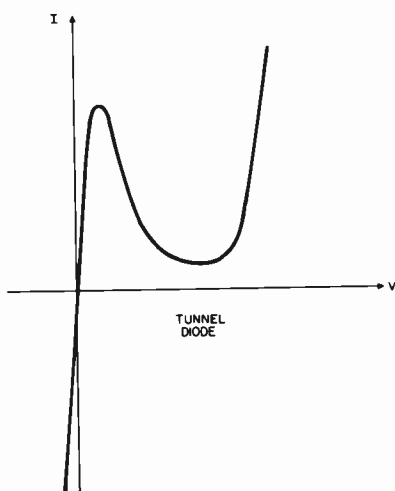


FIGURE 3

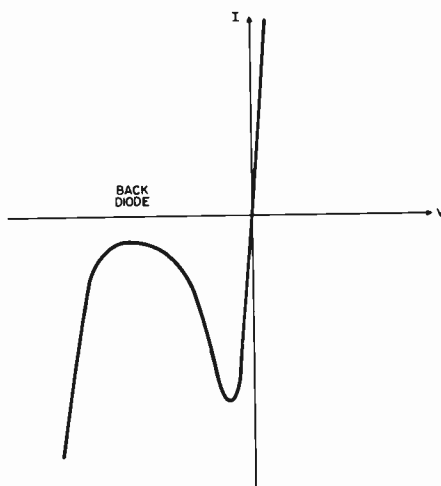


FIGURE 4

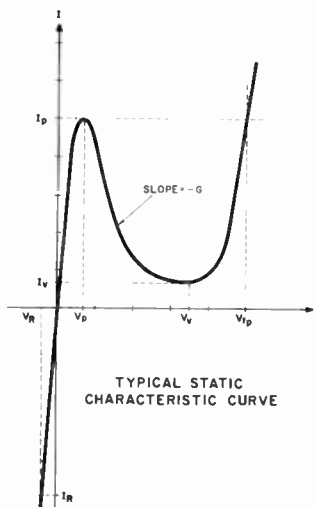
TUNNEL DIODE SPECIFICATIONS

The 1N2939 and 1N2939A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance.

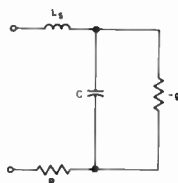
These devices are designed for low level switching and small signal applications with frequency capabilities up to 2.2 Kmc. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.

1N2939, 1N2939A

Outline Drawing No. 1



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (-55 to +100°C)

5 ma

Reverse (-55 to +100°C)

10 ma

Temperature

Storage

T_{STG}

Operating Junction

-55 to +100 °C

Lead Temperature $\frac{1}{8}$ " $\pm \frac{1}{32}$ "

-55 to +100 °C

From Case for 10 seconds

T_L

260 °C

ELECTRICAL CHARACTERISTICS: (25°C) ($\frac{1}{8}$ " Leads)

	Min.	Typ.	Max.	
Peak Point Current 1N2939	I_p	0.9	1.0	1.1 ma
Peak Point Current 1N2939A	I_p	0.975	1.0	1.025 ma
Valley Point Current	I_v		0.10	0.14 ma
Peak Point Voltage 1N2939A	V_p	50	60	65 mv
Valley Point Voltage	V_v		350	350 mv
Reverse Voltage ($I_R = 1.0$ ma)	V_R			30 mv
Forward Peak Point Current Voltage 1N2939	V_{fp}	450	500	600 mv
Forward Peak Point Current Voltage 1N2939A	V_{fp}	450	500	550 mv
Peak Point Current to Valley Point Current Ratio	I_p/I_v		10	
Negative Conductance	$-G$		6.6×10^{-3}	mho
Total Capacity 1N2939	C	5.0	15	15 pf
Total Capacity 1N2939A	C	4.0	10	10 pf
Series Inductance	L_s^*		6	6 nh
Series Resistance	R_s	1.5	4.0	4.0 ohm

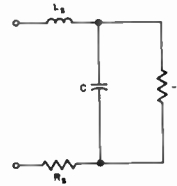
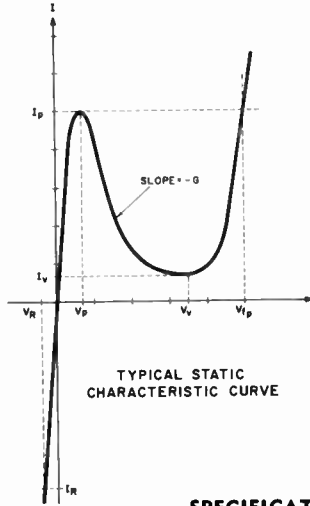
*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

1N2940, 1N2940A

Outline Drawing No. 1

The 1N2940 and 1N2940A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance.

These devices are designed for low level switching and small signal applications with frequency capabilities up to 2.2 Kmc. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (-55 to +100°C)	5	ma
Reverse (-55 to +100°C)	10	ma

Temperature

Storage	T _{STG}	-55 to +100	°C
Operating Junction	T _J	-55 to +100	°C
Lead Temperature $\frac{1}{16}$ " $\pm \frac{1}{32}$ " From Case for 10 seconds	T _L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) ($\frac{1}{8}$ " Leads)

		Min.	Typ.	Max.	
Peak Point Current 1N2940	I _p	0.9	1.0	1.1	ma
Peak Point Current 1N2940A	I _p	0.975	1.0	1.025	ma
Valley Point Current	I _v		0.13	0.22	ma
Peak Point Voltage 1N2940A	V _p	50	60	65	mv
Valley Point Voltage	V _v		350		mv
Reverse Voltage (I _R = 1.0 ma)	V _R			30	mv
Forward Peak Point Current Voltage 1N2940	V _{f_p}	450	500	600	mv
Forward Peak Point Current Voltage 1N2940A	V _{f_p}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I _p /I _v		8		
Negative Conductance	-G		6.6 × 10 ⁻³		mho
Total Capacity 1N2940	C		5.0	10	pf
Total Capacity 1N2940A	C		4.0	7	pf
Series Inductance	L _s *		6		nh
Series Resistance	R _s		1.5	4.0	ohm

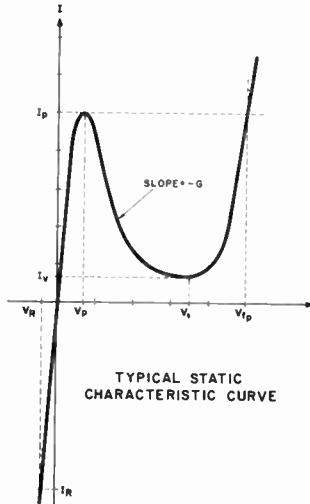
*Inductance will vary 1-12 nh (10⁻⁹ henries) depending on lead length.

1N2941, 1N2941A

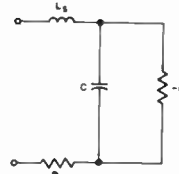
Outline Drawing No. 1

The 1N2941 and 1N2941A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance.

These devices are designed for low level switching and small signal applications. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (-55 to +100°C)	25	ma
Reverse (-55 to +100°C)	50	ma

Temperature

Storage	T _{NTG}	-55 to +100	°C
Operating Junction	T _J	-55 to +100	°C
Lead Temperature 1/8" ± 1/32"	T _L	260	°C
From Case for 10 seconds			

ELECTRICAL CHARACTERISTICS: (25°C) (1/8" Leads)

		Min.	Typ.	Max.	
Peak Point Current 1N2941	I _p	4.2	4.7	5.2	ma
Peak Point Current 1N2941A	I _p	4.58	4.7	4.82	ma
Valley Point Current	I _v		0.6	1.04	ma
Peak Point Voltage 1N2941A	V _p	50	60	65	mv
Valley Point Voltage	V _v		350		mv
Reverse Voltage (I _R = 4.7 ma)	V _R			30	mv
Forward Peak Point Current Voltage 1N2941	V _{FP}	450	500	600	mv
Forward Peak Point Current Voltage 1N2941A	V _{FP}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I _p /I _v		8		
Negative Conductance	-G		30 × 10 ⁻³		mho
Total Capacity 1N2941	C		15	50	pf
Total Capacity 1N2941A	C		10	30	pf
Series Inductance	L _s *		6		nh
Series Resistance	R _s		0.5	2.0	ohm

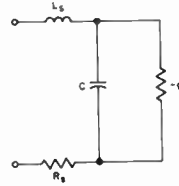
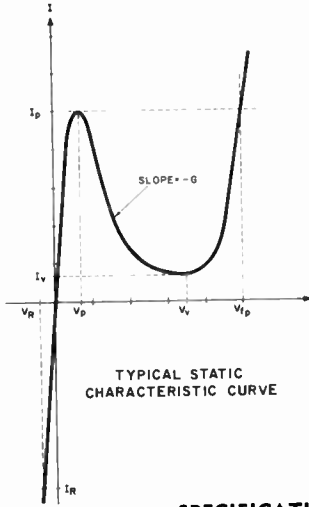
*Inductance will vary 1-12 nh (10⁻⁹ henries) depending on lead length.

1N2969, 1N2969A

Outline Drawing No. 1

The 1N2969 and 1N2969A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance.

These devices are designed for low level switching and small signal applications with frequency capabilities up to 2.5 Kmc. They feature a closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (-55 to +100°C)	10	ma
Reverse (-55 to +100°C)	20	ma

Temperature

Storage	T _{STG}	-55 to +100	°C
Operating Junction	T _J	-55 to +100	°C
Lead Temperature $\frac{1}{8}$ " $\pm \frac{1}{32}$ " From Case for 10 seconds	T _L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) ($\frac{1}{8}$ " Leads)

		Min.	Typ.	Max.	
Peak Point Current 1N2969	I _p	2.0	2.2	2.4	ma
Peak Point Current 1N2969A	I _p	2.145	2.2	2.255	ma
Valley Point Current	I _v		.29	.48	ma
Peak Point Voltage 1N2969A	V _p	50	60	65	mv
Valley Point Voltage	V _v		350		mv
Reverse Voltage (I _R = 2.2 ma)	V _R			30	mv
Forward Peak Point Current Voltage 1N2969	V _{FP}	450	500	600	mv
Forward Peak Point Current Voltage 1N2969A	V _{FP}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I _p /I _v		8		
Negative Conductance	-G		16 × 10 ⁻³		mho
Total Capacity 1N2969	C		8	25	pf
Total Capacity 1N2969A	C		6	15	pf
Series Inductance	L _s *		6		nh
Series Resistance	R _s		1.0	3.0	ohm

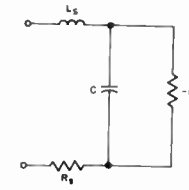
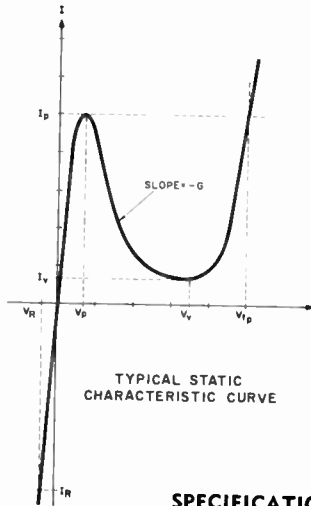
*Inductance will vary 1-12 nh (10⁻⁹ henries) depending on lead length.

The 1N3149 and 1N3149A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance.

These devices are designed for low level switching and small signal applications with frequency capabilities up to 1.5 Kmc. They feature closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.

1N3149, 1N3149A

Outline Drawing No. 1



EQUIVALENT CIRCUIT
(BIASED IN NEGATIVE
CONDUCTANCE REGION)

TYPICAL STATIC
CHARACTERISTIC CURVE

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (-55 to +100°C)	
Reverse (-55 to +100°C)	50 ma
	50 ma

Temperature

Storage	T _{STG}		
Operating Junction	T _J	-55 to +100	°C
Lead Temperature 1/8" ± 1/32"		-55 to +100	°C
From Case for 10 seconds	T _L	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8" Leads)

		Min.	Typ.	Max.	
Peak Point Current 1N3149	I _p	9.0	10.0		ma
Peak Point Current 1N3149A	I _p	9.75	10.0	10.25	ma
Valley Point Current	I _v		1.3	2.2	ma
Peak Point Voltage 1N3149A	V _p	50	60	65	mv
Valley Point Voltage	V _v		350		mv
Reverse Voltage (I _k = 10 ma)	V _r			30	mv
Forward Peak Point Current Voltage 1N3149	V _{f_p}	450	500	600	mv
Forward Peak Point Current Voltage 1N3149A	V _{f_p}	450	500	550	mv
Peak Point Current to Valley Point Current Ratio	I _p /I _v		8		
Negative Conductance	-G		60 × 10 ⁻³		mho
Total Capacity 1N3149	C		30	90	pf
Total Capacity 1N3149A	C		25	50	pf
Series Inductance	L _s *		6		nh
Series Resistance	R _s		.25	1.5	ohm

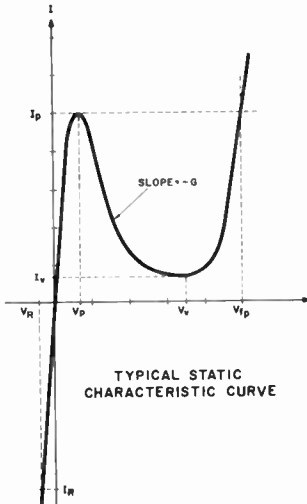
*Inductance will vary 1-12 nh (10⁻⁹ henries) depending on lead length.

1N3150

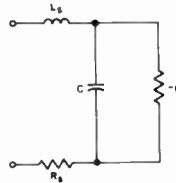
Outline Drawing No. 1

The 1N3150 is a germanium tunnel diode which makes use of the quantum mechanical tunneling phenomenon thereby attaining a unique negative conductance characteristic and very high frequency performance. This

device is designed for low level switching and small signal applications with frequency capabilities up to 1.3 Kmc. It features closely controlled peak point current, good temperature stability and extreme resistance to nuclear radiation.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS: (25°C)

Current

Forward (-55 to +100°C)	100	ma
Reverse (-55 to +100°C)	100	ma

Temperature

Storage	TSTG	-55 to +100	°C
Operating Junction	TJ	-55 to +100	°C
Lead Temperature $1/8'' \pm 1/32''$ From Case for 10 seconds	TL	260	°C

ELECTRICAL CHARACTERISTICS: (25°C) (1/8" Leads)

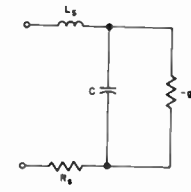
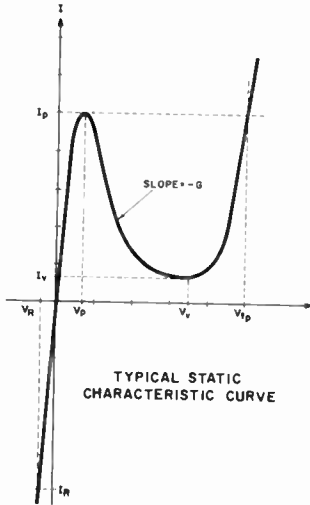
		Min.	Typ.	Max.	
Peak Point Current	Ip	20	22	24	ma
Valley Point Current	Iv		2.9	4.80	ma
Peak Point Voltage	Vp		60		mv
Valley Point Voltage	Vv		350		mv
Reverse Voltage (IR = 22 ma)	Vr			30	mv
Forward Peak Point Current Voltage	Vrfp	450	500	600	mv
Peak Point Current to Valley Point Current Ratio	Ip/Iv		8		
Negative Conductance	-G		100×10^{-3}		mho
Total Capacity	C		60	150	pf
Series Inductance	Ls*		6		nh
Series Resistance	Rs		.15	1.0	ohm

*Inductance will vary 1-12 nh (10^{-9} henries) depending on lead length.

1N3218, 1N3218A

Outline Drawing No. 3

The 1N3218 and 1N3218A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristics and very high frequency performance. These small stripline type packages are designed for microwave communications, radar, very high frequency amplifiers and oscillator applications. The very low series inductance plus controlled low capacity permits very high frequency performance in the S band.



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS

Current

Forward (-55 to +100°C)

Reverse (-55 to +100°C)

5 ma
10 ma

Temperature

Storage

-55 to +100 °C

ELECTRICAL CHARACTERISTICS: (25°C)

		Min.	Typ.	Max.	
Peak Point Current	I_P	0.9	1.0	1.1	ma
Valley Point Current	I_V		0.13	0.22	ma
Peak Point Voltage	V_P		60		mv
Valley Point Voltage	V_V		350		mv
Forward Voltage ($I_F = I_P = 1$ ma)	V_{FP}	400	500	600	mv
Reverse Voltage ($I_R = 1$ ma)	V_R		20		mv
Negative Conductance	-G	5			
Series Resistance	R_S		9	12	$\times 10^{-3}$ mho
Series Inductance	L_S		1.5	4.0	ohm
Total Capacity 1N3218	C		0.3	0.5	nh
Total Capacity 1N3218A	C	1.5	7	10	pf
			4	5	pf

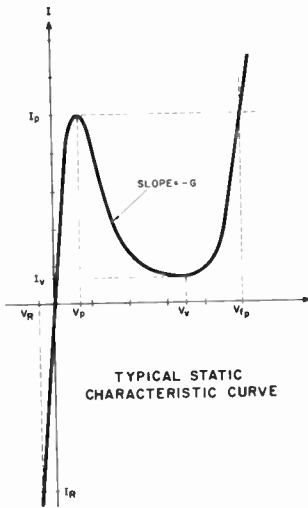
Color Code: (counter clockwise on flange)
 1N3218 Orange-Red-Brown-Gray-Black
 1N3218A Orange-Red-Brown-Gray-Brown

1N3219, 1N3219A

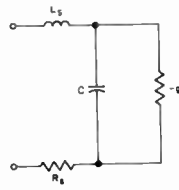
Outline Drawing No. 3

The 1N3219 and 1N3219A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristics and very high frequency performance.

These small stripline type packages are designed for microwave communications, radar, very high frequency amplifiers and oscillator applications. The very low series inductance plus controlled low capacity permits very high frequency performance in the S band.



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS

Current

Forward (-55 to +100°C)	5	ma
Reverse (-55 to +100°C)	10	ma

Temperature

Storage	-55 to +100	°C
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ELECTRICAL CHARACTERISTICS: (25°C)

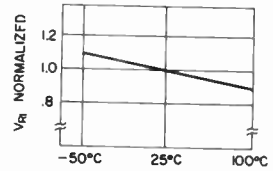
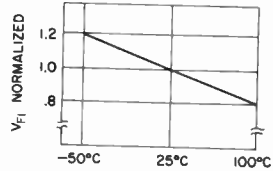
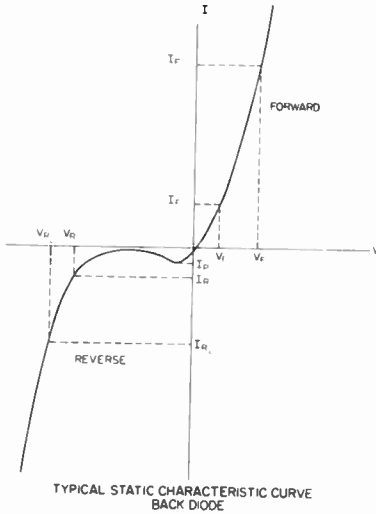
		Min.	Typ.	Max.	
Peak Point Current	I_P	2.0	2.2	2.4	ma
Valley Point Current	I_V		0.28	0.48	ma
Peak Point Voltage	V_P		60		mv
Valley Point Voltage	V_V		350		mv
Forward Voltage ($I_F = I_P = 2.2$ ma)	V_{FP}	450	500	600	mv
Reverse Voltage ($I_R = 2.2$ ma)	V_R		20		mv
Negative Conductance	-G	10	18	25	$\times 10^{-3}$ mho
Series Resistance	R_S		0.7	3.0	ohm
Series Inductance	L_S		0.3	0.5	nh
Total Capacity 1N3219	C	1.5	14	20	pf
Total Capacity 1N3219A	C	1.5	7	10	pf

Color Code: (counter clockwise on flange)
 1N3219 Orange-Red-Brown-White-Black
 1N3219A Orange-Red-Brown-White-Brown

The General Electric Type 4JF2A is a germanium back diode which makes use of the quantum mechanical tunneling phenomenon thereby attaining very low capacity and low forward voltage drop. This device is designed for high speed computer switching circuits. The very low forward voltage is ideal for use with germanium tunnel diodes. It features stable forward and reverse voltage with temperature.

BACK DIODES—GERMANIUM
4JF2A-1, 2, 3, 4, 5, 6, & 7

Outline Drawing No. 2



VOLTAGES VS. AMBIENT TEMPERATURE

SPECIFICATIONS

ABSOLUTE MAXIMUM RATINGS

Current

	TYPE 4JF2A-1							
	-2	-3	-4	-5	-6	-7		
Forward (-55 to +100°C)	25	15	10	5	5	5	5	ma
Reverse (-55 to +100°C)	10	5	5	5	5	5	5	ma

Temperature

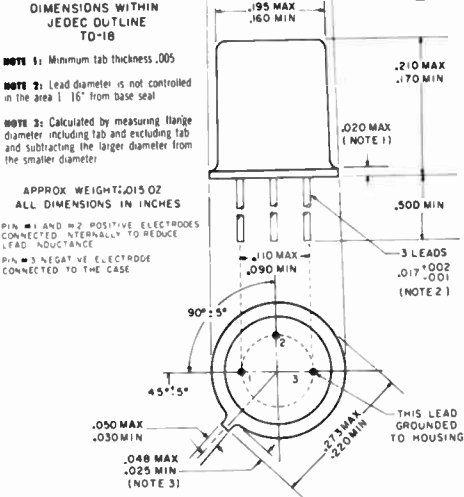
Ambient	-55 to +100		°C
Lead Temperature, $\frac{1}{16}$ " \pm $\frac{1}{32}$ " From case for 10 seconds	260		°C

ELECTRICAL CHARACTERISTICS: (25°C) ($\frac{1}{8}$ " Leads)

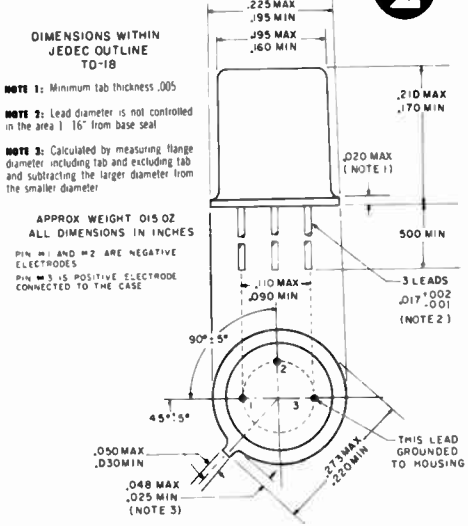
	TYPE 4JF2A-1								
	-2	-3	-4	-5	-6	-7			
Reverse Peak Point Current I_P	1	.5	.2	.1	.05	.02	.01	ma	max.
Reverse Voltage ($I_R = I_P$ max) V_{R1}	440	420	400	380	350	330	300	mv	min.
Reverse Voltage ($I_R = 1$ ma) V_{R2}	440	465	465	465	465	465	465	mv	min.
Forward Voltage at I_{F1} V_{F1}	$\left\{ \begin{array}{l} 80 \\ 100 \\ 10 \end{array} \right.$	$\left\{ \begin{array}{l} 80 \\ 100 \\ 5 \end{array} \right.$	$\left\{ \begin{array}{l} 80 \\ 100 \\ 2 \end{array} \right.$	$\left\{ \begin{array}{l} 80 \\ 100 \\ 1 \end{array} \right.$	$\left\{ \begin{array}{l} 80 \\ 100 \\ .5 \end{array} \right.$	$\left\{ \begin{array}{l} 80 \\ 100 \\ .2 \end{array} \right.$	$\left\{ \begin{array}{l} 80 \\ 100 \\ .1 \end{array} \right.$	mv	min.
								mv	max.
								ma	
Forward Voltage at I_{F2} V_{F2}	$\left\{ \begin{array}{l} 120 \\ 25 \end{array} \right.$	$\left\{ \begin{array}{l} 130 \\ 15 \end{array} \right.$	$\left\{ \begin{array}{l} 170 \\ 10 \end{array} \right.$	$\left\{ \begin{array}{l} 170 \\ 5 \end{array} \right.$	$\left\{ \begin{array}{l} 170 \\ 5 \end{array} \right.$	$\left\{ \begin{array}{l} 160 \\ 2 \end{array} \right.$	$\left\{ \begin{array}{l} 160 \\ 1 \end{array} \right.$	mv	typ.
								mv	
								ma	
Total Capacitance C	$\left\{ \begin{array}{l} 15 \\ 25 \end{array} \right.$	$\left\{ \begin{array}{l} 9 \\ 15 \end{array} \right.$	$\left\{ \begin{array}{l} 6 \\ 15 \end{array} \right.$	$\left\{ \begin{array}{l} 6 \\ 15 \end{array} \right.$	$\left\{ \begin{array}{l} 5 \\ 15 \end{array} \right.$	$\left\{ \begin{array}{l} 5 \\ 15 \end{array} \right.$	$\left\{ \begin{array}{l} 5 \\ 15 \end{array} \right.$	pf	typ.
pf								max.	
Series Inductance L_s	6	6	6	6	6	6	6	nh	typ.

Outline Drawings

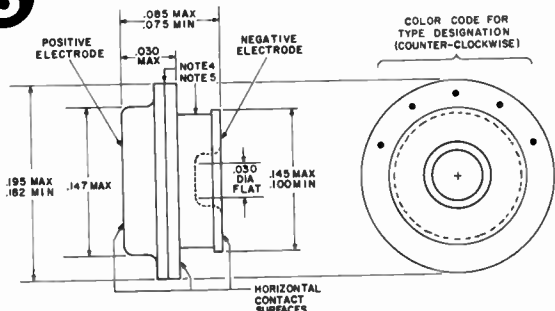
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2



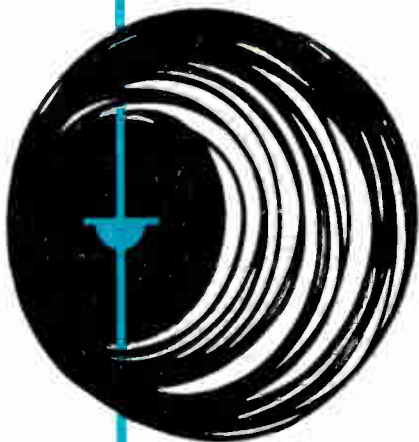
3



- NOTES**
1. ALL DIMENSIONS IN INCHES.
 2. DIMENSIONS ARE REFERENCES UNLESS TOLERANCED.
 3. CONTACT SURFACES TINNED 0005.
 4. WELD FLASH ALLOWED (THIS IS NOT A CONTACT SURFACE).
 5. INSULATION—DO NOT APPLY CLAMPS.

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THIRD EDITION

NINTH PRINTING, 1963

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WETHERSFIELD 9, CONN.

— P R E F A C E —

A non-technical monograph prepared especially for the enlightenment of the high fidelity listener in the interest of providing a better understanding of FM reception for music appreciation. The fundamental considerations for improving and obtaining optimum results with an FM tuner are carefully reviewed for the various conditions of reception. A general knowledge of this important but neglected subject is to be regarded as conducive to the attainment of greater satisfaction in the realistic reproduction of FM broadcasts.

L. F. B. C.

GETTING THE MOST FROM YOUR FM TUNER

It goes without saying that most hi-fi enthusiasts and serious music lovers exercise pre-eminent care and consideration in selecting their equipment and records in order to be assured of the finest in reproduction; yet practically everyone, with the exception of local and metropolitan listeners located near FM stations, fail to realize the importance of employing a proper antenna system.

Today many of us can enjoy excellent musical programs as broadcast from numerous "good music stations". More and more of the FM stations are improving their repertoire to provide a finer diversification of music appreciation. One need no longer depend upon a few local stations as has been the practice in the past when FM was limited to reception of just several stations. Indeed, why should you limit the capabilities of your tuner to local listening, when the installation of a good FM antenna would enable you to listen to many other good music stations if your tuner were given a chance to really perform as it should?

WHY AN FM ANTENNA?

Some music lovers are fortunate to be located within easy reception distance of, say, WQXR-FM or other comparable good music stations, and as a rule many listeners who are within 50 miles of an FM station usually experience little trouble with normal reception. However, there are more and more hi-fi installations being made beyond the primary listening zone, in remote areas now being referred to as the fringe perimeters. It is in such locations that it is especially important that every bit (microvolt) of signal shall be collected and delivered to the tuner in order to receive a station. The degree of listenability resulting is principally the function of three basic factors; (a) signal strength, (b) tuner sensitivity, and (c) an adequate FM antenna. Invariably, *it follows that regardless of the type of tuner used, a good high gain antenna will improve the existing reception.* Bear in mind, however, that a tuner cannot yield what it does not receive in the first place. Even a weak signal, if hardly perceptible at all, may not be capable of properly operating the limiter of the finest tuner made. The logical solution is to deliver a greater input signal strength to the tuner. This service is essentially what a good antenna system should contribute.

WHAT CONSTITUTES A GOOD FM ANTENNA

Ordinarily the usual FM antenna takes the form of a five feet length of ribbon wire ("T" formed) placed under the rug and connected to the tuner input terminals. Sometimes the antenna is mounted outside, such as on the roof-top, and consists of a simple aluminum configuration not unlike those used for television. These forms are usually of the elementary type, known as a folded dipole, Fig. 1, and are considered adequate only for local reception. Inasmuch as the reception of local stations therefore presents no difficult problem, let us concern ourselves with FM reception over greater distances and, in particular, fringe reception over distances in excess of 60 miles. Increasing the distance coverage will automatically provide a greater diversification of programs resulting from the additional FM stations that you will

be able to pick-up. Thus, a high gain antenna offers greater collective capacity for delivering a much stronger signal than any elementary type. This special propensity is a requisite for distant reception and, indeed, it will reveal the latent merits of the tuner in realizing its full sensitivity. So astonishing can be this improvement, that the tuner will appear to be outperforming itself in the seemingly unusual ability to pick-up stations never thought to be within listening range before.

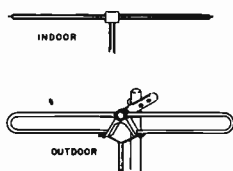


Fig. 1

DESIRABLE CHARACTERISTICS OF FM ANTENNAE

Like any other component comprising your high fidelity equipment, an antenna may have inherent design characteristics that emphasize either desirable or unfavorable features. In evaluating the merits of an antenna, therefore, due considerations should be given to each of the following:

Frequency Coverage

Most antennae are designed to provide a nominal coverage of the FM band, i.e., 88 to 108 megacycles. Like an amplifier or speaker, the degree of coverage over the inclusive frequency range should be normalized to provide good broadband response at a *uniform signal level*. Too many of the so-called broadband antennae now offered are the result of poor engineering design and give their optimum sensitivity in only a portion of the 20 mc. band so that *severe signal attenuation* may result at either upper or lower section of the FM band. Their gain is concentrated at one point and give *peak performance* only over a relatively narrow band. This is particularly true of the ribbon dipole which is resonant to the center of the band and other antennae of poor design as shown in Fig. 2.

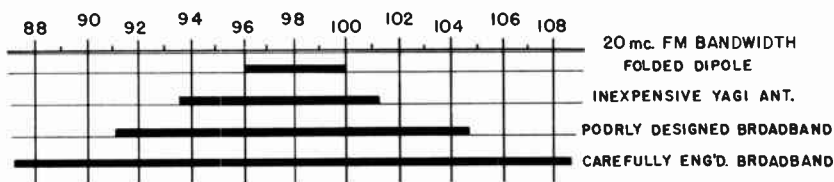


Fig. 2

Other designs usually known as single driven yagi, have a contracted frequency coverage of only four or five megacycles wide. Such an antenna is employed where reception is limited to only one or several stations, all of which have frequencies within the narrow band of a few megacycles, say 88 to 93 or 98 to 103 mc.

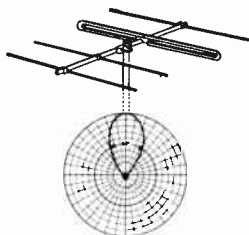


Fig. 3

Usually when only one FM station frequency is desired, a single channel yagi, Fig. 3, may be used in order to obtain optimum gain and greatest sensitivity. For example many music lovers listen only to a favorite station, such as WQXR-FM on 96.3 mc., while employing a crystal controlled or single frequency tuner. For such exacting requirements, we design a single frequency antenna to provide maximum response for the specified frequency only.

Gain and Sensitivity

The performance of an antenna is usually adjudged on the basis of these several characteristics. Among the inherent qualities, gain or sensitivity is indeed a salient merit; the more gain (in db.) it has (up to a certain point) the stronger will be its signal pick-up and yield to the tuner. Obviously, this degree of efficiency at the proper frequency is very important for long distance fringe reception.

The above statement, however, implies a certain qualifying aspect in its true interpretation. Just as automobile manufacturers feature maximum horsepower ratings in establishing the prowess of their engines, some antenna manufacturers do likewise, but the important consideration is in accepting an honest evaluation of *gain for overall performance* and not only for certain "peak" frequencies. This is what some manufacturers offer as grossly exaggerated gain figures used only to deceive the public.

Front-to-back Response

There are a variety of types of antenna designs, some of which are better than others for a specific characteristic, but the one perfect design that affords all of these attributes is yet to evolve in the future.

For greatest diversity of reception, an antenna should receive stations equally well in any direction. This objective is accomplished in one of two manners; (a) with a crossed dipole having omnidirectional signal response or, (b) with a broadband uni-directional antenna controlled with a rotator. See Fig. 4.

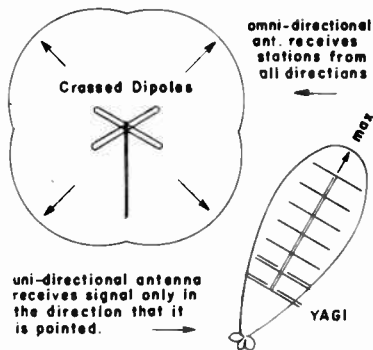


Fig. 4

The former is an elementary type antenna and because of its relatively low gain can serve only for local FM service. However, since signals are received with equal ease in any direction, it is not possible to discriminate between one of two close or adjacent frequencies without some degree of station interference due to the lack of preselection.

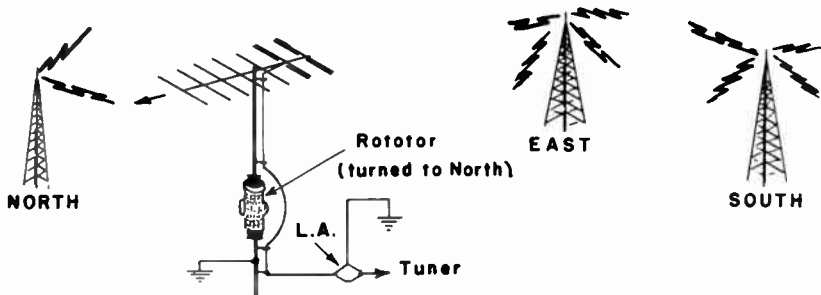


Fig. 5

With a uni-directional antenna and rotator, it is possible to revolve the antenna about a 360° azimuth and because its sense of directivity is very sharp, a station located towards the north can be favored in preference to a station whose location is to the south, notwithstanding the fact that these two stations may occupy adjacent frequencies and be equally distant from the tuner. See Fig. 5. Such an antenna, obviously, has very desirable characteristics



Fig. 6

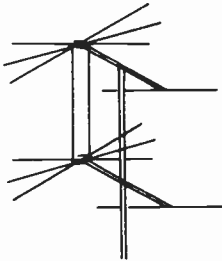
in that it can be beamed like a search light to afford excellent exclusion of unwanted signals and a maximum rejection of pickup from the sides and the rear. Thus the ratio of forward pick-up to that at the rear is expressed as the antenna's "front-to-back" ratio. A high F/B ratio is highly desirable and advantageous in long distance reception and is recommended for any other location where maximum rejection of spurious or noise signals is desired for the suppression of interference. Fig. 6 illustrates this characteristic of the yagi antenna.

CHOOSING THE CORRECT ANTENNA FOR YOUR INSTALLATION

Tuner manufacturers usually furnish a folded dipole, Fig. 1, constructed of ribbon wire in "T" form, which serves as a simple antenna of the most elementary type. In terms of gain, F/B, etc., it leaves much to be desired although it is satisfactory for metropolitan service. Considered as a reference standard for comparison, its gain is 0 db.

Its equivalent in aluminum as constructed for outdoor installation may offer some slight degree of improvement and only if mounted at higher roof-top elevations. The pick-up response pattern is bi-directional as a figure 8 and is varied only when produced as a *crossed dipole*. Fig. 4, in which form it evolves as a broad cloverleaf configuration. Gain is still 0 db.

While singularly popular for television reception, some TV antennae of the conical design, Fig. 7, have also been adapted for FM. Fundamentally, the conical is a very broadband type of antenna and because it covers the VHF TV band 2 to 13, it will also serve for the FM frequencies which, of course, fall in between channels 6 and 7 of the TV band. The conical is better adapted for TV than FM because its gain, F/B and response pattern do not afford the desired performance demand by the more closely spaced FM stations. Its F/B is poor, propensity for noise pick-up is regarded as especially bad for FM, and there is considerable wind resistance due to the large frontal area presented in its structural configuration. Gain is 3 db.



CONICAL ARRAY

Fig. 7

THE YAGI ANTENNA

For optimum performance where reception is difficult, either because of long distance, co-frequency or noise interference, the yagi design has no practical equivalent. A good rugged yagi of superior engineering design can provide more of the desirable attributes that are sought for in an FM antenna than all others combined.

Essentially, the basic yagi consists of a folded dipole, a director (forward) and a reflector at the rear of the dipole as shown in Fig. 8. For the FM band, the standard yagi would normally consist of five elements mounted upon a supporting boom. The folded dipole is the terminal element to which the lead-in or transmission line is attached. The reflector at the rear serves as a shield to reject unwanted signal pick-up from the back while the three director elements mounted ahead of the dipole serve as collectors to focus the signal that is directed to the dipole. Gain is 5 to 7 db. with an average frequency coverage of 4 to 6 mc. bandwidth.

As the basic yagi will cover an average 5 mc. band width, a total of four such antenna would necessarily be required to provide full frequency coverage of the complete FM band of 20 megacycles. Since each yagi would have a specific range of its own, such as 88-94, 94-99, 99-104, 104-108 mc., the

installation would necessitate a four bay array if superposed upon a common mast. Such yagi types are desirable only where a small antenna is to be used and where reception is confined to that portion of the band covered by the particular antenna.

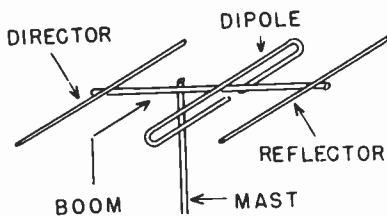


Fig. 8

THE BROADBAND MULTI-ELEMENT YAGI

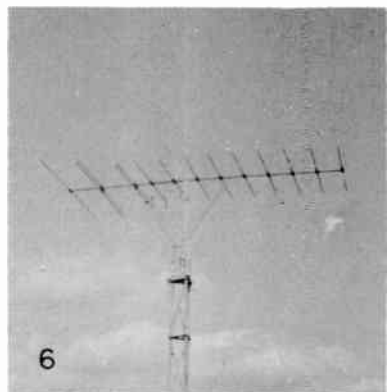
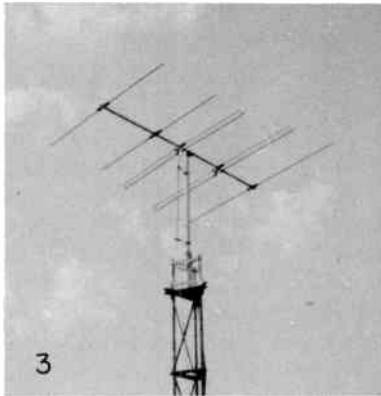
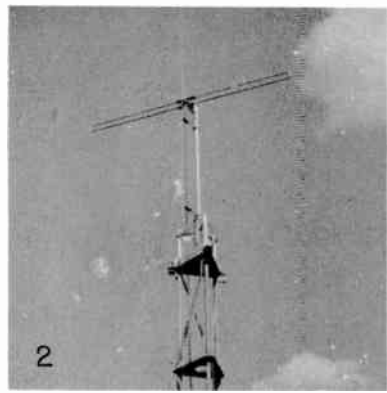
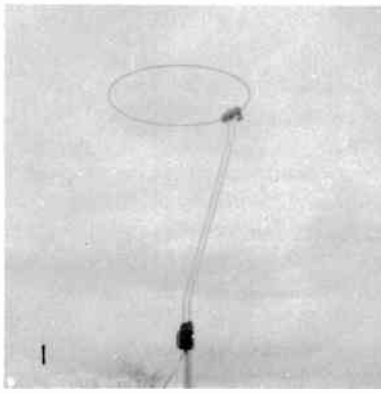
A more ideal antenna to employ is the broadband yagi which is designed for an extended frequency bandwidth. This development is a hybrid version of the basic type and has appeared in various forms, generally distinguished by the use of two dipoles and many more additional director elements. A broadband yagi can be designed to cover the complete FM band without impairing any of the special virtues of the basic type. Some accomplish this feat efficiently and others do not. Given the prime engineering parameters, the solution of the problem is fundamentally one of determining what is the best element size to employ, their optimum spacing, and exactly how many shall be integrated into the overall design. Many varied configurations will necessarily result. However, by knowingly proportioning the dipoles and elements, one outstanding prototype will evolve that is collectively representative of the optimum criteria that has been sought as the ideal objective. Among the various types of antenna now in use, the "FM/Q" design is without reservation the finest broadband yagi available today.

Cognizant engineers have acclaimed it as indisputably offering the very best of engineering features in both electrical and mechanical development. This antenna was perfected by FM engineers after extensive research and tests with actual conditions of reception in locations presenting the most challenging problems. It is recommended as representative of every desirable attribute that a good FM antenna should possess and is endorsed by all of the prominent tuner designers as unsurpassed for top performance.

ANTENNA FOR FRINGE RECEPTION

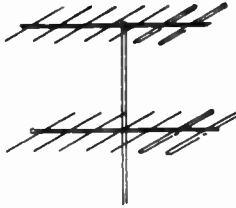
Many tuners today offer excellent sensitivity and are naturally preferred for fringe reception. For distances in excess of 100 miles from an FM station, the tuner should have a sensitivity of at least 3 microvolts for 20 db. of quieting, or better. The less the μv and the more the db. rating, the greater is the sensitivity of the tuner. It follows then, that the antenna should likewise possess equally superior features as it will be necessary to employ a high gain yagi in order to obtain the most signal strength.

In hilly terrain where there is rough topography between tuner and station, a single antenna may not deliver sufficient signal despite the excellence in



1. Alford loop, omni-directional, for primary regional reception.
2. Folded dipole, bi-directional, 0 db. gain, for local reception. (FM/Q Cub)
3. Broadband 5 element yagi, 6 db. gain, for reception to 75 miles. (FM/Q Jr.)
4. Broadband 7 element yagi, 9 db. gain, for reception to 140 miles.
5. Single frequency or narrow band yagi, 12 db. gain, for selective reception or fringe service. (FM/Q LR)
6. Very high gain 12 element broadband yagi for extreme fringe. 14 db. gain single, may be stacked to yield extra 3 db. to produce most powerful FM antenna array available. FM/Q

choice of equipment used. In such locations, the usual resort taken is to "stack" several bays of antennae together as in Fig. 10. Either two (or as many as four) antennae may be interconnected together with a parallel wiring arrangement known as a "harness" to obtain an increase of gain. Each antenna (bay) is superimposed upon a common mast and spaced $\frac{3}{4}$ or $\frac{1}{2}$ wavelength apart for best performance. A common connection of the harness



STACKED YAGI

Fig. 10

(or stacking bars) ensures the correct terminal impedance of 300 ohms for direct attachment to the transmission line and then to the tuner input. Where a single antenna, such as the "FM/Q" Sr. model yields an honest gain of 9 db., a two bay stacked array will deliver an additional gain of from 35 to 41% more signal. Such an installation would then provide the equivalent gain that could otherwise be obtained only by stacking four five element standard type yagi.

Double stacked "FM/Q" yagi will normally fulfill all of the challenging requirements for long distance reception of upwards of 250 miles when employed with a sensitive tuner and an approved antenna system. One should bear in mind that even the best tuner and the best antenna, such as the "FM/Q" Sr., may result in disappointment unless careful attention is also directed to the proper installation of the entire antenna system. The antenna system consists not merely of the antenna itself, but in addition, it includes the transmission line, (lead-in), insulators and method of installation to the tuner. Essential considerations that will determine the ultimate results obtained therefore call for a special regard of the following:

IMPEDANCE MATCHING

All FM tuners have a standard nominal input of 300 ohms corresponding to the equivalent terminal impedance of most antennae and related accessories. A few of the better tuner circuits also have an alternate input of 72 ohms for use with co-axial line. The importance of maintaining compatible impedance relations in the system should not be overlooked as any disparity between these circuits will create an electrical mismatch which results in an attenuation of signal. This condition is analogous to a pipeline made by coupling a $\frac{1}{2}$ inch diameter pipe to a 1 inch size and expecting to obtain the rated water output which the larger size is capable of delivering. While a variation of from 250 to 375 ohms exists in the impedance of many antenna types, the variance is less important here than when introduced in the antenna system between the tuner and the antenna proper. Because it is so easy to lose much of an elusive signal through mismatches after it is delivered by the antenna, it behooves us to pay particular care in installing an antenna system that is to serve for fringe reception.

PERFORMANCE CHARACTERISTICS OF FM/Q YAGI

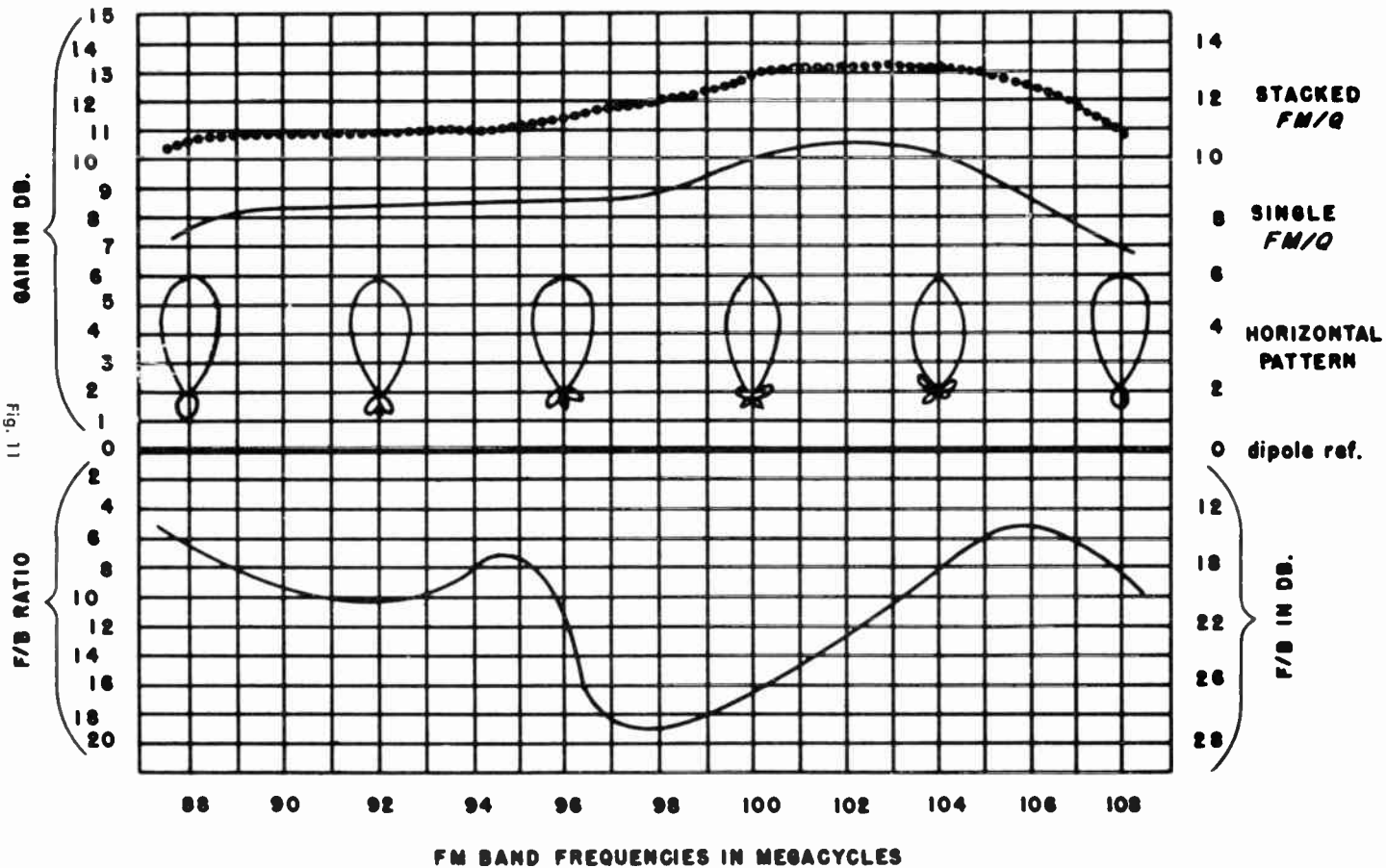


Fig. 11

THE TRANSMISSION LINE

Ordinarily, the common lines such as employed with TV antenna will suffice for most FM installations. Among the numerous types in use, Fig. 12, some offer less losses than others and are preferred because of their special merits, particularly for fringe service where insertion loss must be reduced to a minimum. A reference to the accompanying chart will disclose that there is a great variation according to type. Note also the detrimental losses introduced when a line becomes wet due to rain, snow or ice.

Several noteworthy rules to practice include:

1. Suspend the line over a roof so as to prevent contact with shingle in order to minimize leakage of signal.
2. Whenever possible run the line in a vertical path with minimum horizontal traverse. Noise pick-up is far less in a vertically transported line than in a horizontally strung lead-in.
3. Never coil an unused portion of the line at the tuner.

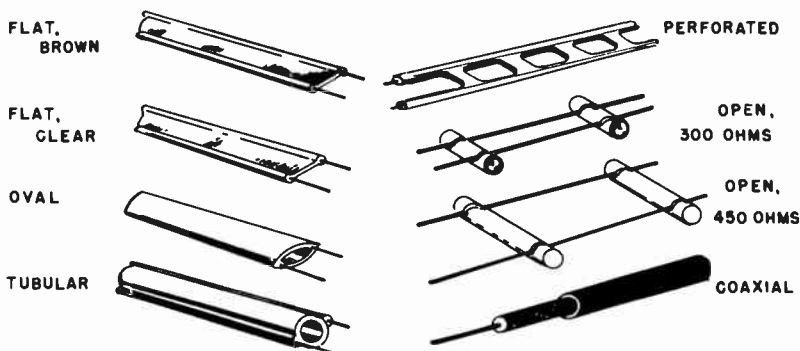


Fig. 12

Standard 300 ohm flat ribbon line (TV type) is available in either clear or dark brown color, the former to be used for indoors and the latter is made especially for outdoor service. These lines employ polyethylene insulation and despite the pigmentation added, their average service life is seldom more than 20 months due to ultra-violet solar depreciation which induces progressive losses in the insulation. Accordingly, it is recommended that poly type lines should be replaced at least every two years to ensure good signal strength at all times. Variations of the standard, such as the perforated and the silicone coated (a film easily applied over the surface) lines are better; however, the lowest losses are to be realized only when open wire line is chosen. For general usage, silicone is highly endorsed as an excellent preservative for the entire antenna system.

INSERTION LOSS OF LINES IN DB. PER 100 FT. AT 100 MC.

<i>Type</i>	<i>Dry</i>	<i>Wet</i>	<i>Impedance</i>	<i>Remarks</i>
Co-axial, RG-59/U	3.7	3.7	72	Shielded cable, small
Co-axial, RG-11/U	2.2	2.2	72	Shielded cable, medium
Co-axial, JT-200S	0.92	0.92	72	Shielded cable, large
shielded twin lead	3.4		300	balanced line
flat ribbon, brown	1.6	7.3	300	polyethylene insulation
oval, air core	1.3	4.8	300	polyethylene insulation
tubular, air core	1.2	2.5	300	polyethylene insulation
flat ribbon, clear	1.1		300	for indoors only
perforated flat	0.7		300	punched ribbon
open wire	0.3		300	5/16 inch spacing
open wire	0.3		450	1 inch spacing

OPEN WIRE LINE

Open wire line has the lowest loss providing the best signal path for either fringe or other installations where the antenna must be located more than 100 feet away from the tuner. As illustrated, this line consists of parallel exposed wires equidistantly separated by a polystyrene insulator. The spacing of the wires determines the line impedance. For lines from 150 to 500 feet, the 300 ohm size is recommended; however, the wider spaced 450 ohm line is preferred for distances of from 500 to 1,000 feet as it is less susceptible to retaining formations of snow and ice. Open wire line is also advised for coastal locations where saline deposition is inescapable and causes serious signal depreciation. Of course, the use of either type of open wire line implies that proper impedance matching shall be provided to tuner and antenna terminations. Fig. 13 shows how this may be accomplished within the line itself. Copperweld wire, which has the equivalent electrical conductivity as the annealed copper conductor, is preferred in any open wire line as it affords greater tensile strength and less expansion, thereby enabling the use of fewer pole supports throughout the suspended length. Open wire line naturally demands a greater care in installation because it is not as flex-

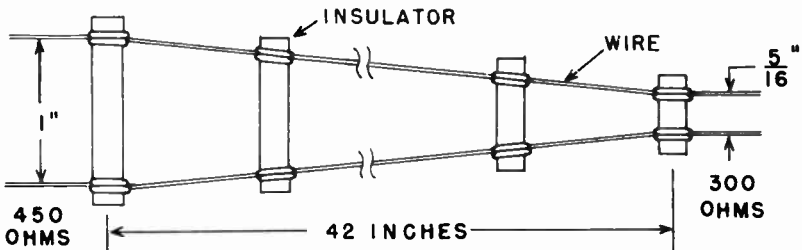


Fig. 13

ible as flat ribbon lead and can easily kink if improperly handled. It is held away from buildings by special all-plastic retainers (or standoff insulators) designed especially for low loss support. When bypassing a rotator, a 'flexoff' insulator, Fig. 14, should be employed to prevent contact and short circuiting the wires to the rotor or mast. This device is helpful with any type of transmission line as it prevents the loss of signal whenever the rotor turns and wraps the line about the mast.

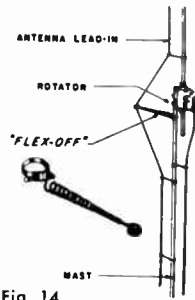


Fig. 14

THE LIGHTNING ARRESTOR AND GROUND

Inasmuch as the antenna is normally erected in an unobstructed area, usually at roof top elevation, the supporting mast must be grounded for protection against lightning. It is recommended that No. 6 or 8 aluminum wire shall be employed in effecting a good contact connection between the mast and the earth.

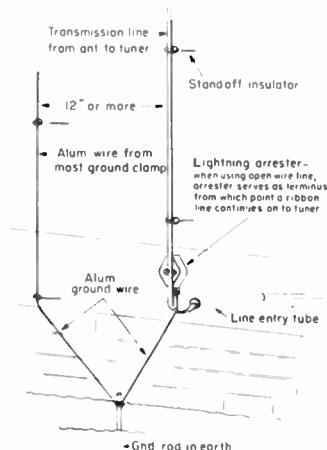


Fig. 15

are concentrated within the antenna system and must be dissipated and neutralized with the ground potential. A lightning arrester accomplishes this by by-passing the charge to the earth before it continues on to the tuner circuit where it would eventually result in component breakdown.

THE TRUTH ABOUT CHIMNEYS

While the chimney affords a simple and convenient support for mounting an antenna, it is nevertheless the worst possible location for the FM antenna. Such installations are a condemnable practice and should not be permitted in the interest perpetuating good reception.

An antenna should never be mounted upon an active chimney. The smoke and fumes discharged are detrimental in the manner that they attack the insulators, line and the very metal itself to induce rapid deterioration of the complete assembly. It is significant to recall that resistances are basically of carbon composition and smoke is essentially carbonaceous. To apply a resistance coating to the antenna, particularly the insulators, is comparable to effecting a high resistance short circuit and by-passing the signal to ground.

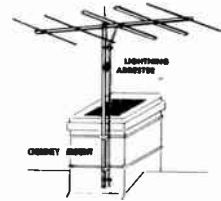


Fig. 16

ATTIC ANTENNA INSTALLATIONS

Some listeners prefer to install the antenna in the attic of their home. While such an installation will give satisfactory results for primary and regional service, it is not recommended for fringe reception. As compared to the usual outdoor mounting, the attic antenna may operate at a loss of from 10 to 30 percent of its rated efficiency. However, mounting the antenna in the attic simplifies installation and offers desirable protection but there are environmental influences which may seriously affect reception and these should be duly considered before deciding on such an installation.

If the antenna is to be installed in the attic, a stationary mounting is easily made by suspending the yagi in free space using a strong string (do not use wire) and positioning the antenna so that it points in the direction of the principal stations to be received. Where the antenna is to be orientated, the rotator mast may be suspended from the overhead rafter and allowing enough clearance to accommodate the turning radius of the antenna. Avoid placing the antenna near to electrical wiring and never point it towards the standpipe or any other metal object. Homes with metal roofs or where aluminum foil is used for insulation under the shingles should not contain attic installations as the shielding thus provided will seriously impair reception. There are also other influences which can affect the reception when using an attic antenna and these may be observed when the roof is wet (particularly slate roofs) or when covered with snow or with reflective aluminum paint.

ADVANTAGES OF A ROTATOR

A uni-directional antenna operates best when it can be properly orientated and beamed directly to the source of signal broadcast. Any degree of rotation for exact antenna positioning is easily accomplished, with a rotator. The device is mounted on the mast, supports the antenna, and is remotely controlled by the listener at the tuner. Inasmuch as a good FM antenna has a sharp directional pattern, the advan-

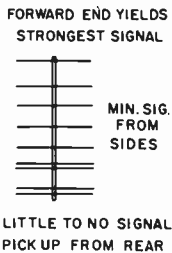


Fig. 17

tage of being able to turn it in any direction enables the listener to not only obtain the best pick up of signal, but also to search for elusive stations. The much improved performance means that this added feature of operational control can be used to augment the tuner in providing for better selectivity. Thus, by careful positioning of the antenna, certain stations may be more easily segregated to the exclusion or satisfactory suppression of others since the antenna receives signals best in its forward plane.

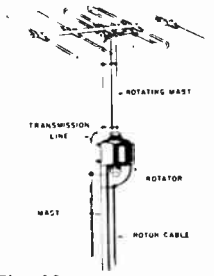


Fig. 18

The inclusion of a rotator should be considered as essential with every high gain broadband installation. This provision is indispensable for fringe reception and also for 72 ohm installations using high gain broadband yagi. With coaxial line, there is no spurious signal pick up in the lead-in and because the antenna will exhibit its true performance characteristics as interceptor, a rotator is necessary for proper orientation.

For FM, the manually controlled models of rotator are preferred to the automatic type. The latter do not index at sufficiently close intervals as required for precise positioning of the antenna and cannot be jogged as may sometimes be necessary. Where stations can be picked up at all points of the compass, the rotator is a definite aid to tuning as it can effectively improve signal strength, reduce adjacent and co-frequency interference, minimize distortion and enhance tuner selectivity.

MOUNTING FM AND TV ANTENNAE

For practical reasons, it is sometimes desirable to install both FM and TV antennae on the same mast. If this type of installation is contemplated, a separation of at least one quarter wavelength (24 to 30 inches) should be allowed for spacing between them in order to prevent electrical interaction. Usually the larger antenna is at the bottom; however, for fringe reception, the FM antenna should be mounted in the most favorable position at the top.

Individual transmission lines for the lead-in should always be installed for the respective antennae. Arrange the lines so that they are on opposite sides of the mast, or to be as far apart as practical and thereafter separate them by increasing the distance to 12 inches or more. Under no condition should these antennae be connected together with a common coupling device to provide for a single lead-in as cross channel interference will result.

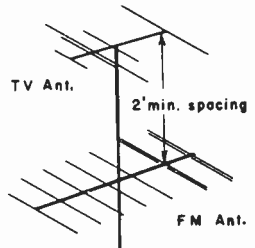


Fig. 19

TV ANTENNA FOR FM?

For the sake of convenience, some listeners would also use their TV antenna for alternate FM reception. Good reception of FM stations cannot be expected when a TV antenna is employed as a compromise for both services.

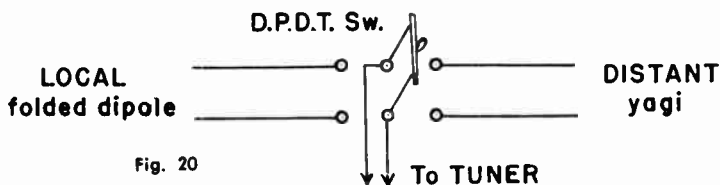
Because TV antennae are especially designed to give the best sound and picture resolution on the specific TV channels, their gain response for the FM band is purposely suppressed to minimize interference. Poor response on FM with a TV antenna is therefore an inherent characteristic, otherwise the FM stations would also be equally well received and there would be a serious source of picture interference as a result. As the FM band is between TV channels 6 and 7, the obvious necessity for a frequency hiatus at this point conforms with good design practice for TV antennae. Accordingly, while the TV antenna may suffice for local FM reception, it will never give the comparable performance delivered by an antenna that is designed especially for FM.

TWO SET COUPLERS

Two set couplers are not recommended as an accommodation for multiple tuner operation unless FM reception from the one available antenna is limited to local stations. It should be noted that the insertion loss as imposed by the use of a coupler is approximately 3 db. This loss is equal to the gain that is obtained by stacking and it is a high sacrifice of signal that can thoughtlessly be lost to such inconsiderate improvisations. If a two set coupler must be employed, first install a high gain antenna that is capable of delivering enough signal to service more than one tuner.

LOCAL-DISTANT RECEPTION

Many urban installations have one or more local stations which provide programs in addition to other more distant sources of broadcast. The installation is best made using a high gain yagi for the distant objectives, notwithstanding the fact that a folded dipole would easily suffice for the local service.



A compromise calls for the use of an antenna transfer switch which is installed behind the tuner to change over from the dipole to yagi according to station demand. The switch eliminates possible overloading of the tuner on local stations were the yagi is to be used alone.

FM RECEPTION IN CITIES

Perhaps the most trying situations in obtaining good reception with FM is to be experienced in cities. Aside from the restrictions on the type of antenna that may be installed, reception may suffer from structural barriers, noise, spurious images and an unfavorable signal-to-noise level. All this is true, notwithstanding that the transmitting antenna may be only several miles away. A solution to these dispiriting problems is not easily offered because the situation is indigenous to the surrounding environment. Unfortunately, FM is influenced by buildings to the extent that they cause an obstruction, reflection or absorption of the signal. This coupled with the fact that in most instances it is only possible to employ an indoor folded dipole is indeed a discouraging matter.

Many apartments are provided with a master TV antenna system and because of restrictions, this outlet is usually the only facility available for FM reception. Such installations will not give satisfactory FM performance as being designed especially for TV service, they include a filter which suppresses the FM frequencies in order to eliminate possible TV picture interference.

The obvious remedy is, of course, to install a roof top antenna. However, where restrictions do not allow its use, the next best alternative is to place the dipole at the window. Too frequently, ribbon dipoles are placed under the rug, along an inside wall or tacked behind a cabinet. These are all poor locations and contribute to a worsening of the condition under the circumstances.

DISTORTION AND MULTIPATH RECEPTION,

FM distortion is a common complaint wherever the antenna, in addition to the direct signal received, also intercepts secondary signals that are being picked up as reflections. This form of distortion is similar to and may be compared with the phenomena of "ghosts" as observed on many TV sets. Because one of these signals is a reflection and therefore arrives slightly later than the direct transmission, a time lag occurs between them that is of sufficient delay to result in distorted reproduction.

The condition is most prevalent in cities where reflected signals abound but may also be encountered in mountainous regions. FM distortion due to multipath reception can be expected with the use of any of the omnidirectional types of antennae such as the folded dipole, crossed dipole and conicals. Their lack of directional pick up, makes them susceptible to spurious signals and when used with conventional ribbon line, such distortion is inevitable.

The uni-directional antenna (yagi), however, offers a solution as its forward response and high order of F/B discrimination may be used advantageously

for directive signal interception while minimizing the pick up of reflections arriving at the sides and rear. Also, the use of such an antenna predicates the utilization of shielded coaxial cable as a necessary provision to prevent the infiltration of secondary signals in the line.

For city reception, a high gain yagi, equipped with an attenuation pad and the coaxial line to tuner will effectively correct the problem provided that the installation is made in an approved manner at roof top elevation or in an area free from immediate obstructions.

THE 72 OHM CO-AXIAL ANTENNA SYSTEM

Several FM tuners of the better type are equipped with an alternate 72 ohm antenna input. Such tuners are particularly desirable for service in locations where an abnormal noise environment exists as created by either automotive ignition or other electrical equipment. Obviously, a sensitive tuner cannot escape noise pickup and be expected to discriminate between good music and man made interference. For relief, any degree of noise suppression is desirable and assuming that the tuner is adequately well shielded, the rest of

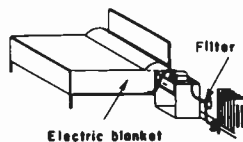


Fig. 21

of the problem is largely dependent upon; (a) type and location of antenna, (b) transmission line employed, (c) nature of noise generator. Some antennae, such as the common dipoles, have poor response patterns that are inherent in their design and afford no directive discrimination for sensitivity. A good yagi, such as the "FM/Q", which has a minimum vertical pick-up and pronounced directivity in the forward plane (like a light beam), will exclude much of the surface pick-up which is where the majority of noises originate. The most troublesome interference is that produced by autos, electrical utility lines and appliances (as in Fig. 21); all of which are usually found near the ground surface level. In other words, *by locating the proper antenna high enough so as to be above the source of noise, much of the interference can be eliminated or effectively reduced.* However, since the transmission line is also a part of the antenna system, the special shielded line known as co-axial cable (type RG-59/U) is employed. Co-axial line will not absorb noise voltages and is ideal for this purpose. Due to the capacitive effect introduced by the close proximity of the braided shield and the center wire conductor, co-axial line has more loss than the unshielded lines. This relatively greater insertion loss of the shielded line results in considerable signal attenuation which explains why (despite the apparent added sensitivity shown by tuner manufacturer's specifications) there is little difference in the relative signal strength obtained when delivered to the tuner input through either a 72 or 300 ohm system.

THE 72 OHM INSTALLATION

The 72 ohm system is recommended as a solution to two prevalent complaints as experienced with installations in city and suburban areas;

- (a) for suppression of noise and interference
- (b) for alleviating signal dominance from a nearby station

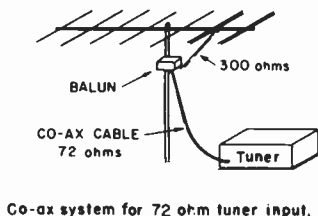


Fig. 22

Its installation may be made with any standard FM antenna while using either of two popular types of co-axial cable, RG-59/U or RG-11/U. For a line length of 100 feet or less, use the former and for lines of greater length use the latter type which has less attenuation loss.

The standard 300 ohm antenna is adapted to this system by employing a balun which is an impedance matching transformer designed to effect a low-loss coupling between 300 and 72 ohms or vice versa. An outdoor type balun is mounted on the mast, below the antenna, and to which the co-axial cable is connected and run to the tuner. If the FM tuner has a 72 ohm input, a direct connection can be made using the proper fitting. Where there is no 72 ohm input, an indoor type balun must be employed to reconvert the cable impedance back to 300 ohms in order to match the standard tuner input. The successful operation of the system is dependent upon the location of the antenna, use of a low-loss balun and good weatherproof cable connections.

Because unshielded transmission line readily picks up a signal as additional to the antenna, too much signal delivery to a tuner can result in overloading. This is an existing problem near to FM stations and the 72 ohm system is therefore desirable in strong signal areas to limit signal interception to the antenna so that the usable signal can be controlled.

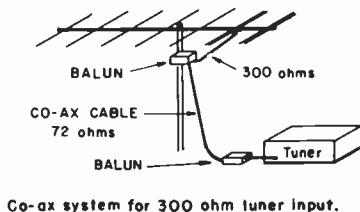


Fig. 23

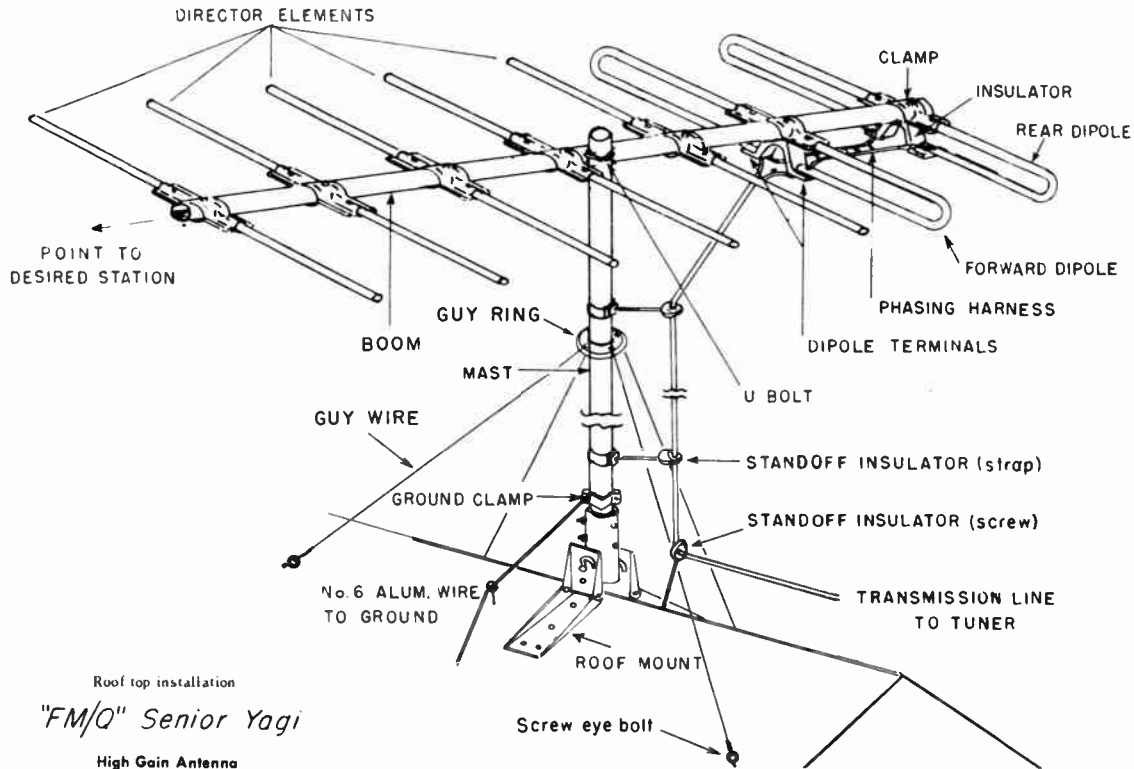
FRINGE AREA RECEPTION

Good music is contagious to many people and we find that there are golden ears located at remote distances striving to catch the elusive signal of a station that is beyond the normal limits of reception.



Fig. 24

To those who would overcome this obstacle of distance, my advice is necessarily tempered by the type of tuner selected or already in use. However, *it is always better to obtain maximum signal strength through the use of stacked yagi* than to resort to other ques-



For an installation where the antenna is to be orientated, a rotator is employed and is mounted on the mast where the guy ring is shown. The supporting guy ring wires are then attached directly to the anchoring lugs as provided on the rotator.

tionable expedients. Any gain that is achieved by the use of stacked antennae has less noise content than by voltage amplification through other alternatives. It is also prescribed that you employ half-wave spacing if at all possible for stacked arrays to realize optimum gain. Open wire line is a must and all the special precautions should be taken to keep losses at a minimum or to eliminate them entirely from the antenna system. Accordingly, avoid the use of standard TV standoff insulators and guard against running the transmission line either close to or in parallel to any other wire or metal surface, not, of course, excluding contact with foliage and roof surfaces.

In some locations, such as hilly or mountainous terrain, the use of a mast or tower to gain elevation cannot be advocated as always advantageous. Often-times, the signal is found to be stronger near the ground than at 60 to 100 feet above.



Fig. 26

This surprising fact is true because the signal wave washes over the mountain (Fig. 24) and spills down along the sides or contours of the mountain. Probing and trial locations will frequently disclose the best spot where the antenna should be placed. Also, because a good yagi antenna possesses a pronounced directivity pattern, it may even be necessary to try tilting the antenna so as to aim it over the ridge of the mountain. It is further suggested that the antenna should be carefully orientated around a full 360 degrees, all the time noting for the best position that gives the maximum signal. The best position indicated may not necessarily point toward the transmitter as it is possible to obtain a very strong signal as a reflection from another mountain as shown in Fig. 26.

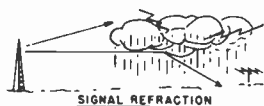


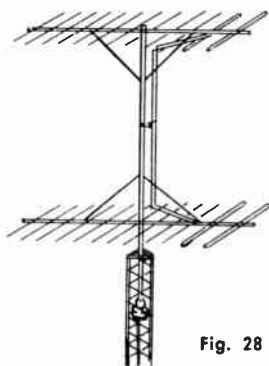
Fig. 27

Last, but not least, a pick-up antenna located on a mountain top can be made to direct or retransmit the intercepted signal down to the antenna proper at the tuner, each of which may be located at varying levels of altitude. At other times, prevailing weather conditions may also influence long distance reception as illustrated in Fig. 27. These are some of the uncertainties of long distance reception and indicate various approaches in the solution of the problem.

VERY LONG DISTANCE RECEPTION

Aside from the important consideration noted in the preceding paragraphs, long distance FM reception implies the use of an antenna system that is designed for exceptionally high gain signal yield. A multi-element yagi, preferably installed as a stacked array and using low-loss line, will assure the most promising performance. It is not unusual, therefore, to anticipate reception over distances of from 200 to 500 miles when such a system is operated in conjunction with a tuner of relatively high sensitivity.

For consistent long distance reception, the considerate selection of a well designed FM tuner that is to be used in conjunction with a high gain stacked array is of equal importance. The tuner should offer a high order of sensitivity, good limiting and adequate IF amplification for a better than average performance. The array should be equipped to permit orientation using a rotator that is capable of supporting the load. A tilt-down type of tower is also highly recommended to facilitate installation, gain elevation and to lower the antenna in the event of impending storms which might cause extensive damage.



Because long distance reception covers a range of considerable territorial extent, special regard should necessarily be allowed for the intervening and perimeter influences which affect propagation. A range that is interspersed with mountains and the variable weather conditions that are associated therewith, can seriously affect the signal delivery to account for fading, pulsing, etc. Experiments disclose that signal stability over extended distances is most favorably controlled when employing a multiplicity of elements with the antennae arranged in a stacked array. These ideals have been realized in special designs as recommended for such application. See page 7.

ANTENNA BOOSTER OR PREAMPLIFIER

Where a stacked antenna fails to bring in sufficient signal, the additional use of an approved FM preamplifier *may*, in some instances, further enhance the objective signal yield. Because FM boosters are of varied design, care should be exercised in appraising the claims made for them. Recently, there have appeared small antenna mounted units of dubious value and which leave much to be desired towards improving FM reception.

The well designed modern tuner, already possessing a high degree of sensitivity, will not benefit from the additional use of an auxiliary signal booster that is of inferior design to the tuner itself. As a result, no specific contribution is to be gained in their usage because in many instances reception may be further impaired due to their unsatisfactory operation in conjunction with an FM tuner. Among the common difficulties experienced are circuit instability resulting in oscillations, stray signal and interference pick up and the generation of spurious images of strong local stations. The latter is particularly troublesome as it causes the station to repeat at many points of the dial.

In general, signal boosters are not recommended for use with tuners of high sensitivity. Older type circuits, lacking good sensitivity, may benefit a little, but the overall signal-to-noise level will usually remain unchanged. Consequently, it is possible for the inexperienced audiophile to invest in the use of a preamplifier and still fail to gain any appreciable improvement in the resulting signal-to-noise level.

LOW SIGNAL INTENSITY AREAS

Occasionally an antenna installation may provoke a great disappointment due to its apparent failure to produce the expected results. Since "shadow spots" are nondescript and because an antenna that is erected within these confines leaves much to be desired in terms of performance, it is usual to discredit the equipment and not the condition that is prevalent. Locations known as low signal intensity areas do exist and may likely be encountered anywhere. In particular, these areas are to be found in canyons, deep valleys and in mountainous terrain with dense foliage. Here poor signal interception can present a problem because the immediate and surrounding topography limits the usual recourse to special provisions that might otherwise be taken to improve the signal deficiency.

A relocation of the antenna from the "shadow" area to a more favorable site offering better signal intensity is about the only practical approach to the problem. A solution suggests placing the antenna on a nearby hilltop and using low-loss open wire line to carry the signal down to the tuner. This in effect is what is accomplished on a commercial scale with community TV service.

SIGNAL FLUTTER DUE TO AIRCRAFT

Many listeners suffer the annoyance of sporadic signal loss coincident with the flight of aircraft traveling enroute between the receiving and transmitting sites. This disturbance is also a form of multipath reception wherein the aircraft diverts the usual path of signal transmission by introducing a moving reflector to distort the normal pattern of wave propagation.

The susceptance to such disturbances may be minimized by the installation of two uni-directional antennae erected at random locations and dispersed approximately ten wave-lengths apart or

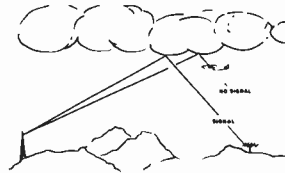


Fig. 29

as to be determined by trial and experiment. Each of the separated yagi is individually terminated to an antenna discriminator using shielded coaxial cable so that the entire system is effectively protected against secondary signal pick up. The output from the discriminator feeds the tuner with a signal voltage that is not below the value as required to maintain the threshold of limiting for a stable signal.

FM WAVE TRAPS

A frequent complaint amongst FM listeners results from a strong adjacent frequency station creating interference which defies suppression. Where this situation exists and cannot be eliminated by either the usual methods of proper antenna orientation or tuner selectivity, a wave trap must be employed.

In most instances it is folly to buy an inexpensive type of trap and hope that it will prove effective. Likewise the use of any of the selective types of signal attenuators as sold for the purpose will not correct the problem. Its proper solution implies the use of a complete 72 ohm system with baluns in conjunction with a well designed high-Q wave trap. In order that the system shall be fully effective, the tuner must also be 100% shielded.

By providing for a completely shielded receiving system, the antenna alone is exposed to intercept the signal. The interfering signal is absorbed by the resonant trap circuit and there being no additional signal reinforcement from spurious pick-up, its cancellation is subject to maximum attenuation. While such a trap may not be as effective as Armstrong's cavity resonator, it will alleviate the most stubborn cases and bring relief.

ANTENNA REQUIREMENTS FOR MULTIPLEX

The new multiplex system utilizes a composite signal (*representing the left channel*) consisting of the main FM signal (*which is the sum of both channels*) and an AM 38 kc. sub-carrier which conveys the difference (*representing the right channel*) between the two signals to provide true stereo reproduction. As *the full composite signal is transmitted as an FM signal*, segregation of the respective FM main carrier and the AM sub-carrier channels results after detection in the FM tuner or adapter. This implies that distortion may occur with some tuners (and locations) as a result of; (a) poor tuner design having inadequate IF bandwidth and discriminator linearity for low distortion reception of the sub-carrier, (b) insufficient signal voltage yield from antenna to tuner.

The type and adequacy of the receiving antenna employed now assumes a new importance for good FM multiplex reception. With added demands being placed upon the tuner conversion operation, *a much stronger signal voltage is required as the range of reception is substantially reduced from 30 to 50% of the normal coverage*. Thus while the range of usual mono signal reception may be 50 miles, a good tuner will now only have a range of 35 miles and a poorly designed tuner may be limited to reception within 25 miles of the station. Because the signal voltage level to tuner input is so essential to good stereo resolution, *a high gain broadband FM antenna is a must* if distortion is to be avoided. This means that ribbon dipoles, turnstile folded dipoles and the inadequate TV antenna (see page 15) are to be discarded for the more efficient FM yagi which is capable of delivering a maximum signal to tuner. In many instances, the indoor antenna is totally ineffective and a roof top installation will be required in order to obtain sufficient signal improvement, or gain, to escape the distortion that is to be attributed to this inadequacy.

FM ANTENNA SHOULD COMPLIMENT THE TUNER

So many audiophiles are prone to overlook the importance of equipping their FM tuner with an efficient antenna system. While the selection of a tuner and the attention devoted to specifications receives careful scrutiny worthy of a professional piece of equipment, most people are content to operate their system with a wholly inadequate antenna. This important oversight defeats the original intent and special advantages that might have been realized as a result of their considerate investment. Therefore, if you select a tuner, believing it to be the best for superior performance, by all means also install a suitable antenna that is complimentary to its design. Otherwise how can these objectives be fully realized? In failing to make this necessary provision, the tuner operates under a handicap because you will be limiting its potential capability by making it subservient to an inefficient antenna. Don't make the mistake of putting the plow before the horse.

If the antenna that is selected is adjudged to be both adequate and efficient, then make certain that it is also properly installed so that it may deliver the most favorable signal to ensure optimum tuner performance. A haphazard installation that is characterized by hasty and careless workmanship, while taking short cuts with unapproved practices, can only result in a progressive loss of signal. An efficient antenna, properly installed, is an asset to any hi-fi installation.

ANTENNA INSTALLATION ACCESSORIES

An outdoor installation is considerably simplified if suitable accessories as designed for use with the antenna system are employed. Below are listed various items of common usage and their application to the job. Their provision, as required, will ensure a low-loss, permanent and safe installation. See Fig. 25.

ANTENNA INSTALLATION KIT — Contains all of the essential accessories needed to make installation of the antenna for an average roof top mounting.

ANTENNA MAST GROUNDING CLAMP — Provides a connection for attaching ground wire to the mast as necessary for grounding the antenna against lightning.

ANTENNA SWITCH — A transfer switch designed to permit the selection of one of several available antennae.

ATTENUATION PAD — An electrical device used to reduce the signal strength from strong local stations to a value that will not overload the tuner.

COAXIAL CONNECTOR — An electrical connector used with shielded cables in a 72 ohm system for terminating the line to balun or wave trap.

GUY RING — Attached to mast to provide anchoring and support for the antenna. When a rotator is employed, guy wires are attached to rotator to dispense with guy ring.

GROUND ROD—The terminal stake that is driven into the earth and to which the ground wire from the mast (and lightning arrestor) is attached. Usually 4 to 6 ft. long.

GUY WIRE—Aluminum stranded wire attached to guy ring and anchored to roof to support the antenna permanently in place.

GUY WIRE THIMBLES—Small "U" bolt clamping devices, two of which are used for securing each of the guy wires.

LIGHTING ARRESTOR—A protective device used in the antenna system to dissipate electro-static discharge.

LINE ENTRY TUBE—An insulated lead-in sleeve used to admit the line into the building with a minimum loss of signal.

MAST—The tubular standpipe upon which the antenna is mounted in order to gain clearance and elevation in free space. Heavy duty aluminum tubing of from 1¼ to 1½ inch diameter is preferred as it is non-rusting, light weight and strong. A mast that is 4 to 12 ft. high is usually employed for roof installations.

ROOF MOUNT—The mast is held in position with a receptacle known as the roof mount. For end of the building installations, an eave or wall mast bracket is needed to support the mast. Specify aluminum for roof mounts and hot dip galvanized steel for wall type brackets.

ROTATOR THRUST BEARING—Multiple antenna installations and stacked arrays designed for rotation, should be supported with a thrust bearing to relieve the load imposed on the rotator mechanism.

STAND-OFF INSULATORS—To support the transmission line, use strap type along the mast or tower. Change to screw type to support the line along sides of a building using one at intervals of every 4 to 8 ft. or as necessary.

TRANSMISSION LINE—The lead-in carrying the signal from antenna to tuner should be of good low loss polyethylene and preferably made with 18 gauge stranded wire conductors. Perforated ribbon line is highly recommended as best for general use. Twisting the line one turn per linear ft. sometimes helps to minimize noise pick-up.

FINAL RECOMMENDATIONS

Recommendations for selecting the correct antenna for your FM tuner should be considered on the same order as when evaluating any other component of the high fidelity system. In the foregoing review, the principal aspects of what constitutes a good FM antenna have been carefully pointed out in an endeavor to give you the significant features that contribute to an attainment of FM reception at its best. With regard to specific brand, I will singularly specify the "FM/Q" yagi for the following outstanding features;

1. Unsurpassed performance with optimum gain, exceptionally high F/B ratio, reflecting superior engineering design throughout.

2. Has permanent mortise type of element attachment to boom. Only FM antenna made assuring rigid assembly as protection against the formation of high resistance grounds which cause weak signals, loss of sensitivity, intermittent noise and disabled elements. Notice that it *does not* use "quick-rig or snap-lock" type of construction, principal cause of the above failures in 90% of antennae.
3. Has recursive dipole in a continuous length assuring the lowest possible electrical resistance and uni-potential conductivity for permanent high-gain sensitivity. Mechanical connections will not loosen up to result in progressive signal attenuation.
4. Only FM antenna using stainless steel bolts, nuts and washers. Offers finest protection against rust corrosion and mechanical breakdown.
5. Employs durable oversize heavy-duty aluminum seamless elements for maximum strength and weather protection from wind, ice and storms.
6. Insulator made of non hygroscopic dielectric material for low-loss efficiency. Minimizes attenuation of signal due to moisture-path leakage.
7. Designed by renowned FM antenna engineer for optimum tuner performance over the entire FM spectrum. The "FM/Q" is not a Chinese copy of an existing prototype nor is it made up of incidental lengths of tubing thrown together for a bargain.
8. Only FM antenna made that approaches perfection.

BIBLIOGRAPHY

For additional collateral research, interested readers are referred to a special article on "Antenna Systems for FM", written by the author of this book and published in the July 1956 issue of "AUDIOCRAFT" magazine.

FREE ENGINEERING COUNSEL

Interested audiophiles who desire to improve their FM status and others who have special problems of reception, are invited to write us for advisement. We shall be pleased to give you our honest analysis of your specific problem and suggest proper remedial measures for your consideration.

To avail yourself of our engineering counsel, please enlighten our antenna authority with pertinent information regarding the following provisions;

1. Make and model of FM tuner, sensitivity, 72 or 300 ohm input?
2. What type of antenna is in use; is rotator employed?
3. FM stations you desire to receive.
4. Distance of the desired stations.
5. General topography.
6. Extent of present FM reception.

Address all correspondence to the attention of L. F. B. CARINI, in care of
APPARATUS DEVELOPMENT COMPANY, INC.
 Drawer 153, Wethersfield 9, Conn.

NOTICE

"FM/Q" is the trade mark of the Apparatus Development Company, Inc., manufacturers and purveyors of the foremost type of antennae systems designed exclusively for FM reception. The more popular types available include those illustrated opposite page 6.

Descriptive literature on "FM/Q" antennae, installation kits and the related accessories comprising a complete antenna system may be obtained by writing directly to the company.

APPENDIX

FREQUENCY IN MC vs. FM CHANNELS

88.1	201	.7	234	.5	268
.3	202	.9	235	.7	269
.5	203	95.1	236	.9	270
.7	204	.3	237	102.1	271
.9	205	.5	238	.3	272
89.1	206	.7	239	.5	273
.3	207	.9	240	.7	274
.5	208	96.1	241	.9	275
.7	209	.3	242	103.1	276
.9	210	.5	243	.3	277
90.1	211	.7	244	.5	278
.3	212	.9	245	.7	279
.5	213	97.1	246	.9	280
.7	214	.3	247	104.1	281
.9	215	.5	248	.3	282
91.1	216	.7	249	.5	283
.3	217	.9	250	.7	284
.5	218	98.1	251	.9	285
.7	219	.3	252	105.1	286
.9	220	.5	253	.3	287
92.1	221	.7	254	.5	288
.3	222	.9	255	.7	289
.5	223	99.1	256	.9	290
.7	224	.3	257	106.1	291
.9	225	.5	258	.3	292
93.1	226	.7	259	.5	293
.3	227	.9	260	.7	294
.5	228	100.1	261	.9	295
.7	229	.3	262	107.1	296
.9	230	.5	263	.3	297
94.1	231	.7	264	.5	298
.3	232	.9	265	.7	299
.5	233	101.1	266	.9	300
		.3	267		

DIRECTORY OF FM STATIONS

● Indicates FM Stereo Multiplex.

Ch. 201-220 are educational stations.

City	Call	Freq.	Ch.	Kw.	City	Call	Freq.	Ch.	Kw.
ALABAMA									
Albertville	WAVU-FM	105.1	286	4.5	Berkeley	● KQXR	101.5	268	14.5
Alexander City	WRFS-FM	106.1	291	4.6		KPFB	89.3	207	0.15
Andalusia	WCTA-FM	98.1	251	10.0	KPFA	94.1	231	54.0	
Anniston	WHMA-FM	100.5	263	3.6	KRE-FM	102.9	275	9.9	
Athens	WJOF	104.3	282	1.05	Beverly Hills ●	● KCBH-FM	98.7	254	75.0
Birmingham	● WSMF	93.7	229	20.0	Bijou	KHVR	99.9	260	1.65
	● WCRT-FM	96.5	243	55.4	Claremont	KSPC	88.9	205	1.0
	WAPI-FM	99.5	258	72.0	Coachella	● KCHV-FM	93.7	229	25.0
	WJLN-FM	104.7	284	23.0	KUFM	93.3	227	3.4	
	WBRC-FM	106.9	295	15.0	Eureka	KRED	96.3	242	5.6
Clanton	WKLF-FM	100.9	265	0.37	Fremont	KHYD	104.9	285	0.81
Cullman	WFMH-FM	101.1	266	5.3	Fresno	KRFM	93.7	229	68.0
Decatur	WHOS-FM	92.5	223	8.6	KJPO	94.5	233	7.5	
Huntsville	● WNDA	95.1	236	3.0	KMAK-FM	95.5	238	26.0	
	● WAHR	99.1	256	72.0	KMJ-FM	97.9	250	7.3	
Montgomery	WAJM	103.3	277	29.2	KARM-FM	101.9	270	7.4	
Mobile	WKRG-FM	99.9	260	3.0	KMER	104.7	284	2.9	
Tuscaloosa	WUOA	91.7	219	15.0	Garden Grove ●	● KGGK	94.3	232	0.1
	WTBC-FM	95.7	239	17.5	Glendale	KFMU	97.1	246	58.0
						KUTE	101.9	270	11.0
ALASKA						KHOF	99.5	238	100.0
Anchorage	● KBYR-FM	102.1	271	3.0	Hayward	KB8M	101.7	269	0.9
	KTVA	105.5	288	0.75	Hollywood	KRHM	94.7	234	58.0
ARIZONA						KHJ-FM	101.1	266	16.8
Globe	KWJB-FM	100.3	262	11.0	Inglewood	KTYM-FM	103.9	280	0.39
Mesa	KBUZ-FM	104.7	284	16.0	Lancaster	KANT-FM	107.9	300	20.5
Phoenix	KFCA	88.5	203	0.1	Lodi	KCVR-FM	97.7	249	0.8
	KOY-FM	92.5	223	27.5	Long Beach	KLON	88.1	201	1.2
	KYEW	93.3	227	4.6		KNOB	97.9	250	11.0
	KOOL-FM	94.5	233	58.0		KFOX-FM	102.3	272	1.0
	KELE	95.5	238	18.0		KLFM	105.5	288	0.3
	KPHO-FM	96.9	245	68.0	Los Altos	KPGM	97.7	249	1.0
	KTAR	98.7	254	57.0	Los Angeles	KCMI	88.5	203	0.01
	KITH	101.3	262	4.5		KXLU	89.1	206	0.6
	KEBJ	103.5	278	25.0		KPFK	90.7	214	75.0
Tempe	KUPD-FM	97.9	250	6.4		KUSC	91.5	218	2.9
Tucson	KTAP	96.1	241	25.0		KFAC-FM	92.3	222	60.0
	KFMM	99.5	258	7.2		KNX-FM	93.1	226	67.0
ARKANSAS						● KPOL-FM	93.9	230	17.0
Blytheville	KLCN-FM	96.1	241	21.0		● KRHM	94.7	234	58.0
Forest City	KXJK-FM	93.3	227	0.67		KABC-FM	95.5	238	4.3
Jonesboro	KASU	91.9	220	0.76		KRKD-FM	96.3	242	54.0
	KBTM-FM	101.9	270	8.0		● KFMU	97.1	246	58.0
Little Rock	KMMK	94.1	231	31.0		KHOF	99.5	258	16.5
	KFMA	103.7	279	22.5		● KMLA	100.3	262	58.0
Mammoth Sprs.	KAMS	103.9	280	0.34		KLAC-FM	102.7	274	13.6
Osceola	KOSE-FM	98.1	251	2.2		KGLA	103.5	278	12.5
Pine Bluff	KOTN-FM	92.3	222	3.2		KBIQ	104.3	282	53.0
Siloam Sprs.	KUOA-FM	105.7	289	2.6		KBCA	105.1	286	6.0
CALIFORNIA						● KBMS	105.9	290	18.0
Alameda	KJAZ	92.7	224	1.0		KBBI	107.5	298	19.5
Anaheim	KEZE	95.9	240	0.48	Marysville	KMYC-FM	99.9	260	4.9
Arcata	KHSC-FM	90.5	213	0.01	Modesto	KBEE-FM	103.3	277	11.5
Arlington	KNFP	89.7	209	1.7		KTRB-FM	104.1	281	4.7
Atherton	KPEN	101.3	267	1.5	Mt. View	KFJC	88.5	203	0.01
Auburn	KAFI	101.1	266	4.6	Nat'l City	KEAX	99.3	257	0.84
Bakersfield	KERN-FM	94.1	231	9.1	Newport Beach	KAJS	103.1	276	0.86
	KLYD-FM	96.5	243	3.0	Oceanside	● KJDE-FM	102.1	271	20.0
					Ontario	KASK-FM	93.5	228	1.0
					Oxnard	KAAR	104.7	287	10.5
					Palm Sprs.	KPSR-FM	100.7	264	4.0
					Pasadena	KPCS	89.3	207	0.37
					Redlands	KCAL-FM	96.7	244	1.0

City	Call	Freq.	Ch.	Kw.	City	Call	Freq.	Ch.	Kw.
Redondo Bch.	KAPP	93.5	228	1.0		KWG-FM	105.7	289	3.5
Redwood City	KCUF	107.7	299	20.0		KHIP	106.9	295	39.0
Riverside	KACE-FM	92.7	224	0.7		KSRF	107.3	297	8.3
	• KDUO	97.5	248	72.0	Ventura	KVEN-FM	100.7	264	12.0
	KPLI	99.1	256	1.5	Ventura-Oxnard	• KUDU-FM	95.1	236	24.0
Sacramento	KFBK-FM	92.5	223	1.9	Visalia	• KONG-FM	92.9	225	2.9
	KCRA-FM	96.1	241	45.0	Walnut Creek	• KWME-FM	92.1	221	0.25
	• KSFM	96.9	245	6.4	W. Covina	KDWC	98.3	252	1.0
	KXRQ	98.5	253	3.5	Woodland	• KATT	102.5	273	33.0
	KEBR	100.5	263	15.0					
	KROY-FM	102.5	273	18.0					
	KHIQ	105.1	286						
	KJML	106.5	293	2.7					
	KXOA-FM	107.9	300	12.5					
Salinas	KSBW-FM	102.5	273	18.5					
San Bernardino	KVCR	91.9	220	0.7					
	KFMW	99.9	260	31.0					
San Diego	KSDS	88.3	202	0.37					
	KEBS	89.5	253	78.0					
	KFSF-FM	94.1	231	33.0					
	• KLRO	94.9	235	7.5					
	KFMX	96.5	243	30.0					
	KSEA	97.3	247	29.5					
	KJLM	98.1	251	4.5					
	KfMB-FM	100.7	264	18.5					
	• KGB-FM	101.5	268	36.0					
	KBBW	102.5	273	38.0					
	KSON-FM	103.7	279	11.5					
	KITT	105.3	287	13.0					
	• KPRI	106.5	293	3.5					
San Fernando	KJPL	94.3	232	0.86					
	KSFV	106.3	292	1.0					
San Francisco	KSFX	90.1	211	1.4					
	KALW	91.7	219	3.3					
	KYA-FM	93.3	227	4.9					
	• KSFJ	94.9	235	41.0					
	KOBY-FM	95.7	239	10.5					
	KRON-FM	96.5	243	13.0					
	KEAR	97.3	247	125.0					
	• KAFE	98.1	251	1.7					
	KCBS-FM	98.9	255	35.0					
	KNBC-FM	99.7	259	45.0					
	KDFC	102.1	271	33.0					
	KGO-FM	103.7	279	3.8					
	• KBAY-FM	104.5	283	29.0					
	• KCBO	105.3	287	17.5					
	KFRC	106.1	291	71.0					
	KHIP	106.9	295	39.0					
San Jose	• KSJO-FM	92.3	222	1.0					
	KPRM	98.5	253	3.3					
	KEEN-FM	100.3	262	16.2					
San Luis Obispo	KATY-FM	96.1	241	3.8					
	KSCM	90.9	215	0.27					
	• KVEC	93.3	227	30.2					
Santa Ana	KWIZ-FM	96.7	244	1.0					
	KFIL	106.3	292	1.0					
Santa Barbara	KDB	93.7	229	5.0					
	KRCW	97.5	248	17.5					
	• KMUZ	103.3	277	50.0					
	KGUD-FM	107.9	300	29.5					
Santa Maria	• KEYM-FM	99.1	256	3.2					
	KSMA-FM	102.5	273	11.0					
Santa Monica	KCRW	89.9	210	0.4					
	KSRF	103.1	276	1.0					
Sierra Madre	KMAX	107.1	296	1.0					
Stockton	KCVN	91.3	217	3.4					

COLORADO									
Boulder	KRNW	97.3	247	2.0					
Colo. Sprgs.	KSHS	90.5	219	0.6					
	KRCC	91.3	217	0.16					
	KVOR-FM	92.9	225	1.35					
	KFMH	96.5	243	5.0					
Cortez	KZFM	94.1	231	3.6					
Denver	KDEN	99.5	238	9.1					
	KALH	101.3	267	18.0					
	KLIR-FM	100.3	262	2.4					
	KRKY	102.1	271	8.5					
	KOA-FM	103.5	278	50.0					
	KTGM	105.1	286	10.0					
Golden	• KFML-FM	98.5	253	2.3					
Grand Junction	KREX-FM	92.3	222	2.8					
Manitou Sprgs.	• KCMS-FM	102.7	274	10.0					

CONNECTICUT									
Bridgeport	WJZZ	99.9	260	9.0					
Brookfield	• WGHF	95.1	236	20.0					
Danbury	WLAD-FM	98.3	252	0.34					
Darien	WDRN	95.9	240	0.75					
Hamden	WDEE-FM	101.3	267	20.0					
Hartford	WRTC-FM	89.3	207	0.145					
	WFNQ	93.7	229	7.0					
	• WTIC-FM	96.5	243	8.0					
	WDRG-FM	102.9	275	7.0					
	WHCN	105.9	290	7.0					
	WCCC-FM	106.9	295	6.6					
Manchester	WINF-FM	107.9	300	5.4					
Meriden	• WBMI	95.7	239	7.0					
Middletown	WESU	88.1	201	0.01					
New Haven	WYBC	94.3	232	0.30					
	WNHC-FM	99.1	256	10.0					
Stamford	WSTC-FM	96.7	244	0.65					
Storrs	WHUS	90.5	213	0.01					
Waterbury	WATR-FM	92.5	223	7.0					
	WWCO	104.1	281	20.0					

DELAWARE									
Dover	WDOV-FM	94.7	234	8.3					
Wilmington	WDEL-FM	93.7	229	20.0					
	• WJBR	99.5	258	20.0					

DISTRICT OF COLUMBIA									
Washington	WAMU-FM	88.5	203	12.0					
	WGTB-FM	90.1	211	0.84					
	WRC-FM	93.9	230	20.0					
	WPGC-FM	95.5	238	15.5					
	WTOP-FM	96.3	242	20.0					
	• WASH	97.1	246	15.0					
	WOL-FM	98.7	254	20.0					
	WFAN	100.3	262	20.0					
	WWDC-FM	101.1	266	20.0					
	WGMS-FM	103.5	278	20.0					
	WMAL-FM	107.3	297	20.0					

City	Call	Freq.	Ch.	Kw.	City	Call	Freq.	Ch.	Kw.
FLORIDA					IDAHO				
Coral Gables	● WVCG-FM	105.1	286	18.5	Boise	● KBOI-FM	97.9	250	17.5
Daytona Beach	WNUB-FM	94.5	233	8.5		KBFI	99.9	260	19.0
Ft. Lauderdale	WMFP	100.7	264	56.0	Coldwell	KBGN-FM	94.1	231	7.6
	WWIL-FM	103.5	278	31.0	Lewiston	KOZE-FM	96.7	244	0.9
	● WFLM	105.9	290	2.9					
Ft. Pierce	WIRA-FM	95.5	238	2.48					
	WARN-FM	98.7	254	1.3					
Gainesville	● WRUF-FM	104.1	281	12.0	ILLINOIS				
Jacksonville	WJAX-FM	95.1	236	7.7	Anna	WRAJ-FM	92.7	224	1.0
	WMBR-FM	96.1	241	50.0	Arlington Hts.	WNWC	92.7	224	1.0
	WZFM	96.9	245	9.7	Aurora	WMRO-FM	95.1	236	3.6
Miami	WKAT-FM	93.1	226	17.5		WKKD-FM	95.9	240	1.0
	WMET-FM	93.9	230	13.0	Bloomington	WJBC-FM	101.5	268	5.1
	● WAEZ-FM	94.9	235	56.0	Carbondale	WSIU	91.9	220	22.0
	WGBS-FM	96.3	242	1.5	Carmi	WROY-FM	97.3	247	11.0
	WCKR-FM	97.3	247	50.0	Champaign	WDWS-FM	97.5	248	27.0
	WINZ-FM	99.9	260	105.0	Chicago	WUCB-FM	89.1	206	0.01
	WVPB-FM	101.5	268	8.5		WMBI-FM	90.1	211	47.0
Miami Beach	WAFM	93.1	226	17.5		WBEZ	91.5	218	16.0
Ocala	WMOP-FM	93.7	229	4.3		● WSBC-FM	93.1	226	27.5
Naples	WMFM	94.5	233			WEBH	93.9	230	35.0
Orlando	WDBO-FM	92.3	222	25.0		WENR-FM	94.7	234	4.3
	● WHOO-FM	96.5	243	59.0		WDHF	95.5	238	52.0
	WKIS	100.3	262	5.5		WBBM-FM	96.3	242	10.5
Palm Beach	WQXT-FM	97.9	250	3.6		WNIB	97.1	246	11.0
Pensacola	● WPEX-FM	94.1	231	2.45		WEHS	97.9	250	21.0
St. Petersburg	● WTCX	99.5	258	31.0		● WEMT	98.7	254	29.5
	WVST	105.9	290	105.0		● WEFM	99.5	258	44.0
Sarasota	● WYAK	102.5	273	2.7		WFMF	100.3	262	33.0
Tallahassee	WFSU-FM	91.5	218	0.01		● WMAQ-FM	101.1	266	24.0
Tampa	WTUN	88.9	205	0.77		WCLM	101.9	270	17.0
	WFLA-FM	93.3	227	46.0		● WKFM	103.5	278	50.0
	WDAE-FM	100.7	264	65.0		WJJD-FM	104.3	282	40.0
	WPKM	104.7	284	10.5		● WFMQ	107.5	298	11.0
Winter Park	WPRK	91.5	218	0.33	Decatur	WSOY-FM	102.9	275	31.0
					DeKalb	WNIC	91.1	216	0.01
					E. St. Louis	WAMV-FM	101.1	261	
					Effingham	WSEI	95.7	239	20.0
					Elgin	WEPS	90.9	215	0.38
						WRMN-FM	94.3	232	1.0
						WELG	103.9	280	1.0
					Elmwood Park	WXFM	105.9	290	32.0
					Evanston	WNUR	89.3	207	0.01
						WEAW-FM	105.1	286	36.0
					Glen Ellyn	WELF	107.1	296	3.6
					Harrisburg	WEBQ-FM	99.9	260	4.2
					Jacksonville	WLDS-FM	100.5	263	9.0
					Joliet	WAJP	93.5	228	1.0
					Joliet	WJOL-FM	96.7	244	1.0
					Kewanee	WKSD	91.9	220	0.01
					Litchfield	WSMI	106.1	291	
					Macomb	WWKS	91.3	217	17.0
					Mattoon	WLBH-FM	96.9	245	23.0
					Mt. Carmel	WSAB	94.9	235	37.4
						WVMC-FM	101.1	266	4.0
					Mt. Vernon	WMIX-FM	94.1	231	15.2
					Oak Park	WOPA-FM	102.7	274	1.0
					Olney	WVLN-FM	92.9	225	18.5
					Paris	WPRS-FM	98.3	252	1.0
					Park Forest	WRHS	88.1	201	0.01
					Park Ridge	WMTH	88.5	203	0.01
					Peoria	WMBD-FM	92.5	223	2.5
					Quincy	WTAD-FM	99.5	258	27.0
						● WGEM-FM	105.1	286	8.3
					Rockford	WROK-FM	97.5	248	16.0
					Rock Island	● WHBF-FM	98.9	255	39.0
GEORGIA									
Athens	● WGAU-FM	102.5	273	4.4					
Atlanta	WABE	90.1	211	4.8					
	WGKA-FM	92.9	225	9.4					
	WDJK	94.1	231	30.0					
	WAVQ	94.9	235	7.1					
	● WSB-FM	98.5	253	49.0					
	WPLO-FM	103.3	277	36.0					
Augusta	WBBQ-FM	103.7	279	19.5					
	WAUG-FM	105.7	289	5.5					
Columbus	● WBRL-FM	102.9	275	25.0					
Gainesville	WDUN-FM	103.9	280	0.34					
La Grange	WLAG-FM	104.1	281	2.3					
Macon	WMAZ-FM	99.1	256	3.0					
	WNEX-FM	96.9	245	4.1					
Marietta	WKLS	96.1	241	6.4					
	WBIE	101.5	268	1.35					
Newman	WCOH-FM	96.7	244	0.33					
Savannah	WTOC-FM	97.3	247	7.9					
Toccoa	WLET-FM	106.1	291	0.73					
HAWAII									
Honolulu	KVOK	88.1	201	0.01					
	● KAIM-FM	95.5	238	5.6					
	KEFW	96.3	242	7.7					
	KPOI	97.5	248	27.5					

City	Call	Freq.	Ch.	Kw.
MARYLAND				
Annapolis	WNAV-FM	99.1	256	16.6
	WXTC	107.9	300	20.0
Baltimore	WBJC	88.1	201	0.12
	WYOU	92.3	222	11.0
	WFMM-FM	93.1	226	18.0
	WRBS	95.1	236	15.5
	WBAL-FM	97.9	250	8.7
	• WAQE-FM	101.9	270	
	WCAO-FM	102.7	274	20.0
	WITF-FM	104.3	282	20.0
	WCBM-FM	106.5	293	17.0
Bethesda	• WJMD	94.7	234	20.0
	• WHFS	102.3	272	1.0
Cumberland	WCUM-FM	102.9	275	1.3
	WTBO-FM	106.1	291	7.7
Frederick	WFMD-FM	99.9	260	9.0
Hagerstown	WJEJ-FM	104.7	284	9.4
	WARK-FM	106.9	295	2.2
Havre de Grace	WASA-FM	103.7	279	3.8
Oakland	WPGC-FM	95.5	238	16.5
Silver Sprs.	WGAY	99.5	258	20.0
Takoma Park	WGTS-FM	91.9	220	0.01
Towson	WAQE-FM	101.9	265	20.0
Westminster	WTTR-FM	100.7	264	4.6

City	Call	Freq.	Ch.	Kw.
MASSACHUSETTS				
Amherst	WFCR	88.5	203	34.4
	WAMF	89.5	203	0.01
	WMUA	91.1	216	0.01
Boston	WERS	88.9	205	18.0
	WGBH-FM	89.7	209	15.5
	WBUR	90.9	215	20.0
	WHDH-FM	94.5	233	3.3
	WRKO-FM	98.5	253	20.0
	WCOF-FM	100.7	264	20.0
	WEEI-FM	103.3	277	20.0
	• WBCN	104.1	281	20.0
	WBZ-FM	106.7	294	2.8
Brockton	WBET-FM	97.7	249	0.8
Brookline	WBOS-FM	92.9	225	15.0
Cambridge	WTBS	88.1	201	0.01
Cambridge	WHRB-FM	95.3	237	0.8
	WXHR	96.9	245	20.0
Fitchburg	WFGM-FM	104.7	284	20.0
Framingham	WKOX	105.7	289	15.5
Greenfield	WHAJ-FM	98.3	252	1.0
Haverhill	WHAV-EM	92.5	223	20.0
Lawrence	WGHJ-FM	93.7	229	1.35
Lowell	WLLH-FM	99.5	258	12.2
Lynn	• WUPY	105.3	287	1.05
Medford	WISK	107.9	300	20.0
Mt. Greylock	WAMC	90.3	212	10.0
New Bedford	WBSM-FM	97.3	247	2.5
	WNBH-FM	98.1	251	20.0
Plymouth	WPLM-FM	99.1	256	20.0
So. Hadley	WMHC	90.7	214	0.01
Springfield	WSCB	89.9	210	0.01
	WEDK	91.7	219	15.0
	WHYN-FM	93.1	226	3.2
	WMAS-FM	94.7	234	1.35
Waltham	• WCRB-FM	102.5	273	9.2
W. Yarmouth	WOCB-FM	94.3	232	1.0
Williamstown	WCFM	91.3	217	0.01
Winchester	WHSR-FM	91.9	220	0.01

City	Call	Freq.	Ch.	Kw.
Worcester	• WTAG-FM	96.1	241	10.0
	WAAB-FM	107.3	297	1.4
MICHIGAN				
Ann Arbor	WUOM	91.7	219	115.0
Bay City	WBCN-FM	96.1	241	41.0
	• WNEM-FM	102.5	273	42.0
Benton Harbor	WHFB-FM	99.9	260	9.2
Cadillac	WWTV-FM	92.9	225	100.0
Coldwater	WTVB-FM	98.5	253	6.3
Dearborn	WKMH-FM	100.3	262	7.1
Detroit	WDTR	90.9	215	2.0
	WIPE	92.3	222	10.0
	WJBK-FM	93.1	226	30.0
	WHFI	94.7	234	20.0
	• WLDM	95.5	238	20.0
	WJR-FM	96.3	242	24.0
	WWJ-FM	97.1	246	50.0
	WMZK	97.9	250	20.0
	WFME	98.7	254	8.7
	• WABX	99.5	258	36.0
	WXYZ-FM	101.1	266	27.0
	WDET-FM	101.9	270	79.0
	WMUZ	103.5	278	20.0
	• WOMC	104.3	282	61.0
	WQRS	105.1	286	28.5
	WCHD	105.9	290	34.0
	• WDTM	106.7	294	61.0
	WQTI	107.5	298	32.0
E. Lansing	WKAR-FM	90.5	213	74.0
	• WSWM	99.1	256	30.0
Flint	WFBE	95.1	236	1.3
Grand Rapids	• WJEF-FM	93.7	229	11.5
	WLAV-FM	96.9	245	1.75
	WMAX	101.3	262	7.2
	WFUR	102.9	275	40.7
Highland Park	WHPR	88.1	201	0.01
Holland	WJBL-FM	94.5	233	37.0
Interlochen	WGYA	103.1	276	1.0
Jackson	WBBC	94.1	231	5.4
Kalamazoo	WMCR	102.1	271	36.0
Lansing	WJIM-FM	97.5	248	46.0
	WMRT-FM	100.7	264	92.0
	WBRB-FM	102.7	274	80.0
Mt. Clemens	WOAK	89.3	207	0.01
Royal Oak	WSAM-FM	98.1	251	1.7
Saginaw	WSTR-FM	103.1	276	0.28

City	Call	Freq.	Ch.	Kw.
MINNESOTA				
Brainerd	KLIZ-FM	95.7	239	
Duluth	KUMD-FM	89.1	206	0.01
Mankato	• WAYL	96.1	241	10.1
	• KYSM-FM	103.5	278	47.0
Minneapolis	• KWFM	97.1	246	9.7
	KTIS-FM	98.5	253	5.3
	WLOL-FM	99.5	258	9.7
	KFMV	100.3	262	8.9
	WPBC-FM	101.3	267	5.2
St. Cloud	KFAM-FM	104.7	284	15.0
St. Paul	KNOF	95.3	237	25.0
Worthington	KWOA-FM	94.9	235	26.5

City	Call	Freq.	Ch.	Kw.
MISSISSIPPI				
Jackson	WJDX-FM	102.9	275	50.0
Laurel	WNSL-FM	100.3	277	
Meridian	WMMI	91.5	223	0.65

City	Call	Freq.	Ch.	Kw.	City	Call	Freq.	Ch.	Kw.
MISSOURI					Patterson	● WPAT-FM	93.1	226	20.0
Clayton	KFUO-FM	99.1	265	6.7	Princeton	WPRB	103.5	278	17.0
Crestwood	KSHE	94.9	235	5.1	Red Bank	WFHA-FM	106.3	292	1.0
Joplin	● KSYN	92.5	223	3.6	S. Orange	WSOU	89.5	208	2.0
Kansas City	KCUR-FM	89.3	207	0.36	Trenton	WTOA	97.5	248	14.5
	KTSR	90.1	211	0.01	Wildwood	WCMC-FM	100.7	264	3.5
	KCMK	93.3	227	35.0	Zarepath	WAWZ-FM	99.1	256	4.8
	● KCMO-FM	94.9	235	65.0	NEW MEXICO				
	KXTR-FM	96.5	243		Albuquerque	KANW	89.1	206	0.35
	KISA	99.7	259	1.95		KDEF	94.1	231	1.3
	WDAF-FM	102.1	271	56.0	● KHFM	96.3	242	1.45	
	KBEY	104.3	282	16.5		KARA	99.5	258	3.4
	KBOA-FM	98.9	255	6.9	Aztec	KNDE-FM	94.9	235	3.4
	Poplar Bluff	KWOC-FM	94.5	233	14.0	Los Alamos	KRSN-FM	98.5	253
St. Charles	KADI	96.5	243	25.0	Roswell	KBIM-FM	97.1	246	2.9
St. Louis	KSLH	91.5	218	12.5	NEW YORK				
	WIL-FM	92.3	222	2.0	Albany	WROW-FM	95.5	238	20.0
	● KCFM	93.7	229	60.0	Allegheny	WHDL-FM	95.7	239	43.0
	KPLR-FM	97.3	247	25.5	Amherst	WYSL-FM	103.3	227	4.9
	KSTL-FM	98.1	251	76.0	Auburn	WMBO-FM	96.1	241	18.0
	● KWIX	102.5	273	25.0	Babylon	WBAB-FM	102.3	272	0.7
	KMOX-FM	103.3	277	28.0	● WTFM	103.5	278	15.0	
	KEMO	104.1	281	76.0	Binghampton	WNBF-FM	98.1	251	4.6
	KANG	106.9	295	82.0		WKOP-FM	99.1	256	39.0
Springfield	KTTS-FM	94.7	234	9.1	Buffalo	WBFO	88.7	204	0.19
West Plains	KWPM-FM	93.9	230	3.4		WBUF	92.9	225	94.0
NEBRASKA						WEBR-FM	94.5	233	105.0
Kearney	KHOL-FM	98.9	255	57.0		WGR-FM	96.9	245	16.0
Lincoln	KFMQ	95.3	237	0.32		WBEN-FM	102.5	273	110.0
Omaha	WOW-FM	92.3	222	4.0		WWOL-FM	104.1	281	0.75
	● KQAL-FM	94.3	232	1.0	Central Square	WCSQ	89.3	207	1.0
	KOIL	96.1	241	3.4	Cherry Valley	WRRR	101.9	270	5.4
	KSWI	98.7	254	3.1	Corning	WCLI-FM	106.1	291	4.2
	KFAB-FM	99.9	260	58.0	Cortland	WKRT-FM	99.9	260	14.0
Scotts Bluff	KNEB-FM	94.1	231	5.6	De Ruyter	WRRD	105.1	286	5.3
NEVADA					Elmira	WECW	88.1	201	0.01
Las Vegas	KRAM-FM	100.5	263	2.9	Foral Park	WSHS	90.3	212	0.35
Reno	KNEV	95.5	238	35.0	Foral Park	WOSC-FM	104.7	284	3.1
NEW HAMPSHIRE					Gul'n	● WLIR	92.7	224	75.0
Berlin	WMOU-FM	103.7	279	10.0	Hempstead	WVHC	88.7	204	0.01
Claremont	WTSV-FM	106.1	291	3.4		WHLI-FM	98.3	252	1.0
Manchester	WKBR-FM	95.7	239	3.6	Hornell	WWHG-FM	105.3	287	8.3
Mt. Wash.	WMTW-FM	94.9	235	48.0	Ithaca	WICB	91.7	219	0.01
Nashua	WOTW-FM	106.3	292	1.0		WVBR-FM	93.5	228	0.7
NEW JERSEY						WHCU-FM	97.3	247	40.0
Asbury Park	WJLK-FM	94.3	232	1.0		WRRR	103.7	279	5.6
Atlantic City	WRNJ	95.1	236	3.5	Jamestown	WJTN-FM	93.3	227	9.5
	WOSJ-FM	103.7	279	33.1	Massena	WMSA-FM	105.3	287	13.0
Bridgeton	WSNJ-FM	107.7	299	15.2	Mt. Kisco	WRNW	107.1	295	0.3
Camden	WKDN-FM	106.9	295	7.3	New Rochelle	WVOX-FM	93.5	228	1.0
Dover	● WDHA-FM	105.5	288	1.0	New York	WKCR-FM	89.9	210	4.2
E. Orange	WFMU	91.1	220	0.01		WFUV	90.7	214	3.5
Eatontown	WHTG-FM	105.	288	1.0		WNYE	91.5	218	20.0
Hackettst'n	WNTI	91.9	220	4.8		WHOM-FM	92.3	222	11.0
Long Branch	● WRLB	107.1	296	0.76	● WNYC-FM	93.9	230	18.0	
Newark	WBGO	88.3	202	20.0	WABC-FM	95.5	238	1.5	
	WNTA-FM	94.7	234	13.5	● WQXR-FM	96.3	242	11.0	
	WVNJ-FM	100.3	262	20.0	WNBC-FM	97.1	246	1.1	
New Brunswick	WCTC-FM	98.3	252	1.0	WEVD-FM	97.9	250	20.0	
Newton	WNNJ-FM	103.7	279	1.65	WOR-FM	98.7	254	1.7	
					WBAI	99.5	258	18.0	
					WCBS-FM	101.1	266	1.5	
					WBFM	101.9	270	10.3	

City	Call	Freq.	Ch.	Kw.	City	Call	Freq.	Ch.	Kw.
	WNEW-FM	102.7	274	2.8	Raleigh	WPTF-FM	94.7	234	50.0
	WNCN	104.3	282	15.0		WKIX-FM	96.1	241	29.5
	WRFM	105.1	286	20.0		WRAL-FM	101.5	268	54.0
	WRVR	106.7	294	20.0	Reidsville	WREV-FM	102.1	271	1.6
	WRLB	107.1	296		Rocky Mount	WEED-FM	92.1	221	0.27
Niagara Falls	WHLB-FM	98.5	253	46.0		WFMA	100.7	264	33.0
Olean	WHDL	95.7	239	43.0	Roxboro	WROX-FM	96.7	244	0.6
Patchogue	WALK-FM	97.5	248	15.0	Salisbury	WSTP-FM	106.5	293	15.0
	● WPAC-FM	106.1	291	8.3	Sanford	WWGP-FM	105.5	288	0.49
Peekskill	WLNA-FM	100.7	264	20.0	Shelby	WHOS-FM	96.1	241	2.6
Plattsburgh	WEAV-FM	99.9	260	3.7	Statesville	WDBM-FM	96.9	245	3.6
Poughkeepsie	WKIP-FM	104.7	284	2.3		WFMX	105.7	289	2.4
Rochester	WIRQ	90.9	215	0.01	Tarboro	WCPS-FM	104.3	282	7.0
	● WCMF	96.5	243	1.35	Thomasville	WTNC-FM	98.3	252	0.45
	WRQC-FM	97.9	250	18.0	Wilmington	WPRV	93.9	230	6.8
	WHFM	98.9	255	8.7	Wilson	WVOT-FM	106.1	291	25.0
	WRVM-FM	100.5	263	5.0	Winston-Salem	WFBS-FM	88.1	201	0.01
	WBBF-FM	101.3	267	27.5		WAIR-FM	93.1	226	34.0
Schnectady	● WGFN	99.5	258	6.0		WSJS	104.1	281	12.5
S. Bristol	WRRE	95.1	236	5.3		WYFS	107.5	298	1.55
Springville	WSPE	88.1	201	0.01					
Syracuse	WAER	88.1	201	0.73					
	WDDS-FM	93.1	226	97.0					
	● WSYR-FM	94.5	233	10.0					
	WONO	100.9	265	0.26					
	WHEN-FM	102.9	275	9.0					
Troy	WRPI	91.5	218	0.7					
	WFLY	92.3	222	5.4					
Utica	WRUN	104.3	282	4.3					
Wethersfield	WRRL	107.7	299	5.4					
White Plains	WFAS-FM	103.9	280	0.12					

OHIO

	WAPS	89.1	206	1.3	Akron	WAPS	89.1	206	1.3
	WCUE-FM	96.5	243	14.5		WCUE-FM	96.5	243	14.5
	WAKR-FM	97.5	248	4.4		WAKR-FM	97.5	248	4.4
	WFAH-FM	101.7	269	1.0	Alliance	WFAH-FM	101.7	269	1.0
	WNCO-FM	101.3	267	10.2	Ashland	WNCO-FM	101.3	267	10.2
	WREO-FM	97.1	246	200.0	Ashtabula	WREO-FM	97.1	246	200.0
	WOUB-FM	91.5	218	0.01	Athens	WOUB-FM	91.5	218	0.01
	● WDBN	94.9	235	118.0	Barberton	● WDBN	94.9	235	118.0
	● WCNO	106.9	295	28.0		● WCNO	106.9	295	28.0
	WOMP-FM	100.5	263	9.0	Bellaire	WOMP-FM	100.5	263	9.0
	WBWC	88.3	202	0.01	Berea	WBWC	88.3	202	0.01
	WBGU	88.1	201	0.01	Bowling Green	WBGU	88.1	201	0.01
	WHBC-FM	94.1	231	15.0	Canton	WHBC-FM	94.1	231	15.0
	WMER	94.3	232	0.74	Celina	WMER	94.3	232	0.74
	WGUC	90.9	215	4.7	Cincinnati	WGUC	90.9	215	4.7
	WZIP-FM	92.5	223	82.6		WZIP-FM	92.5	223	82.6
	WOIO	98.5	253	14.0		WOIO	98.5	253	14.0
	WFFM	101.1	266	91.0		WFFM	101.1	266	91.0
	WKRC-FM	101.9	270	14.5		WKRC-FM	101.9	270	14.5
	WSAI-FM	102.7	274	14.7		WSAI-FM	102.7	274	14.7
	● WAEF-FM	104.3	282	4.8		● WAEF-FM	104.3	282	4.8
	WCPO-FM	105.1	286	16.5		WCPO-FM	105.1	286	16.5
	WBOE	90.3	212	9.0	Cleveland	WBOE	90.3	212	9.0
	● WDGO	95.5	238	40.0		● WDGO	95.5	238	40.0
	WERE-FM	98.5	253	40.0		WERE-FM	98.5	253	40.0
	WGAR-FM	99.5	258	31.0		WGAR-FM	99.5	258	31.0
	WHK-FM	100.7	264	11.5		WHK-FM	100.7	264	11.5
	● WDOK-FM	102.1	271	31.0		● WDOK-FM	102.1	271	31.0
	WCRF-FM	103.3	277	21.3		WCRF-FM	103.3	277	21.3
	WJW-FM	104.1	281	2.8		WJW-FM	104.1	281	2.8
	KYW-FM	105.7	289	27.0		KYW-FM	105.7	289	27.0
	WABQ-FM	106.5	293	16.5		WABQ-FM	106.5	293	16.5
	● WNOB	107.9	300	135.0		● WNOB	107.9	300	135.0
	● WCUY	92.3	222	26.0	Cleveland Hts.	● WCUY	92.3	222	26.0
	WOSU-FM	89.7	209	14.0	Columbus	WOSU-FM	89.7	209	14.0
	WCBE	90.5	213	11.0		WCBE	90.5	213	11.0
	WCOL-FM	92.3	222	7.2		WCOL-FM	92.3	222	7.2
	WVKO-FM	94.7	234	52.0		WVKO-FM	94.7	234	52.0
	WTVN-FM	96.3	242	26.0		WTVN-FM	96.3	242	26.0
	● WBNS-FM	97.1	246	54.0		● WBNS-FM	97.1	246	54.0
	WRDF	97.9	250	185.0		WRDF	97.9	250	185.0
	WHIO-FM	99.1	256	19.0	Dayton	WHIO-FM	99.1	256	19.0
	WIFE	104.7	284	24.0		WIFE	104.7	284	24.0
	WAVI	107.7	299	31.0		WAVI	107.7	299	31.0

NORTH CAROLINA

Albermarle	WABZ-FM	100.9	265	0.75
Asheboro	WGWR-FM	92.3	222	10.0
Asheville	WLOS-FM	99.9	260	2.1
Burlington	WBAG-FM	93.9	230	2.8
	● WBBB-FM	101.1	266	3.8
Chapel Hill	WUNC	91.5	218	15.5
Charlotte	WSOC-FM	103.5	278	35.0
	WYFM	104.7	284	1.0
Clingmans Pk.	WMIT	106.9	295	36.0
Concord	WEGO-FM	97.9	250	3.35
Durham	WDNC-FM	105.1	286	36.0
Elkin	WIFM-FM	100.9	265	0.35
Fayetteville	WFNC-FM	98.1	251	14.0
Forest City	WBBO-FM	93.3	227	1.5
	WAGY-FM	105.3	287	3.3
Gastonia	WGNC-FM	101.9	270	11.1
Goldensboro	WEQR	96.9	245	25.5
Greensboro	WGPS	89.9	210	0.01
	● WMDE	98.7	254	5.8
Greenville	WWWS	91.3	217	4.0
Henderson	WHNC-FM	92.5	223	9.0
Hendersonville	WHKP-FM	102.5	273	4.0
Hickory	WHKY-FM	102.9	275	14.8
High Point	WHPS	89.3	207	0.01
	WHPE-FM	95.5	238	13.0
	WMFR-FM	99.5	258	8.0
	WNOS-FM	100.3	262	2.7
Laurinburg	WEWO-FM	96.5	243	9.2
Leaksville	WLOE-FM	94.5	233	13.5
Lexington	WBUY-FM	94.3	232	0.3
Lumberton	WTSB-FM	95.7	239	6.5
N. Wilkesboro	WKBC-FM	97.3	247	4.5

City	Call	Freq.	Ch.	Kw.
Delaware	WLSN	91.1	216	0.01
Eaton	WCTM	92.9	225	7.6
E. Liverpool	WOHI-FM	104.3	282	27.0
Elyria	WEOL-FM	107.3	297	15.0
Findlay	● WFIN-FM	100.5	263	8.2
Fostoria	WFOB-FM	96.7	244	1.0
Fremont	WFRD-FM	99.3	257	1.0
Granville	WDUB	91.3	217	0.01
	WFOL	94.9	235	3.4
Hamilton	WQMS	96.7	244	0.29
	WHOH	103.5	278	8.7
Hillsboro	WSRW-FM	98.1	251	36.3
Kent	WKSU-FM	88.1	201	0.1
Lancaster	WHOK-FM	95.5	238	16.0
Lima	WIMA-FM	102.1	271	15.0
Marion	WMRN-FM	106.9	295	25.0
Marietta	WCMO	89.3	207	0.01
Miamisburg	WFCJ	93.9	230	59.0
Middletown	● WPFB-FM	105.9	290	30.0
Mt. Vernon	WMVO-FM	93.7	229	2.9
Newark	WCLT-FM	100.3	262	8.5
Oxford	WMUB	88.5	203	0.81
	WOXR	97.7	249	0.6
Piqua	WPTW-FM	95.7	239	36.0
Portsmouth	● WPAY-FM	104.1	281	7.0
Salem	WSOM	105.1	286	100.0
Sandusky	WLEC-FM	102.7	274	5.9
Springfield	● WBLV-FM	103.9	280	1.0
Steubenville	WSTV-FM	103.5	278	3.1
Toledo	WTDS	91.3	217	0.73
	WMHE	92.5	223	10.0
	WTRT	99.9	260	8.8
	WSPD-FM	101.5	268	23.0
	● WTOL-FM	104.7	284	1.3
Westerville	WOBN	91.5	218	0.01
Wooster	WWST-FM	104.5	283	13.0
Yellow Sprgs.	WYSO	91.5	218	0.01
Youngstown	WBBW-FM	93.3	227	19.5
	WKBN-FM	98.9	255	50.0
	WRED	101.1	266	24.5

OKLAHOMA

Durant	KSEO-FM	107.3	297	2.75
Norman	WNAD-FM	90.9	215	7.0
Okla. City	KOKH	88.9	205	11.0
	KEFM	94.7	234	3.0
	KYFM	98.9	255	1.5
	KIOO	100.5	263	40.4
	KJEM-FM	102.7	274	29.2
Stillwater	● KOSU-FM	91.7	219	0.38
	KSPI-FM	93.9	230	3.9
Tulsa	KWGS	89.5	203	3.2
	KOGM-FM	92.9	225	4.1
	KIHI	95.5	238	2.95
	● KOCW	97.5	248	1.5
	KAKC	99.5	258	15.5

OREGON

Eugene	KWAX	91.1	216	0.01
	KRVM	91.9	220	0.4
	KEED-FM	93.1	226	1.35
	● KFMY	97.9	250	3.6
	KUGN-FM	99.1	256	0.39
Grants Pass	KGPO	96.9	245	1.3
Medford	KBOY-FM	95.3	237	0.8
Oretech	KTEC	88.1	201	0.01

City	Call	Freq.	Ch.	Kw.
Portland	KRRC	89.3	207	0.01
	KEX-FM	92.3	222	56.0
	● KGMG	95.5	238	68.0
	● KPFM	97.1	246	33.0
	KPOJ-FM	98.7	254	4.4
	KFMC	99.5	258	7.8
	KQFM	100.3	262	17.0
	KOIN-FM	101.1	266	48.0
	KMCP	102.7	274	16.9

PENNSYLVANIA

Allentown	WFMZ	100.7	264	4.8
	WAEB	104.1	281	9.0
Altoona	WFBG-FM	98.1	251	33.0
	WVAM-FM	100.1	261	0.36
Beaver Falls	WBVP-FM	106.7	294	16.5
Bethlehem	WGPA-FM	95.1	236	10.0
Bloomsburg	WHLM-FM	106.5	293	8.8
Boyetown	WBYO-FM	107.5	298	20.0
Braddock	● WLOA-FM	96.9	245	68.0
Butler	WBUT-FM	97.7	249	0.72
Carlisle	WHYL-FM	102.3	272	0.76
Chambersburg	● WCHA-FM	95.1	236	2.8
Dubois	WCED-FM	102.1	271	3.2
Easton	WEEX-FM	99.9	260	3.3
	WEST-FM	105.7	289	11.0
Erie	● WERC-FM	99.9	260	9.7
	WIFI	92.5	223	20.0
Hanover	WHVR-FM	98.5	253	7.2
Harrisburg	WHP-FM	97.3	247	1.55
Havertown	WHHS	89.3	207	0.01
Hazleton	WAZL-FM	97.9	250	2.6
Jenkintown	WIBF	103.9	280	0.75
Johnstown	WARD-FM	92.1	221	0.28
	● WJAC-FM	95.5	238	8.3
Lancaster	WDAC	94.5	233	20.0
	WLAN-FM	96.9	245	6.7
	WGAL-FM	101.3	267	3.6
Lebanon	WLCB	100.1	261	0.72
Levittown	WBCB-FM	100.1	261	1.0
Meadville	WMGW-FM	100.3	262	10.0
Media	WXUR-FM	100.3	262	2.7
Montrose	WPEL-FM	96.5	243	57.0
Oil City	WDJR	98.5	253	3.4
Palmyra	WJWR	92.1	221	0.76
Philadelphia	WXPB	88.9	205	0.01
	WRTI-FM	90.1	211	0.79
	WHYY	90.9	215	20.0
	WPWT	91.7	219	0.18
	WIP-FM	93.3	227	20.0
	WIBG-FM	94.1	231	20.0
	● WFLN	95.7	239	20.0
	● WHAT-FM	96.5	243	20.0
	WCAU-FM	98.1	251	10.0
	WPBS	98.9	255	4.6
	WFIL-FM	102.1	271	6.3
	WPEN-FM	102.9	275	20.0
	WPCA-FM	104.5	283	20.0
	WDAS-FM	105.3	287	20.0
	● WQAL	106.1	291	7.8
Pittsburgh	WDUQ	91.5	218	2.75
	KDKA-FM	92.9	225	47.0
	● WKJF	93.7	229	40.0
	● WWSW-FM	94.5	233	20.0
	WCAE-FM	96.1	241	31.0
	WJAS-FM	99.7	259	24.0

City	Call	Freq.	Ch.	Kw.
	WFMP	99.7	259	24.0
	WPGH	104.7	284	2.82
	WAZZ	105.9	290	4.5
	WINE	107.9	300	36.0
Pottsville	WPPA-FM	101.9	270	2.8
Red Lion	WGCB-FM	96.1	241	20.0
Scranton	WUSV	89.9	210	1.35
	WGBI-FM	101.3	267	1.8
Sharon	WPIC-FM	102.9	275	26.0
State College	WDFM	91.1	216	0.25
Sunbury	WKOK-FM	94.1	231	4.4
Towanda	WTTC-FM	92.7	224	7.2
Warren	• WRRN	92.3	222	3.0
Waynesboro	WAYZ-FM	95.1	236	3.0
Wilkes-Barre	WBRE-FM	98.5	253	2.25
	• WYZZ	103.3	277	3.1
Williamsport	WRAK-FM	102.7	274	54.0
	WLYC-FM	105.1	286	0.84
York	WNOW-FM	105.7	289	1.2

RHODE ISLAND

Cranston	WLOV	99.9	260	3.4
Providence	WPRO-FM	92.3	222	15.0
	• WPFM	95.5	238	3.2
	WHIM-FM	98.9	255	3.36
	• WXCN	101.5	268	20.0
	WPJB-FM	105.1	286	20.0
Warwick	WDJD	107.7	299	3.3
Westerly	WERI-FM	103.7	279	17.5
Woonsocket	WWON-FM	106.3	292	0.39

SOUTH CAROLINA

Anderson	WCAC	101.1	266	6.4
Charleston	WTMA-FM	95.1	236	10.0
	WCSC-FM	96.9	245	68.0
Clemson	WS9F-FM	88.1	201	0.01
Columbia	WUSC-FM	89.9	210	0.01
	WCOS-FM	97.9	250	5.3
	WNOK-FM	104.7	284	3.4
Dillon	WDSC-FM	92.9	225	8.8
Greenville	WESC-FM	92.5	223	12.0
	WFBC-FM	93.7	229	9.9
	WMMU-FM	94.5	233	0.84
Laurens	W19G-FM	100.5	263	3.5
Seneca	• WSNW-FM	98.1	251	9.4
Spartanburg	• WSPA-FM	98.9	255	4.9

TENNESSEE

Bristol	WOPI-FM	96.9	245	9.7
Chattanooga	WDOD-FM	96.5	243	13.0
	WLOM	106.5	293	33.0
Collegedale	WSMC-FM	88.1	201	1.0
Gallatin	WFMG	104.5	283	8.2
Greeneville	WGRV-FM	94.9	235	4.6
Jackson	WTJS-FM	104.1	281	50.0
Johnson City	WJCW-FM	100.7	264	8.1
Kingsport	WKPT-FM	98.5	253	10.0
Knoxville	WKCS	91.1	216	0.31
	WUOT	91.9	220	70.5
	WBIR-FM	93.3	227	3.3
Memphis	WQMM	95.5	238	1.3
	WMPS-FM	97.1	246	6.6
	WMC-FM	99.7	259	30.0
	WVIA	102.7	274	5.2

City	Call	Freq.	Ch.	Kw.
Nashville	WSIX-FM	97.9	250	30.0
	WFMB	105.9	290	3.4
Sevierville	WSEV-FM	102.1	271	22.0

TEXAS

Amarillo	KGNC-FM	93.1	226	14.5
Austin	KUT-FM	90.7	214	4.1
	• KTBC-FM	93.7	229	94.0
	KAZZ	95.5	238	1.35
	KHFI	98.3	252	0.78
Beaumont	KRIF-FM	97.5	248	9.1
Brownwood	KHPC	88.1	201	0.01
Cedar Hill	KDFW	99.5	258	25.0
Cleburne	KCLE-FM	94.9	235	6.35
Corpus Christi	KMFM	95.5	238	10.0
Dallas	KNER	88.1	201	0.01
	KVTT	91.7	219	0.78
	• KRLD-FM	92.5	223	40.0
	KCPA-FM	94.1	231	4.8
	WFAA-FM	97.9	250	38.0
	KOST	98.7	254	16.5
	WRR-FM	101.1	266	68.0
	KQRO	102.9	275	7.6
	• KIXL-FM	104.5	283	20.0
	• KPSD	105.3	287	8.75
Denton	KDNT-FM	106.3	292	0.76
Diboll	KSPF-FM	95.5	238	6.5
Dumas	KDDD	95.3	237	
El Paso	KVOF-FM	88.5	203	0.01
	KHMS	94.7	234	2.55
Ft. Worth	• WBAP-FM	96.3	242	25.0
	KFJZ-FM	97.1	246	22.0
	KFMF	102.1	271	1.68
Gainesville	KGAF-FM	94.5	233	2.5
Harlingen	KELT	94.5	233	3.1
Hillsboro	KHBR-FM	102.5	273	3.3
Houston	KUHF	91.3	217	9.8
	KJSB	93.7	229	18.5
	KARO	94.5	233	7.6
	KHUL	95.7	239	15.5
	• KXYZ	96.5	243	14.6
	• KFMK	97.9	250	10.5
	• KHGM	99.1	256	49.0
	KROW	100.3	262	16.5
	KTRH-FM	101.1	266	29.5
	• KQUE	102.9	275	64.2
	• KRBE	104.1	281	40.0
Lubbock	KTXT-FM	91.9	220	0.01
	KRKH-FM	93.7	229	9.6
	KBFM	96.3	242	3.6
Marshall	KMHT-FM	97.3	247	5.9
Midland	• KNFM	92.3	222	4.8
Odessa	KQIP	96.9	245	1.7
	KWMO	99.1	256	3.4
Pampa	KBMF-FM	100.3	262	3.0
Plainview	KHBL	88.1	201	0.01
	KVOP-FM	97.3	247	3.0
Porth Arthur	• KFMP	93.3	227	3.2
San Antonio	KITY	92.9	225	3.8
	KEEZ	97.3	247	17.0
	KAKI	98.1	251	25.0
	KISS	99.5	258	12.9
Texarkana	KCMC-FM	98.1	251	1.4
Waco	KEFC	95.5	238	3.2
	WACO-FM	99.9	260	3.9
Waxahachie	KBEC-FM	93.5	228	0.35

City	Call	Freq.	Ch.	Kw.	City	Call	Freq.	Ch.	Kw.
UTAH									
Logan	KVSC	88.1	201	0.01		KTNT-FM	97.3	247	10.2
Provo	KBYU-FM	88.9	205	0.01		KTWR	103.9	280	0.83
Salt Lake City	KUER	90.1	211	0.35		KLAY-FM	106.3	292	0.83
	KLUB-FM	97.1	246	4.8	WEST VIRGINIA				
	KCPX-FM	98.7	254	1.2	Beckley	WBKW	99.5	258	34.0
	KSL-FM	100.3	262	5.9	Charleston	WKAZ-FM	97.5	248	9.0
VIRGINIA						• WKNA	98.5	253	2.65
Arlington	WAVA-FM	105.1	286	2.05	Huntington	WKEE	100.5	263	53.0
Charlottesville	WTJU	91.3	217	0.01	Martinsburg	WEPM-FM	94.3	232	0.81
	WINA-FM	95.3	237	0.62	Morgantown	WAJR-FM	99.3	257	1.0
	WCCV	97.5	248	3.4	Oak Hill	WOAY-FM	94.1	231	14.5
Crewe	WSVS-FM	104.7	284	14.0	Parkersburg	WAAM-FM	106.5	293	8.9
Fredericksburg	WFVA-FM	101.5	268	4.6	Wheeling	WKWK-FM	97.3	247	10.0
Gretna	WMNA-FM	103.3	277	3.0		WWVA-FM	98.7	254	7.4
Hampton	WVEC-FM	101.3	267	28.0	WISCONSIN				
Harrisonburg	WEMC	91.7	219	0.01	Appleton	WLFM	91.1	216	10.5
	WSVA-FM	100.7	264	7.2	Baraboo	WWCF	94.9	235	37.0
Lynchburg	WWOD-FM	100.1	261	0.94	Brule	WWSA	89.9	210	38.0
Marion	WMEV-FM	93.9	230	3.7	Chilton	WHKW	89.3	207	51.0
Martinville	• WMVA-FM	96.3	242	2.9	Colfax	WHWC	88.3	202	5.0
Newport News	WGH-FM	97.3	247	36.0	Delafield	WHAD	90.7	214	52.0
Norfolk	WFOS	90.5	213	0.01	Eau Claire	• WIAL	94.1	231	60.0
	WMTI	91.5	218	1.45	Ft. Atkinson	WFAW	107.3	297	3.0
	• WYFI	99.7	259	11.5	Green Bay	WDUZ-FM	98.5	253	2.85
	WRVC	102.5	273	8.7	• WBAY-FM	101.1	266	13.5	
Portsmouth	WAVY-FM	98.9	255	25.8	Highland	WHHI	91.3	217	43.0
Richmond	WRFK	91.1	216	3.6	Janesville	WCLO-FM	99.9	260	15.5
	WRVA-FM	94.5	233	25.0	Madison	WHA-FM	88.7	204	22.0
	• WCOD	98.1	251	34.0	• WISC-FM	98.1	251	1.0	
	WRNL-FM	102.1	271	50.0	WIBA-FM	101.5	268	45.0	
Roanoke	WLRJ	92.3	222	2.8	WRVB-FM	102.5	273	3.9	
	WDBJ-FM	94.9	235	14.5	• WMFM	104.1	281	7.5	
	• WSLS-FM	99.1	256	21.0	WQFM	93.3	227	30.0	
So. Boston	WHLF-FM	97.5	248	2.4	• WTMJ-FM	94.5	233	4.6	
Staunton	WSGM-FM	93.5	228	0.5	WMIL-FM	95.7	239	29.0	
Williamsburg	WCWM	89.1	206	0.01	• WFMR	96.5	243	15.0	
	WBCI-FM	96.5	243	3.6	WISN-FM	97.3	247	25.5	
Winchester	WRFL	92.5	223	22.0	WMLW	98.1	251	141.0	
Woodbridge	WWCN	105.9	290	9.5	WEMP-FM	99.1	256	25.0	
					• WMKE	102.1	271	115.0	
WASHINGTON						WRIT-FM	102.9	275	22.0
Bellingham	KGMI	92.9	225	20.0	WBON	107.7	299	38.0	
Edmonds	KGFM	105.3	287	120.0	Monroe	WEKZ-FM	93.7	229	15.0
Lynden	KLYN-FM	106.5	293	1.45	Racine	WRJN-FM	100.7	264	15.0
Opportunity	KZUN-FM	96.1	241	3.3	Rice Lake	WJMC-FM	96.3	242	50.0
Seattle	KOTO	93.3	227	6.64	Ripon	WRPN-FM	90.1	211	0.01
	KUOW	94.9	235	45.0	S. Beloit	WBEL	103.1	276	
	• KGMJ	95.7	239		Stevens Pt.	WSPT-FM	97.9	250	15.5
	• KLSN	96.5	243	14.5	Tomah	WTMB-FM	98.9	255	175.0
	KING-FM	98.1	251	15.0	Watertown	WTTN-FM	104.7	284	9.6
	KMCS	98.9	255	10.5	Sparta	WCOW-FM	97.1	246	
	• KISW	99.9	260	10.5	Waukesha	WAUX-FM	106.1	291	3.3
	KIRO-FM	100.7	264	16.5	Wausau	WHRM	91.9	220	75.0
	• KETO-FM	101.5	268	17.7		WLIN	95.5	238	20.5
	KPRN	102.5	273	19.0	Wauwatosa	WTOS	103.7	279	
Spokane	• KZAM	92.5	223	175.0	West Bend	• WBKV-FM	92.5	223	18.0
	KREM-FM	92.9	225	4.8	W. Salem	WHLA	90.3	212	39.0
	KXLY-FM	99.9	260	2.0	Wisc. Rapids	WFHR-FM	103.3	277	2.1
	KHQ-FM	98.1	251	47.0	WYOMING				
Tacoma	KGPS	90.9	215	0.8	Cheyenne	KVWO-FM	106.3	292	
	KTOY	91.7	219	3.5					

