Since SEOR requires the system matrix to possess Young's property A, one has to use a seventeen-point operator [8] in conjunction with the five-point Laplacian instead of the thirteen-point operator as used in [7].

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Highly Stabilized IMPATT Oscillators at Millimeter Wavelengths

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***Abstract*—Highly stabilized IMPATT oscillators at millimeter wavelengths have been developed. The IMPATT diode is mounted in the coaxial-waveguide circuit at the detuned open position, and is series-resonant at the design frequency. The frequency stability and power output of $\pm 5\times 10^{-5}/\mp 20$ °C and 50 mW, respectively, have been obtained at 80 GHz.**

Highly stabilized solid-state oscillators at millimeter wavelengths are presently required, especially for the realization of a phase-shift keyed (PSK) guided millimeter-wave transmission system. Several types of oscillators have been developed, mainly in the microwave region [1]–[3], but also in the millimeter-wave region [4]. This short paper describes the design considerations and the experimental results of the reaction-cavity controlled IMPATT oscillator having good frequency stability in the millimeter-wave region.

The construction of the oscillator is essentially the same as the highly stabilized *Ka*-band Gunn oscillator [5], but the operating point is made more suitable for the stabilization. In the millimeter-wave region it is difficult to obtain superior frequency stability, because the unloaded Q of the cavity is much smaller than that of the cavity in the microwave region (for instance, Q_0 at 100 GHz is about one-third of Q_0 at 10 GHz, even if the cavity is ideally constructed). This problem can be solved by utilizing a higher order resonant mode of the cavity and also making the oscillator operate at the point where the increment of the susceptance against frequency is maximum. In order to realize the latter, we have placed the diode at the detuned open position of the cavity and also made the diode series-resonant at the design frequency.

The equivalent circuit of the oscillator is shown in Fig. 1. The assumptions are made that 1) the electronic susceptance of the device is negligibly smaller than that due to the junction capacity C_j, 2) the diode has only a series lumped inductance L_e, except for the negative conductance of the device G_d and the capacitance C_j, 3) the diode is connected to the transmission line of characteristic impedance Z_2 at the detuned open position of the reaction cavity ($1-1'$ in Fig. 1), and 4) the diode is also connected to the line of characteristic impedance Z_0 through an ideal transformer of turns ratio n:1. The locus of the admittance Y_c, which the negative conductance sees, is shown in Fig. 2. The assumptions are made that the series-

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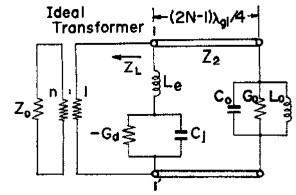

Fig. 1. Equivalent circuit of stabilized IMPATT oscillator.

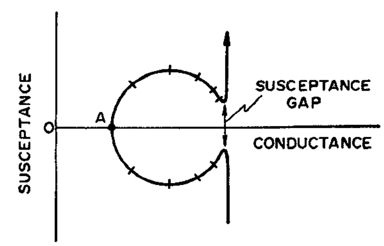

Fig. 2. Admittance locus of reaction-cavity controlled oscillator.

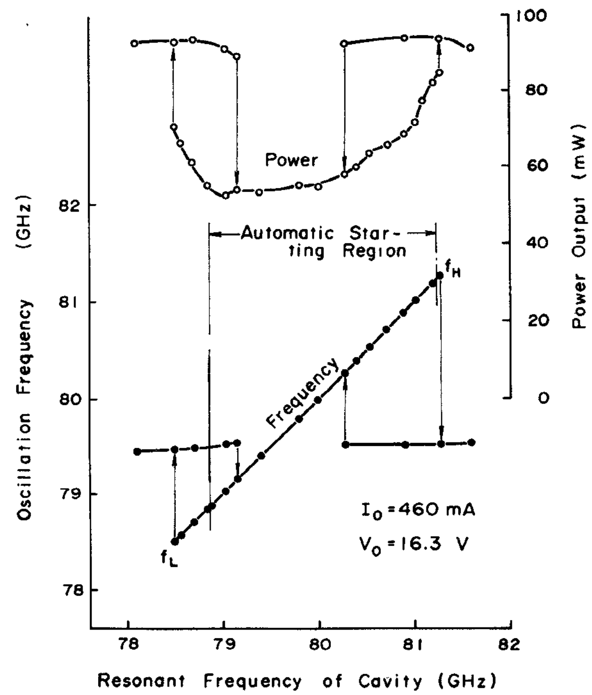

Fig. 3. Tuning characteristic of an 80-GHz IMPATT oscillator.

resonant frequency of the diode $\omega_s(=1/\sqrt{L_eC_j})$ is equal to the resonant frequency of the cavity $\omega_0(=1/\sqrt{L_0C_0})$, and $Q_e(=\omega_0C_0Z_2) \gg Q_1(=\omega_sL_e/Z_2)$. The arrow represents the direction of the increment of the frequency, and the cross lines equal increments of frequency. The oscillator can be stably operated at point A, where the increment of the susceptance against frequency is maximum. As the admittance locus has a "susceptance gap" (refer to Fig. 2), the single-mode operation can be obtained in a narrow tuning range.

An 80-GHz IMPATT oscillator has been designed, based on the above design considerations. The reaction cavity is made of "super-Invar," whose resonant mode is the cylindrical TE_{013} mode. The measured value of Q_0 was 11 000 (cf. theoretical value is about 16 000). In order to make the diode series-resonant at 80 GHz, a coaxial-waveguide circuit [6] has been adopted, where the dimensions of the waveguide cross section are 3.1×1.0 mm. The distance between the diode and the detuned short position of the cavity is $3\lambda_g/4$ at 80 GHz.

Fig. 3 shows the tuning characteristic of the oscillator. In the frequency range from 79.2 to 80.2 GHz, the oscillator operates only

TABLE I

ELECTRICAL PERFORMANCE OF 80-GHz STABILIZED IMPATT OSCILLATOR

Frequency	80.000	(GHz)
Power output	55	(mW)
Pushing figure	50	(kHz/mA)
Frequency stability*	-2.5×10^{-6}	($°C^{-1}$)
Power stability*	∓ 1	(dB/$\pm$20°C)
External Q	10000	
Dc current	460	(mA)

*) Temperature range : 5 ~ 45°C

in a cavity-controlled mode where the oscillation frequency is almost the same as the resonant frequency of the reaction cavity. The "pushing figure" was minimum near 80 GHz. The electrical performance at 80 GHz is listed in Table I. The frequency stability of $\pm 5 \times 10^{-6}/\mp 20$ °C and 50 kHz/mA and power output of 55 mW have been obtained.

In a V-band IMPATT oscillator, nearly the same frequency stability has been obtained.

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Channel Multiplexing Network for a 20-GHz Radio-Relay Transmission System

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***Abstract*—The Nippon Telegraph and Telephone Public Corporation (NTT) plans to make a field test for the practical application of the 20-GHz radio-relay transmission system. This short paper describes channel multiplexing–demultiplexing networks fabricated for use in this test. Overall loss of the network constructed by three stages, namely, a Vertical–horizontal (V–H) polarizing filter, a transmit–receive filter, and a channel-dropping filter, is 5 dB, even though a 2-dB loss of a flexible waveguide is included.**

A radio-relay pulse-code-modulation transmission system in the 20-GHz band has been studied as a large-capacity communication system to prepare for an increase in transmission medium demands caused by new communication services, for instance, picture phone and data transmission [1], [2].

The Nippon Telegraph and Telephone Public Corporation (NTT) is going to make a field test for the practical application of a 20-GHz radio-relay communication system, called the 20G–400M system [2]. The trial link with two terminal stations will be installed over an 2.8-km route.

Fig. 1 shows the construction of the multiplexer–demultiplexer for the field test, of which specifications are shown as follows:

frequency band	17.7–21.0 GHz (overall band—3.2 GHz) and common use of both polarizations;
clock frequency	200 MBd (a bit rate of 400 Mb/s with 4-phase phase-shift keyed);
channel spacing	300 MHz;
guard band	500 MHz.

The multiplexer consists of one vertical–horizontal (V–H) polarizing filter, two transmit–receive filters, 20 channel-dropping filters, and BRF and BPF filters suppressing frequency crosstalk to received signals from transmitted signals [3], [4]. The common use of both polarized waves, that is, a V wave and an H wave, is adopted in order to use restricted frequency bands efficiently. However, heavy rainfall degrades the polarization isolation. This may make it necessary to stagger horizontally and vertically polarized channels. Operation of the trial test link will provide additional data on degradation of cross polarization by rainfall.

First, two orthogonal polarized waves are separated by a V–H polarizing filter. Next, the 17.7–21.0-GHz band is divided into two by the transmit–receive filter, resulting in four groups: A_V, A_H, B_V, and B_H, each having a bandwidth of 1.6 GHz. It is planned to achieve two-way transmission by a single antenna with A_V and A_H groups of lower frequencies in one direction and B_V and B_H groups of higher frequencies in the opposite direction.

A concentrated coupled type like the V–H polarizing filter, a circulator like the transmit–receive filter, and a ring-type filter like the channel-dropping filter are adopted, as described later.

The multiplexer occupies only $1055 \times 860 \times 200$ mm^3 (box size) except for the V–H polarizing filter, which is set on back of the antenna mounted at the top of a pole [2]. The multiplexer is maintained at a low dry air pressure of 0.05 kg/cm^2 above atmospheric pressure.

The trial V–H polarizing filter is a concentrated coupled type, as shown in Fig. 2. One polarized wave is reflected by a plate and emerges to port 2 and the other is transduced from circular TE_{11} to rectangular TE_{10} by a taper-type mode transducer and emerges to port 3. The insertion loss is 0.16 dB for each polarized wave from port 1 to port 2 or 3.

The size of the inner circular waveguide, operated in a dominant TE_{11} mode, is 10.8 mm and the rectangular waveguide is WRJ-180 ($a = 12.954$ mm and $b = 6.477$ mm).

The polarization isolation is over 45 dB and each input VSWR is under 1.10 for the entire frequency range.

Since the frequency crosstalk is most strongly expected at the boundary channels, filters such as BPF and BRF, as well as the transmit–receive filter, must be provided to suppress this crosstalk. The two dominant interference paths, namely, to R_6 from T_5 in one station and to R_5 from T_6 in the next station, must be considered. Also, the crosstalk level must be suppressed to under 114 dB in order to realize required $D/U = -33$ dB when transmitted power, fading margin, section loss, and antenna gain are taken into consideration [2], [3].

There are two types of transmit–receive filters: one consists of two hybrids and two high-pass cutoff filters and the other uses a circulator. In the case of the cutoff-filter-type transmit–receive filter, the former crosstalk, namely, to R_6 from T_5, is sufficiently suppressed by the very sharp frequency response and high attenuation of the cutoff area, but the latter one is not sufficiently suppressed. On the other hand, for the circulator type, the two crosstalks must be suppressed by inserting some filters. In this test, the circulator type is adopted because it is smaller in size and lower in loss than the cutoff-filter type. Moreover, it is confirmed [3] that the crosstalks can be sufficiently suppressed by installing a receiving BPF with a five-cavity Butterworth response and a BRF with a two-cavity Butterworth response, as shown in Fig. 1. The former, with 400 MHz of 3-dB bandwidth, is composed of 6 inductive metal posts. The construction of the latter is the same as for the ring-type channel-dropping filter using the response of ports 1 and 2 as BRF.

The forward loss of the trial circulator (Fig. 3) is under 0.15 dB and the backward loss is over 35 dB for the entire band.

Fig. 4 shows the trial ring-type filter, which has been developed as a channel-dropping filter for the proposed millimeter-wave multiplexing network in the NTT [5], [6].

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